WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

OCTOBER 1953

VOL. 30 No. 10 THREE SHILLINGS AND SIXPENCE

20,000 ohms per volt *plus*

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Produced in response to a demand for a high sensitivity version of the world-famous Universal AvoMeter, this model incorporates the traditional design features of its predecessors, so highly valued for simplicity of operation and compact portability.

It has a sensitivity of 20,000 ohms per volt on all D.C. voltage ranges and 1,000 ohms per volt on A.C. ranges from 100V. upwards. A decibel scale is provided for audio frequency tests. In addition, a press button has been incorporated which reverses the direction of current through the moving coil, and thus obviates the inconvenience of changing over test leads when the current direction reverses. It also simplifies the testing of potentials, both positive and negative, about a common reference point. A wide range of resistance measurements can be made using internal batteries, separate zero adjustment being provided for each range.

It is of importance to note that this model incorporates the "AVO" automatic cut-out for protection against inadvertent overloads

£23 : 10s.

Size 8½″ × 7¼″ × 4½″
Weight 6½lbs. (including leads)

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For your Valve Characteristic Meter or Valve Tester

Owing to the very large number of valves which have been issued within the last two years, no further amendments will be issued for the original "Avo" Valve Testing Manual. A new, completely revised and fully up-to-date Valve Data Manual is now available from the Company at 15/- post free.

D.C. VOLTAGE	D.C. CURRENT	A.C. VOLTAGE	A.C. CURRENT	RESISTANCE
2.5V.	50µA.	2.5V.	100mA.	*First indication 0.5Ω.*
10V.	250µA.	10V.	1A.	*Maximum indication 20MΩ.*
25V.	1mA.	25V.	2.5A.	0—2,000Ω using internal batteries
100V.	10mA.	100V.	10A.	0—200,000Ω using internal batteries
250V.	100mA.	250V.	—	0—20MΩ using internal batteries
1,000V.	1A.	1,000V.	—	
2,500V.	10A.	2,500V.	—	0—200MΩ using external batteries

THE AUTOMATIC COIL WINDER & ELECTRICAL EQUIPMENT CO. LTD.
WINDER HOUSE · DOUGLAS STREET · LONDON S.W.1 *Telephone* VICtoria 3404—9

A8 1

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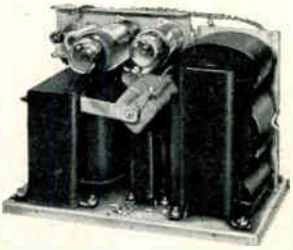
(A.C. OUTPUT)

- ● EXTREMELY PRECISE CONTROL (CONSTANCY 0·15%)
- ● INSENSITIVE TO CHANGES OF LINE FREQUENCY
- ● EXCELLENT REGULATION FROM ZERO TO 100% LOAD
- ● VERY COMPACT ● LIGHTWEIGHT ● FOOLPROOF

"BAVR" Automatic Voltage Stabilizers employ a basically "new" arrangement, having very many real advantages which will be welcome. They are quite unaffected by frequency changes of a commercial nature, have excellent regulation from zero to maximum load, and provide the hitherto impossibly close constancy of 0·15% which puts this in a performance class by itself. Complete data will gladly be sent on written application.

Rear View
BAVR-200

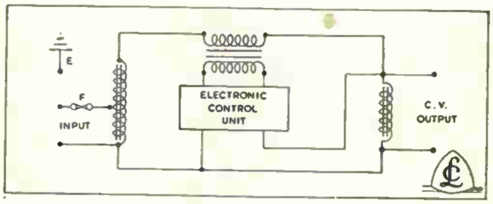
The control unit is a magnetic amplifier, the inductance of which varies with the DC passed through control coils. Stabilization is achieved by monitoring the output side and regulating automatically the D.C. component so as to adjust the AC output voltage and keep it constant within precise limits, the stability being 0·15%. The functional circuit is as shewn. The Electronic Control Unit employs three tubes—one each EL37, 90C1 and A2087, and a selenium rectifier: it is foolproof.



Inside View
BAVR-200

Usual Input Voltage "Swing" is 230V—12% to + 6%, and usual output voltage 230V, tolerance 0·15%, and means are provided so that the instruments may be changed by the customer, to suit them to other input voltages from 200-250V, nominal.

We are also now in a position to accept orders for three larger models, type BMVR: these work on a different and still newer principle, which includes all the good features of "BAVR" plus the ability to cater for large or very large loads, plus undistorted output (sinusoidal, and free from troublesome harmonics).



Functional Circuit
BAVR

TYPE	MAXIMUM LOAD	NET PRICE
BAVR-200	200 VA	£50 0 0
BAVR-500	500 VA	£57 10 0
BAVR-1,000	1 kVA	£70 0 0
BMVR-1,725*	Ca. 2 kVA	£75 0 0
BMVR-7,000*	Ca. 7 kVA	Prices on application
BMVR-25,000*	Ca. 25 kVA	

* Full details now in active preparation, and will gladly be sent on written request. These three Instruments have sinusoidal output waveform (no percentage of harmonics).

Model
BMVR-1725

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Accuracy	±0.005%
Output	0 to 30 volts r.m.s. metered
Attenuator	0 to 110db in 1db steps (constant 600Ω)

The equipment will measure any frequency in the range 10c/s to 100kc/s and any time in the range 10μsec to 10^4sec both to an accuracy within ±0.005%. It will also count up to a maximum rate of 10^5 pulses/second.

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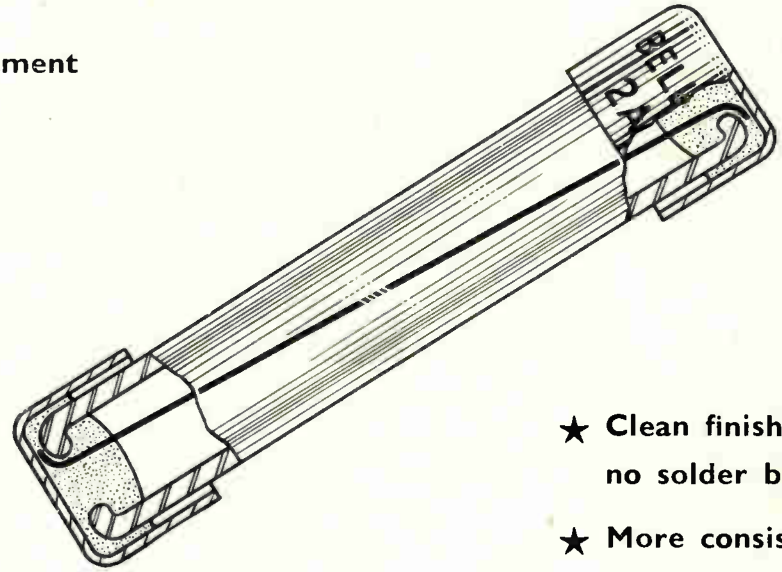
- ★ **Non-sag filament**



- ★ **Clean finish caps— no solder blobs**
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They are made in ratings from 60mA to 25 amps, but we do not guarantee that ratings above 7 amps will clear the high prospective currents specified for the lower ratings.

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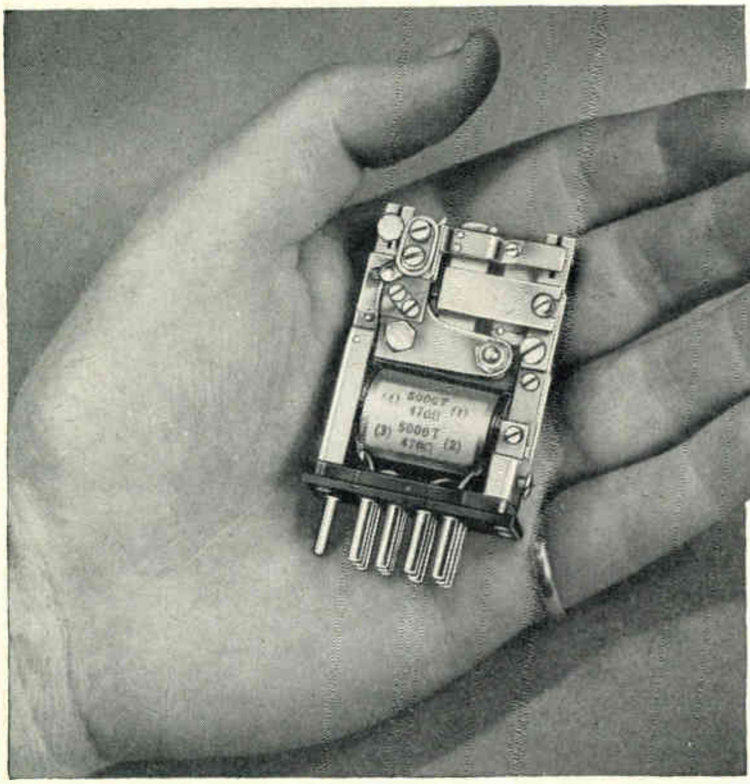

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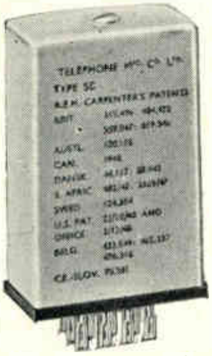

Complete specification and further details of the complete range of Carpenter Relays may be had on request.

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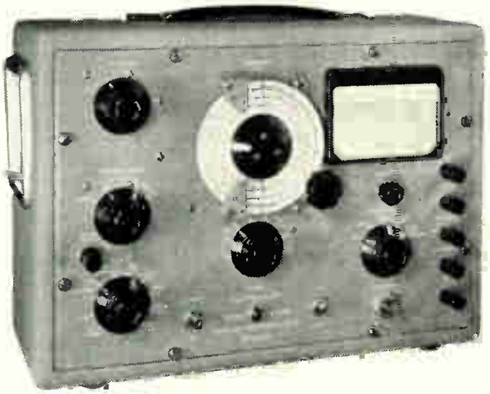

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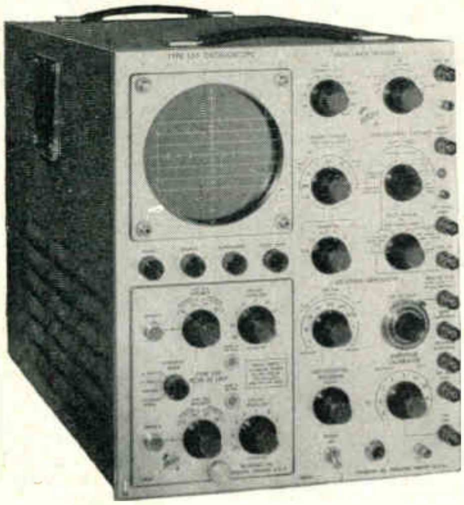
Now a single cathode-ray oscilloscope with its associated plug-in preamplifiers does the work of three or more separate high-quality laboratory oscilloscopes.



... the Tektronix Type 535

Quickly Adaptable to Changing Requirements with Plug-in Vertical Preamplifiers

1. Dual-Trace Unit—Performance superior to dual-beam scope in most applications.
2. Differential High-Gain DC Unit—Sensitivity to 1 millivolt/cm.
3. Wide-Band DC Unit—10 megacycle bandwidth.

New Sweep Circuit Provides Widest Range Ever Offered

1. Accurately calibrated time bases from 0.02 μsec/cm to 5 sec/cm.
2. New sweep delay permits jitter-free display of details in pulse trains.
3. Calibrated interval between sweeps

High Intensity for Low Repetition Rates and Single Sweeps

1. 10-KV accelerating potential
2. New Tektronix 5" flat-faced metallized cathode-ray tube.

The fast growth of the electronic industry is bringing more complex problems to the engineer every day. The need for an oscilloscope quickly adaptable to almost any phase of laboratory investigation is evident. The Type 535 Cathode-Ray Oscilloscope is designed to fill this need.

Plug-in vertical preamplifiers enable the Type 535 to cope with a wide variety of signals. Wide range time base has single shot, recurrent, and triggered operation to permit observation of the very slow as well as the very fast. Flexible sweep delay circuitry provides both jitter-free delay and stable viewing of time-modulated or jittery signals. High crt accelerating potential increases the brightness at low repetition rates. With dual-trace plug-in unit precise time comparisons may be made, free from errors inherent in dual-beam cathode-ray tubes.

CHARACTERISTICS

Basic Unit

TIME-BASE RANGE—MAIN SWEEP

24 calibrated time bases from 0.1 μsec/cm to 5 sec/cm. Continuously variable uncalibrated time base from 0.1 μsec/cm to 10 sec/cm.

SWEEP MAGNIFIER

Accurate 5x magnification on all ranges including the 0.1 μsec/cm range.

DELAYING SAWTOOTH

Automatic main sweep lockout with contrallable reset. Permits delaying the start of the main sweep from 10 μsec to 20,000 μsec in 9 calibrated ranges. Calibrated interval between sweeps, 20 to 20,000 μsec (50 cps to 50 kc).

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Automatic triggering or amplitude selection.

BALANCED DELAY NETWORK

0.25 μsec, superior transient response.

10-KV ACCELERATING POTENTIAL

New Tektronix 5" metallized cathode-roy tube, Type 51P2.

AMPLITUDE CALIBRATOR

0.2 mv to 100 v square wave.

Price (with Type 53A Plug-in Unit)—$1385.

Plug-in Vertical Preamplifiers

Type 53C—Dual-trace Unit

Two identical amplifier channels, activated on alternate sweeps—or nonsynchronously at approximately 100 kc.

Risetime—0.04 μsec.

Bandpass (with Type 535)—DC to 8.5 mc.

Sensitivity—0.05 v/cm to 20 v/cm, ac or dc, in 9 calibrated steps. 0.05 v/cm to 50 v/cm continuously variable. 4-position ac-dc and polarity reversal switches.

Price—$275

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Sensitivity—1 mv/cm to 50 v/cm in 24 calibrated steps. Sensitivity variable between steps.

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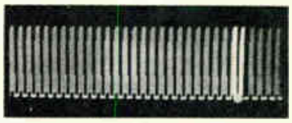
Type 53A—Wide-band DC Unit

Risetime—0.035 μsec.

Bandpass (with Type 535)—DC to 10 mc.

Sensitivity—0.05 v/cm to 20 v/cm, ac or dc, in 9 calibrated steps. 0.05 v/cm to 50 v/cm continuously variable.
Two signal inputs—60 db isolation.

Price—$85



Input signal displayed on the delaying sawtooth. Position of the main sweep is indicated by a bright region on the delaying sawtooth.

Same signal displayed on the main sweep. The main sweep may be ranged out along the delaying sawtooth and positioned to start after the desired amount of delay.

PLEASE WRITE FOR COMPLETE SPECIFICATIONS

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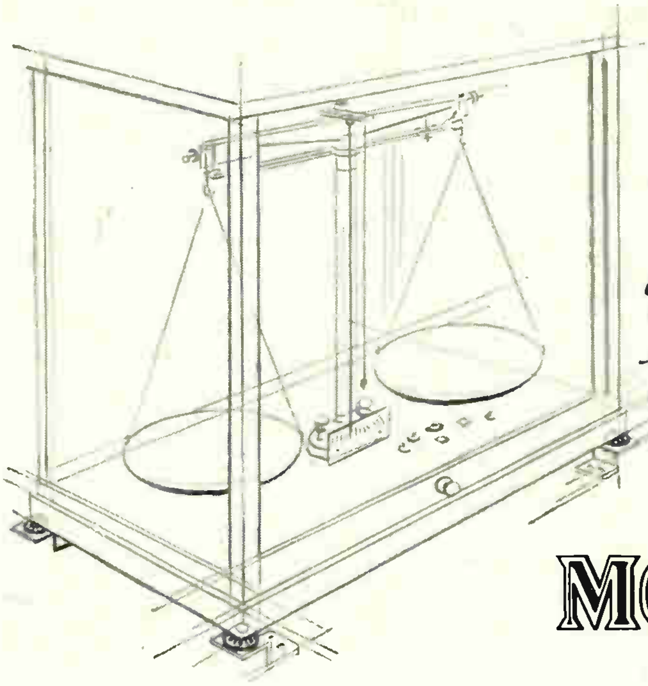

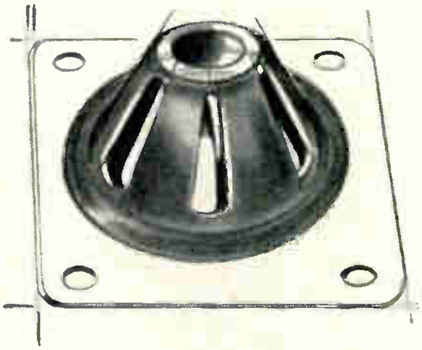


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write for Publication 2654

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Accuracy:	0.001 µF steps ±5% (±1% for an extra charge.) 0.01 µF and 0.1 µF steps ±1% (±0.1% for an extra charge).
Stability:	Within ±0.1%.
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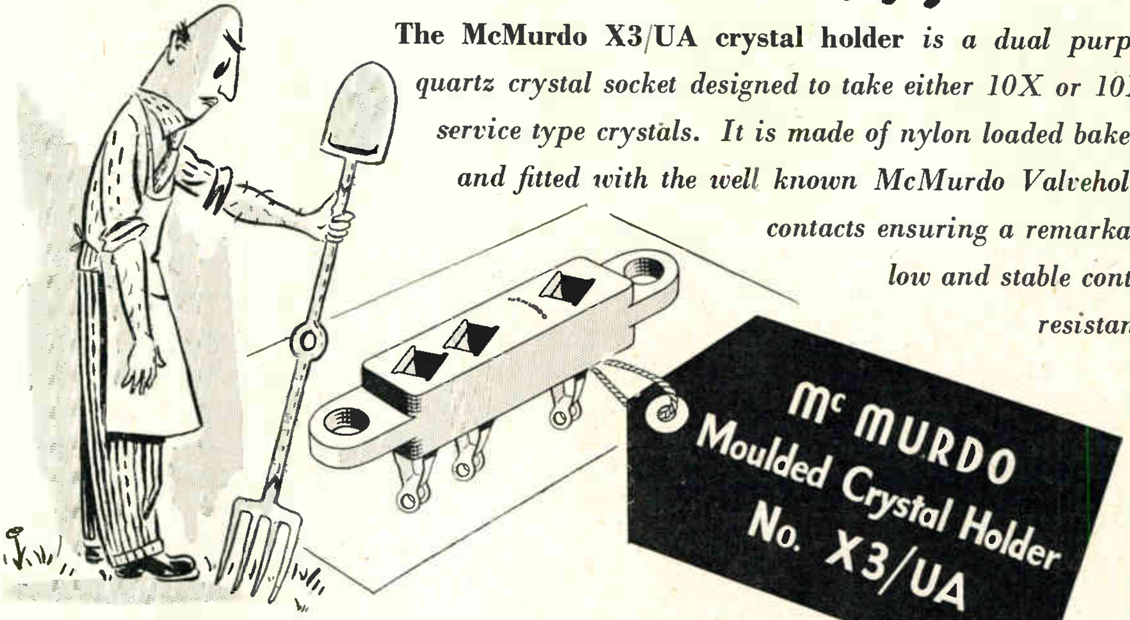

Wholesale Enquiries:—
CYRIL FRENCH HOLDINGS LTD., Hampton Wick, Middlesex. KIN. 2240

Manufacturers' Enquiries: THE McMURDO INSTRUMENT CO. LTD., VICTORIA WORKS, ASHTEAD, SURREY. ASHTEAD 3401


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- **Excellent stability.**
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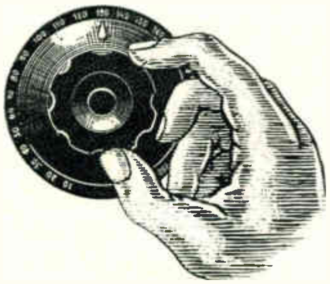

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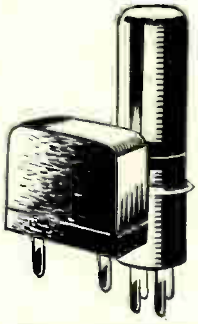

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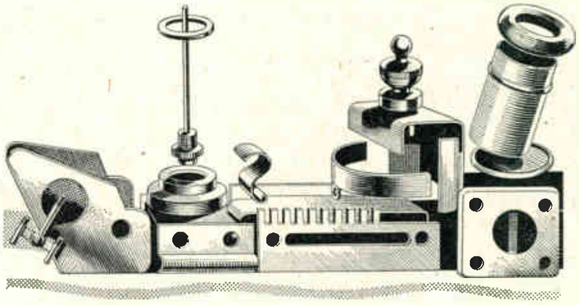

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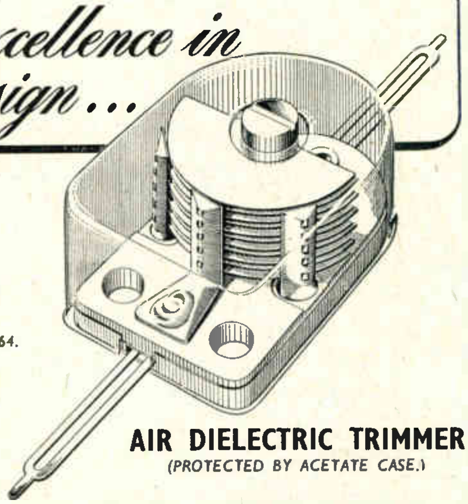

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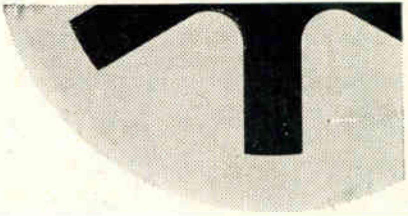

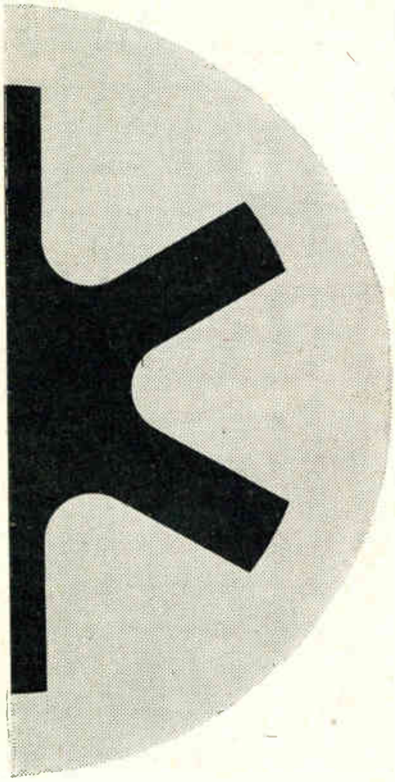

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The Journal of Radio Research and Progress

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Volume 30 · Number 10

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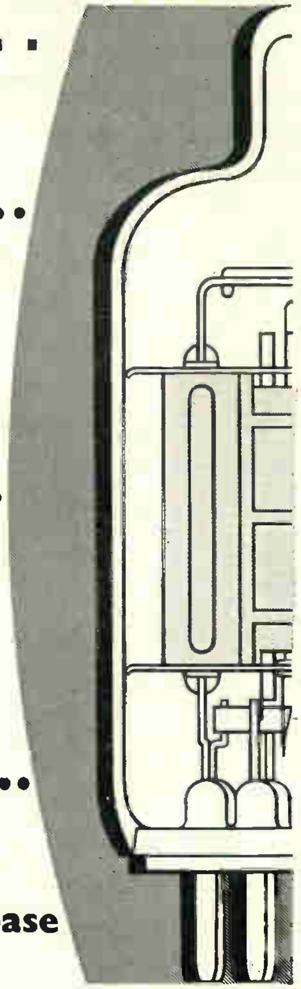
THREE NEW DOUBLE TRIODES . . .

ECC81

ECC82

ECC83



. . . On the Services' Preferred Noval base

TECHNICAL DATA

ECC81

V_a max.	300 V
p_a max.	2.5 W
I_k max.	15mA
g_m (at $V_a = 200$, $I_a = 11.5$mA)	6.4mA/V

The extremely high slope of this valve recommends it for use in grounded grid and cathode R.F. input stages working up to 300 Mc/s. The ECC81 is directly interchangeable with American type 12AT7.

ECC82

V_a max.	300 V
p_a max.	2.75 W
I_k max.	20mA
μ (at $V_a = 250$V, $I_a = 10.5$mA)	17

The ECC82 is a low-μ valve with an anode dissipation of 2.75 watts. These features make it particularly suitable for use as an R.F. oscillator or frequency multiplier. The ECC82 is directly interchangeable with American type 12AU7.

ECC83

V_a max.	300 V
p_a max.	1 W
I_k max.	8mA
μ (at $V_a = 250$V, $I_a = 1.2$mA)	100

An important feature of the ECC83 is its exceptionally high μ. It is an ideal valve for use as a resistance-coupled audio-amplifier, as a phase splitter, or as an inverter. The ECC83 is directly interchangeable with American type 12AX7.

These three double triodes, the latest additions to the Mullard range of noval-based communications valves, provide equipment designers with types suitable for almost every triode application. Features common to all three valves include independent triode sections with separate cathode connections, and centre-tapped heaters that allow either series or parallel wiring (12.6V, 0.15A or 6.3V, 0.3A). Brief descriptions of these triodes are given here. More comprehensive information on these and other valves in the Mullard range of noval-based types will be gladly supplied on request.

Mullard

MULLARD LTD.
COMMUNICATIONS & INDUSTRIAL VALVE DEPT.,
CENTURY HOUSE, SHAFTESBURY AVENUE, LONDON, W.C.2

MVT139

WIRELESS ENGINEER

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Sintered Magnetic Materials

As long ago as 1909 experiments were made with ferrites, and patents taken out by Hilpert, but the results were not very satisfactory and the matter dropped until about 20 years ago when it was revived by Snoek. Since then great advances have been made in the manufacture of sintered non-metallic magnetic material, both magnetically soft and hard, by ceramic methods; as they all have a high resistivity the skin effect is small and they can be used at high frequencies.

The basic materials employed are ferrites; a ferrite is obtained by replacing one of the iron atoms in magnetite or lodestone, Fe_3O_4, by another metal, thus obtaining copper ferrite, $CuFe_2O_4$, manganese ferrite, $MnFe_2O_4$, etc. These materials can now be prepared in such a way that their magnetic properties make them suitable for cores of inductors in telephone apparatus, wave filters, radio receivers and transmitters, etc. Another type of material, magnetically hard, is obtained from the mineral magneto-plumbite, $PbO.6Fe_2O_3$, by replacing the lead by barium, giving $BaFe_{12}O_{19}$.

The crystals of the ferrites have a cubic lattice structure, whereas those of magneto-plumbite have a hexagonal structure. Although the magnetically-hard material has a low retentivity, its coercive force is so great that it is being used in the form of discs as permanent magnets without danger of demagnetization. In all these materials the practical elimination of skin effect makes it possible to study magnetic phenomena, such as ferro-magnetic resonance, within the material at high frequencies. In the soft materials high initial permeability and low hysteresis loss depend on the absence of stresses in the structure of the material, and in this connection it is interesting to note that, on cooling after sintering at a temperature between 1,000° and 1,400°C, the shrinkage with a cubic crystal structure is practically the same in all directions. Although basically cubic the structure of such compounds is naturally somewhat complex, and is known as the spinel structure, 'spinel' being the name given to the mineral $MgAl_2O_4$.

The magnetically-soft ferrites now being employed in the industry are usually a mixture of two different kinds such as Mn and Zn ferrites or Ni and Zn ferrites, depending on the desired results. Mixtures of the latter type have a much higher resistivity (10^5 ohm cm) and can therefore be used at frequencies up to 200 Mc/s but with a greatly reduced permeability. In the former type the initial permeability is between 850 and 1,500 and the maximum value of B between 3,000 and 4,000, whereas some varieties of the latter type have an initial permeability as low as 20 and a maximum induction of less than 2,000.

From an electrical point of view the materials may be regarded as inferior dielectrics with enormous permittivities; the MnZn type may have a permittivity κ of 150,000 and an effective resistance of 100 ohm cm at 1 kc/s, changing to 50,000 and 10 ohm cm at 10 Mc/s. The NiZn type has entirely different values; at 1 kc/s the permittivity may be 400 and the resistivity 5×10^6 ohm cm, while at 10 Mc/s κ may be only 15 and the resistivity $0{\cdot}2 \times 10^6$ ohm cm. The high values of κ in the former type may lead to an unexpected increase in the losses since, combined with the high values of μ, they cause a very low rate of electromagnetic propagation, which in cores of fairly large dimensions may give rise to resonance effects and consequent increased losses; this cannot happen in the NiZn type.

With ordinary iron both the initial and the maximum permeabilities increase gradually with temperature, have a sharp maximum and then at the Curie point, about 770°C, fall practically to zero. The same thing occurs with the ferrites, but at a much lower temperature, the Curie point for the MnZn types lying between 155° and 190°C, and for the various NiZn types increasing from 135° to 550°C as the initial permeability is decreased from 650 to 20. No permanent harm is done to the material by heating it beyond the Curie point.

To render small internal stresses harmless it is advisable to make magnetostriction as small as possible. Fortunately, all the ferrites except one have negative magnetostriction, that is, they contract on magnetization; the exception is magnetite, Fe_3O_4, and by adding a small amount of this, the magnetostriction can be reduced practically to zero, but at the price of a comparatively low resistivity.

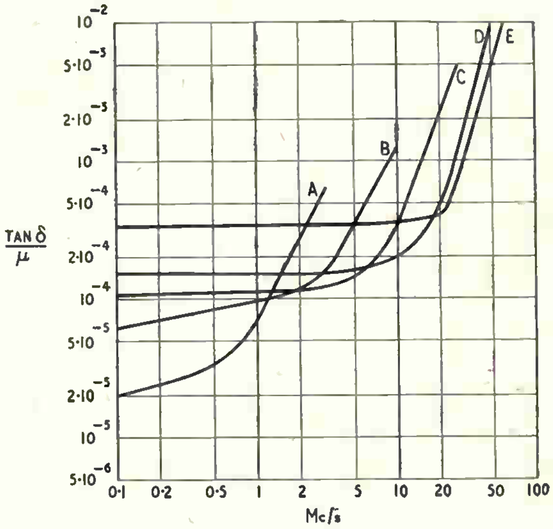

Fig. 1.

After sintering, the component is of a ceramic nature, hard, unaffected by moisture and chemically inert. The raw material is milled into a fine powder and pressed to shape in dies or extruded as rods or tubes, and then sintered in a suitable atmosphere in a furnace. Any alteration of size, which may be necessary to adjust the air-gap in a core, must be done by grinding, a piece of emery paper on a flat surface being often sufficient. As we have seen, high permeabilities can be obtained in MnZn ferrites with Curie points in the neighbourhood of 130°C; the permeabilities are from 10 to 20 times those of iron-dust cores, but they vary considerably with the temperature. This temperature dependence is reduced by making the core with an air-gap, but this also, of course, reduces the effective permeability of the core as a whole. A low Curie point also involves a low value of saturation magnetization, since the magnetization vanishes entirely at the Curie point; it may be between 3,000 and 4,000 gauss at room temperature instead of the 21,000 of ordinary iron. The theory suggested in explanation of this is that in crystals of the spinel type the exchange forces between the atoms do not align all the neighbouring atoms in the same direction, but some in the opposite direction, so that they are anti-parallel but with the majority in the direction of the applied field.

The various types of material are characterized by the quotient $(\tan\,\delta)/\mu$ where δ is the loss angle and μ the effective permeability, allowing for any air-gap. If the core losses are represented by a resistance R in series with the inductance L of the coil, δ is the angle between the total voltage and that across L, and $\tan\,\delta = 1/Q = R/\omega L$. If the gap is increased so that μ is decreased from μ_1 to μ_2 then, with the same current, to maintain B unchanged, the number of turns N would have to be increased so that $\mu_1 N_1 = \mu_2 N_2$, and L would be increased in the same ratio, hence $L_2/L_1 = N_2/N_1 = \mu_1/\mu_2$, and $\mu_1 L_1 = \mu_2 L_2$. The small change in the air gap would not appreciably affect the losses, hence the value of $R/\mu\omega L$, that is, of $(\tan\,\delta)/\mu$ would be unchanged. At low frequencies $\tan\,\delta$ has a value of about 0·01 for all the various types but it increases with the frequency. The value of $(\tan\,\delta)/\mu$ remains practically constant up to a certain frequency, beyond which it increases very rapidly; for this reason many different grades of the material must be made to suit different working frequencies. Materials suitable for very high frequencies have high losses and low permeabilities, but for low frequencies other grades should be used, with small values of $(\tan\,\delta)/\mu$, that is, with low losses and high permeabilities. Fig. 1 shows the variation of $(\tan\,\delta)/\mu$ with frequency for five different grades of material.

Although eddy-current losses may be negligible in the ferrite core, there are still hysteresis losses and residual losses. Assuming low values of B, for which Rayleigh's law is valid, the hysteresis loop resembles a pointed ellipse with a length proportional to B and a width proportional to B^2 and therefore with an area proportional to B^3; hence the loss per cm³ per cycle is proportional to B^3 where B is the maximum induction during the cycle. If R_h is the equivalent series resistance and I the r.m.s. current, $I^2R_h = \text{const.} \times B^3fv$ where v is the volume of the core. For the maximum energy stored in the core we have $I^2L = B^2/\mu.v/8\pi$, and on eliminating I from these two equations we have $R_h/\mu fL = C_hB$ in which C_h is a material constant. If, however, we eliminate B instead of I, since I is easier to measure, we can write $R_h/L = q_2.\sqrt{L}.I.f/800$, in which the frequency is expressed in terms of 800 because this is the standard frequency for telephone measurements. q_2 is not merely a material constant but involves the volume of the core and the effective permeability; we give it here because it has been adopted by the C.C.I.F. and is known as the hysteresis factor; when quoted, the appropriate volume and effective permeability must be stated.

This treatment is all based on Rayleigh's law, which is only valid for low values of B, and quoted values of the constant are not strictly applicable if B exceeds about 30 gauss.

For the total iron loss the above expression becomes

$$\frac{R_{tot}}{\mu fL} = C_e.f + C_hB_{max} + C_r$$

where the first term refers to the eddy-current losses, which we are assuming to be negligible, the second term to the hysteresis losses, which we have just considered, and the third term to the so-called residual losses; the constant C_r is independent of f and B_{max}. The residual losses are supposed to be due to the damping associated with gyromagnetic phenomena of the spinning electrons, which we considered in the Editorial of July 1952 in connection with microwaves in ferrites.

If, as an approximation, one neglects the pointed ends of the Rayleigh hysteresis loop and regards it as an ellipse, both H and B can be regarded as sine functions of time with a phase-angle difference δ which will depend on the ratio of the width to the length of the ellipse. If the permeability μ be divided into real and imaginary components we can write $B = \mu'H - j\mu''H$ in which the first term represents the inductive and the second term the resistive component, and $\tan\delta = \mu''/\mu'$. This has been done for the five grades of ferrite shown in Fig. 1, and the results are plotted in Fig. 2. The full curves refer to μ' and the dotted ones to μ''. It is interesting to note how the real permeability remains practically constant up to a certain frequency and then falls rapidly away after a slight increase in most cases; round about the frequency at which this occurs the value of μ'' undergoes a very great increase.

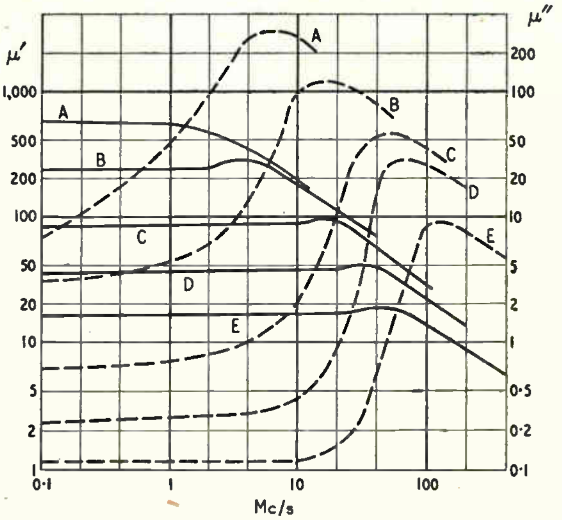

Fig. 2.

Although we have mentioned the magnetically-hard ferrites derived from magneto-plumbite, we have confined our attention mainly to the magnetically-soft varieties. There is no doubt that their development has opened up a wide field in the application of magnetic components in high-frequency apparatus.

G. W. O. H.

TRIODE TRANSFORMATION GROUPS

By A. W. Keen, M.I.R.E., A.M.I.E.E.

SUMMARY.—Group theory is applied to the triode transmission network to obtain transformation groups by means of which the properties of any one method of connection may be derived by routine transformation of those of any other method.

1. Introduction

It has been shown elsewhere[1] that a triode valve, or a multi-grid type operated or connected as a triode, may generally be regarded as a transmission network through which transmission paths may be taken in six possible ways, as indicated in Fig. 1. These may be denoted

gca, cga, agc, gac, cag, acg,

where g, a, c represent grid, anode and cathode, respectively, the first letter-pair being the input and the second pair the output electrodes. Thus gca is the normal amplifier connection, cga the cathode-input or inverted amplifier, agc the series regulator of stabilized h.t. supplies and gac the cathode-follower. Paths cag and acg are employed, in parallel with a control network, as 'dead-end' transmission networks for their input-impedance properties, the latter connection (acg) being typified by the so-called 'electronic reactance valve'.

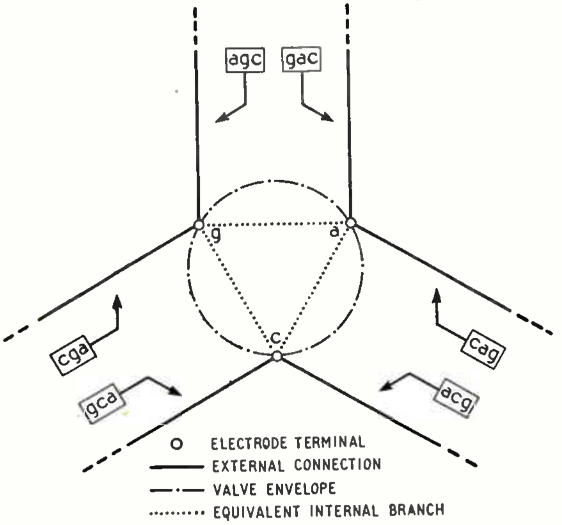

Fig. 1. *Set of available transmission paths in the triode network.*

In so far as connections agc and cag have inverted inputs by comparison with the more normal forms gac and acg they may, by analogy with the description of cag as the inverted amplifier, be designated the inverted cathode-follower and inverted reactance valve, respectively.

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The purpose of the present paper is to present the results of a mathematical study of the inter-relationship of the six operating modes, the aim of which was to discover a method of obtaining from the configuration and electrical properties of the normal amplifier connection (gca), by a regular routine transformation process, those of all other possible methods of connection. By using group theory[2] and matrix algebra[3,4] this objective has been completely achieved, as will be demonstrated by a typical example. The underlying philosophy of the method is similar to that adopted by G. Kron[5] in his tensor analysis of rotating electrical machinery and electrical networks, and can be handled in tensor notation, but the more conventional matrix method of representation will be preferred.

2. Connection Group

Comparison of the set of basic transmission forms of the triode network given in Fig. 2 shows that any one of these may be obtained from any other member of the set by re-orientation of the triode with respect to the network terminals (c.f. also, Fig. 6). The complete set of transformations required may be specified as follows, using bold face to avoid confusion with the use of the same letter groups for the transmission path in cases (2) and (3):—

(1) **I** $= a \to a, c \to c, g \to g$ (no change)
(2) **gca** $= g \to c, c \to a, a \to g$ or $g \to c \to a \to g$
(3) **gac** $= g \to a, a \to c, c \to g$ or $g \to a \to c \to g$
(4) **ga** $= g \to a, a \to g, c \to c$; i.e., $g \to a \to g$ or $g \leftrightarrow a$
(5) **gc** $= g \to c, c \to g, a \to a$; i.e., $g \to c \to g$ or $g \leftrightarrow c$
(6) **ac** $= a \to c, c \to a, g \to g$; i.e., $a \to c \to a$ or $a \leftrightarrow c$.

Taking **gca** as an example, this representation implies that the grid is connected to the network terminal previously connected to the cathode ($g \to c$), the cathode is connected to the terminal previously connected to anode ($c \to a$), and the latter to the terminal previously connected to the grid ($a \to g$). **gca** is, therefore, a cyclical interchange of connections and involves rotation of the triode, by one step, with respect to the external network. **gac** is the interchange of connections produced by rotation one step in the opposite direction. **ga, gc,** and **ac** are simple transpositions of the electrodes concerned leaving the connection of the remaining electrode (e.g., the cathode in the case **ga**) unchanged. The 'transformation' **I**, which leaves the connections unchanged, is called the identity transformation. It is of no practical value in the present context but must be included for mathematical consistency. Thus, if **I** is included, the six transformations, which will be called:—

(1) Identity transformation (**I**)
(2) Anti-clockwise rotation (**gca**)
(3) Clockwise rotation (**gac**)
(4) Grid-anode transposition (**ga**)
(5) Grid-cathode transposition (**gc**)
(6) Anode-cathode transposition (**ac**)

respectively, constitute in the strict mathematical sense a group, in particular a symmetrical group of order six, principally because if any number of them are performed in turn, in arbitrary order, their resultant will always be one of the members of the group.

A clear picture of the results of applying these transformations to the triode operating in any particular network may be obtained from a 'transformation network' in which each node represents one of the six possible methods of connection noted in Section 1 and is joined to every other node by a set of branches, one for each type of transformation. Omitting the identity transformation one obtains the diagram shown in Fig. 3. Thus the set of transformations **gca, gac, ga, gc, ac** convert the normal amplifier connection gca into inverted reactance (cag), inverted follower (agc), reactance valve (acg), inverted amplifier (cga) and cathode-follower (gac) respectively. One cycle of clockwise rotation (transformation **gca** repeated twice), starting with the gca connection, gives the sequence: normal amplifier (gca), inverted reactance (cag), inverted follower (agc) and normal amplifier again. Writing a sequence of transformations as a continued product one can obtain the cathode-follower (for example) from the normal amplifier in such ways as **gc. ga. gc; gc. gac; gca. gc; ac; gac. ga; ga. gc. ga** all of which have the single resultant **ac.** It will be clear that any sequence of transformations may be traced as a continuous path in this network and will always have a single transformation of the set as resultant; this is the characteristic group property.[2]

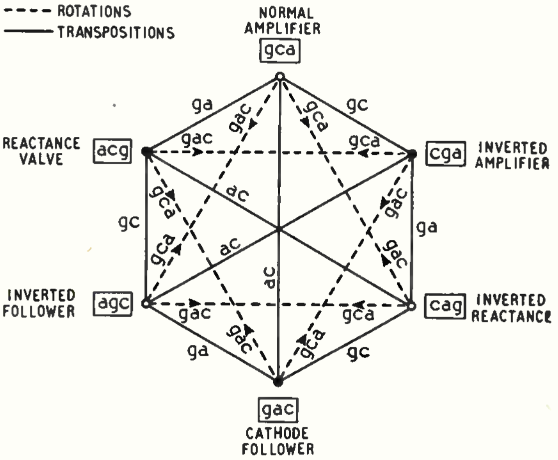

Fig. 3 (above). *Connection transformation network.*

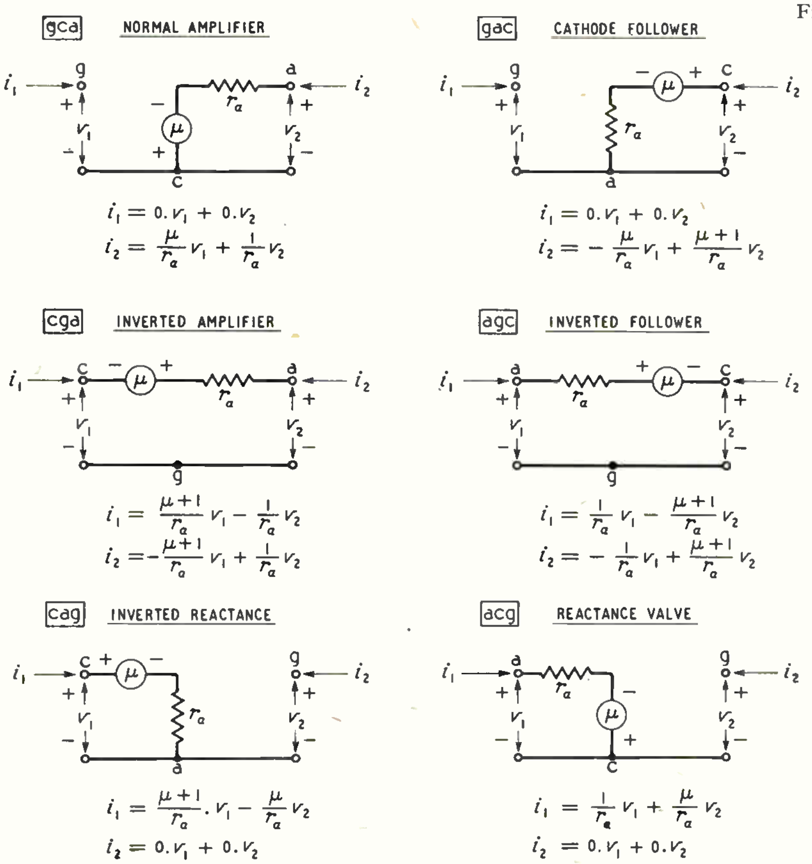

Fig. 2. *Basic triode transmission networks; current-voltage equations and equivalent networks.*

3. Admittance Matrix Transformation Group

The feasibility of deriving all six methods of connection from any one particular (the 'reference') connection suggests the corresponding possibility of obtaining the electrical properties for all cases from a reference case. For this purpose it is necessary to find a suitable algebraic characterization of the triode and to determine the set of algebraic transformations which correspond one-to-one with the geometric transformation group.

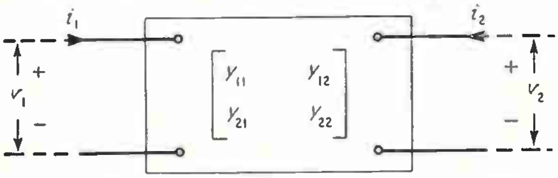

Fig. 4. *Matrix notation for the generalized quadripole.*

In general, the valve and its associated network may be represented by a four-element matrix, as shown, using the conventional notation, in Fig. 4. Any two of the four terminal electrical quantities (v_1, i_1, v_2, i_2) may be combined into a two-element vector from which the other pair may be obtained as a dependent vector by linear transformation with the appropriate matrix representing the action of the valve circuit. For example, taking input and output voltages (v_1, v_2) as the independent (i.e., given) pair, pre-multiplication by the valve admittance matrix yields the dependent pair of currents (i_1, i_2), thus:—

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \cdot \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \quad \ldots \quad \ldots \quad (1)$$

Since the two independent quantities may be taken in six different ways there is a choice of six distinct matrices[3] for the representation of the valve, any one of which may be obtained from any other by linear transformation, but for the present only the admittance matrix will be considered.

TABLE 1

Triode admittance matrices Y_{gca}, etc.

$\begin{bmatrix} 0 & 0 \\ \frac{\mu}{r_a} & \frac{1}{r_a} \end{bmatrix}$ *gca*	$\begin{bmatrix} 0 & 0 \\ -\frac{\mu}{r_a} & \frac{\mu+1}{r_a} \end{bmatrix}$ *gac*
$\begin{bmatrix} \frac{\mu+1}{r_a} & -\frac{1}{r_a} \\ -\frac{\mu+1}{r_a} & \frac{1}{r_a} \end{bmatrix}$ *cga*	$\begin{bmatrix} \frac{1}{r_a} & -\frac{\mu+1}{r_a} \\ -\frac{1}{r_a} & \frac{\mu+1}{r_a} \end{bmatrix}$ *agc*
$\begin{bmatrix} \frac{\mu+1}{r_a} & -\frac{\mu}{r_a} \\ 0 & 0 \end{bmatrix}$ *cag*	$\begin{bmatrix} \frac{1}{r_a} & \frac{\mu}{r_a} \\ 0 & 0 \end{bmatrix}$ *acg*

The admittance matrices for the six triode connections may be readily obtained by conventional analysis (see Fig. 2) and are given in the accompanying Table 1. It may be checked that all are of the form:—

$$\begin{bmatrix} \alpha_1 & \alpha_2 + \beta_1 \\ \alpha_2 + \beta_2 & \alpha_3 \end{bmatrix} \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

where the elements α_i, β_i have the values given in Table 2 for the different methods of connection. Any one set of these values is obtainable from any other set by transpositional inter-changes or cyclical permutations corresponding to the connection transpositions and rotations distinguished in (1) of Section 2. Taking first the α coefficients it is found that under:—

(1) Identity transformation.

When a→c, c→c, g→g

then $\alpha_1 \rightarrow \alpha_1$, $\alpha_2 \rightarrow \alpha_2$, $\alpha_3 \rightarrow \alpha_3$

(2) Anticlockwise rotation.

When g→c, c→a, a→g

then $\alpha_1 \rightarrow \alpha_2$, $\alpha_2 \rightarrow \alpha_3$, $\alpha_3 \rightarrow \alpha_1$

(3) Clockwise rotation.

When g→a, a→c, c→a

then $\alpha_1 \rightarrow \alpha_3$, $\alpha_3 \rightarrow \alpha_2$, $\alpha_2 \rightarrow \alpha_1$

(4) Grid-anode transposition.

When g↔a, then $\alpha_1 \leftrightarrow \alpha_3$

(5) Grid-cathode transposition.

When g↔c, then $\alpha_1 \leftrightarrow \alpha_2$

(6) Anode-cathode transposition.

When a↔c, then $\alpha_3 \leftrightarrow \alpha_2$

Thus the α_i transform in exactly the same manner as the network connections and the α transformations may be denoted in the same way as the

TABLE 2

Admittance matrix elements

	α_1	α_2	α_3	β_1	β_2
gca	0	$\frac{\mu+1}{r_a}$	$\frac{1}{r_a}$	$-\frac{\mu+1}{r_a}$	$-\frac{1}{r_a}$
cga	$\frac{\mu+1}{r_a}$	0	$\frac{1}{r_a}$	$-\frac{1}{r_a}$	$-\frac{\mu+1}{r_a}$
cag	$\frac{\mu+1}{r_a}$	$\frac{1}{r_a}$	0	$-\frac{\mu+1}{r_a}$	$-\frac{1}{r_a}$
gac	0	$\frac{1}{r_a}$	$\frac{\mu+1}{r_a}$	$-\frac{1}{r_a}$	$-\frac{\mu+1}{r_a}$
agc	$\frac{1}{r_a}$	0	$\frac{\mu+1}{r_a}$	$-\frac{\mu+1}{r_a}$	$-\frac{1}{r_a}$
acg	$\frac{1}{r_a}$	$\frac{\mu+1}{r_a}$	0	$-\frac{1}{r_a}$	$-\frac{\mu+1}{r_a}$

changes of connections; i.e.,

$$\mathbf{I,\ gca,\ gac,\ ga,\ gc,\ ac}$$

$$\mathbf{I},\ \alpha_1\alpha_2\alpha_3,\ \alpha_1\alpha_3\alpha_2,\ \alpha_1\alpha_3,\ \alpha_1\alpha_2,\ \alpha_3\alpha_2 \qquad (3)$$

The β_i transform in a simpler manner for β_1, β_2 remain unchanged under rotation of the triode connections, but interchange under any transposition of the triode terminals. The transformation group is, therefore,

$$\mathbf{I},\ \mathbf{I},\ \mathbf{I},\ \beta_1\beta_2,\ \beta_1\beta_2,\ \beta_1\beta_2 \qquad \ldots \quad \ldots \quad (4)$$

Alternatively, one may treat the matrix as a single entity Y and pass from one case to another by linear transformation. Any one of the six admittance matrices may be obtained as a congruent transform[4] of any other member of the set by a transformation **T** such that

$$\mathbf{T}' . Y_1 . \mathbf{T} = Y_2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

where Y_1, Y_2 are any two of the admittance matrices and **T'** is the transpose of **T**; i.e., the transformation whose matrix is obtained from that of **T** by transposition of its rows and columns. The required **T**-matrices are

$$\mathbf{T_I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}_{\mathbf{I}} \mathbf{T_{gca}} = \begin{bmatrix} -1 & 1 \\ -1 & 0 \end{bmatrix}_{\mathbf{gca}} \mathbf{T_{gac}} = \begin{bmatrix} 0 & -1 \\ 1 & -1 \end{bmatrix}_{\mathbf{gac}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

$$\mathbf{T_{ga}} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}_{\mathbf{ga}} \mathbf{T_{gc}} = \begin{bmatrix} -1 & 0 \\ -1 & 1 \end{bmatrix}_{\mathbf{gc}} \mathbf{T_{ac}} = \begin{bmatrix} 1 & -1 \\ 0 & -1 \end{bmatrix}_{\mathbf{ac}}$$

using the subscripts to show the correspondence with the connection transformations (1)-(6) in Section 2 and in Table 3. These matrices constitute a group under matrix multiplication since the product of any two is in all cases a third member of the set. Substituting each in turn in (5) together with (2) gives:—

$$\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} = \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} \quad \ldots \quad (7.1)$$

$$\begin{bmatrix} -1 & -1 \\ 1 & 0 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} -1 & 1 \\ -1 & 0 \end{bmatrix} = \begin{bmatrix} \alpha_2 & \alpha_3+\beta_1 \\ \alpha_3+\beta_2 & \alpha_1 \end{bmatrix} \quad (7.2)$$

$$\begin{bmatrix} 0 & 1 \\ -1 & -1 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} 0 & -1 \\ 1 & -1 \end{bmatrix} = \begin{bmatrix} \alpha_3 & \alpha_2+\beta_1 \\ \alpha_1+\beta_2 & \alpha_2 \end{bmatrix} \quad (7.3)$$

$$\begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix} = \begin{bmatrix} \alpha_3 & \alpha_2+\beta_2 \\ \alpha_2+\beta_1 & \alpha_1 \end{bmatrix} \quad \ldots \quad (7.4)$$

$$\begin{bmatrix} -1 & -1 \\ 0 & 1 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} -1 & 0 \\ -1 & 1 \end{bmatrix} = \begin{bmatrix} \alpha_2 & \alpha_1+\beta_2 \\ \alpha_1+\beta_2 & \alpha_3 \end{bmatrix} \quad (7.5)$$

$$\begin{bmatrix} 1 & 0 \\ -1 & -1 \end{bmatrix} . \begin{bmatrix} \alpha_1 & \alpha_2+\beta_1 \\ \alpha_2+\beta_2 & \alpha_3 \end{bmatrix} . \begin{bmatrix} 1 & -1 \\ 0 & -1 \end{bmatrix} = \begin{bmatrix} \alpha_1 & \alpha_3+\beta_2 \\ \alpha_3+\beta_1 & \alpha_2 \end{bmatrix} \quad (7.6)$$

For example, operating on the admittance matrix representing gca transmission, with more compact notation, one finds in turn:

$$\mathbf{T'_I} . Y_{\text{gca}} . \mathbf{T_I} = Y_{\text{gca}} \quad \ldots \quad \ldots \quad (8.1)$$

$$\mathbf{T'_{gca}} . Y_{\text{gca}} . \mathbf{T_{gca}} = Y_{\text{cag}} \quad \ldots \quad \ldots \quad (8.2)$$

$$\mathbf{T'_{gac}} . Y_{\text{gca}} . \mathbf{T_{gac}} = Y_{\text{agc}} \quad \ldots \quad \ldots \quad (8.3)$$

$$\mathbf{T'_{ga}} . Y_{\text{gca}} . \mathbf{T_{ga}} = Y_{\text{acg}} \quad \ldots \quad \ldots \quad (8.4)$$

$$\mathbf{T'_{gc}} . Y_{\text{gca}} . \mathbf{T_{gc}} = Y_{\text{cga}} \quad \ldots \quad \ldots \quad (8.5)$$

$$\mathbf{T'_{ac}} . Y_{\text{gca}} . \mathbf{T_{ac}} = Y_{\text{gac}} \quad \ldots \quad \ldots \quad (8.6)$$

where the $\mathbf{T_i}$ are the transformation matrices of (6) and the Y_i the admittance matrices. Operation on any other one of the latter matrices in a similar manner yields the complete set of these matrices, but in a different order. All of these results may, of course, be checked by substitution for the elements α_i, β_i from Table 2.

All the transformation groups derived have been collected together in Table 3 to facilitate their use on electrical functions of the triode, as will be illustrated in Section 4.

TABLE 3

Corresponding transformation groups

I	gca	gac	ga	gc	ac
I	$\alpha_1\alpha_2\alpha_3$	$\alpha_1\alpha_3\alpha_2$	$\alpha_1\alpha_3$	$\alpha_1\alpha_2$	$\alpha_3\alpha_2$
I	I	I	$\beta_1\beta_2$	$\beta_1\beta_2$	$\beta_1\beta_2$
$\mathbf{T_I}$	$\mathbf{T_{gca}}$	$\mathbf{T_{gac}}$	$\mathbf{T_{ga}}$	$\mathbf{T_{gc}}$	$\mathbf{T_{ac}}$

The $\mathbf{T_i}$ are written in abbreviation of the transformations $\mathbf{T'_i} \cdot \mathbf{Y} \cdot \mathbf{T_i}$ used in Equ. (8).

4. Voltage Gain and Input Admittance Functions

An advantage of deriving a single form which, under appropriate transformation, will give the complete set of particular cases is that any function of the form, such as the determinant of a matrix, will follow the same transformation rules. Two important electrical functions are the voltage gain G_{12} and input admittance Y_{11}. Both of these may be expressed in terms of the admittance matrix elements, so that if the matrix is known in terms of transformable elements, as at (2), it is possible to obtain corresponding transformable expressions for the two functions.

In the general case of any transmission network operating into an arbitrary load impedance

R, we have:—

$$G_{12} = \frac{-y_{21} \, . \, R}{1 + y_{22} \, . \, R} \quad .. \quad .. \quad .. \quad (9.1)$$

$$Y_{11} = \frac{y_{11} + \Delta Y . R}{1 + y_{22} \, . \, R}, \ \Delta Y = y_{11}y_{22} - y_{12}y_{21} \quad (9.2)$$

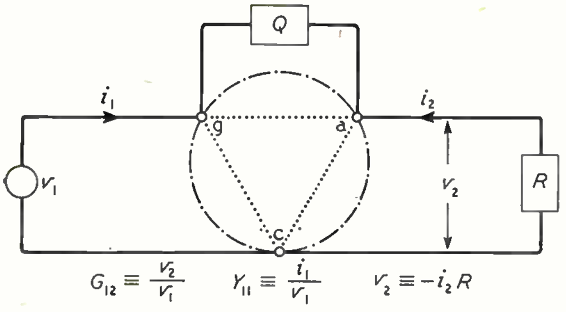

Fig. 5. *Basic circuit of typical triode network, shown for reference case of* gca *transmission.*

Taking, as an elementary example, a transmission network consisting of a linear negative-grid triode having an arbitrary impedance Q bridging input and output terminals (Fig. 5), as for the application of shunt negative feedback, one obtains the resultant network admittance matrix by addition of the two component matrices, thus

$$\begin{bmatrix} \alpha_1 & \alpha_2 + \beta_1 \\ \alpha_2 + \beta_2 & \alpha_3 \end{bmatrix} + \begin{bmatrix} Q' & -Q' \\ -Q' & Q' \end{bmatrix}$$
$$= \begin{bmatrix} \alpha_1 + Q' & \alpha_2 + \beta_1 - Q' \\ \alpha_2 + \beta_2 - Q' & \alpha_3 + Q' \end{bmatrix}, Q' = \frac{1}{Q}$$

whence

$$y_{11} = \alpha_1 + Q' \qquad y_{12} = \alpha_2 + \beta_1 - Q'$$
$$y_{21} = \alpha_2 + \beta_2 - Q' \qquad y_{22} = \alpha_2 \, Q'$$

Using the results:

$$\alpha_1 + \alpha_2 + \alpha_3 + \beta_1 + \beta_2 = 0$$
$$\alpha_1 \, \alpha_3 - (\alpha_2 + \beta_1)(\alpha_2 + \beta_2) = 0$$

which remain invariant under the group of transformations α, β the determinant reduces to $\Delta Y = \alpha Q'$

Substituting these results into (9.1) and (9.2) gives:—

$$G_{12} = \frac{\{1 - (\alpha_2 + \beta_2)Q\}R}{Q + \alpha_3 \, QR + R} \quad .. \quad .. \quad (10.1)$$

$$Y_{11} = \frac{1 + \alpha_1 \, Q + \alpha_2 \, R}{Q + \alpha_3 \, QR + R} \quad .. \quad .. \quad .. \quad (10.2)$$

Starting with the reference case (gca) for which

$$\alpha_1 = 0, \quad \alpha_2 = \frac{\mu + 1}{r_a}, \quad \alpha_3 = \frac{1}{r_a};$$

$$\beta_1 = -\frac{\mu + 1}{r_a}, \quad \beta_2 = -\frac{1}{r_a}$$

one obtains:

$$G_{12}(\text{gca}) = \frac{(r_a - \mu Q)R}{r_aQ + QR + Rr_a} \quad .. \quad .. \quad (11.1)$$

$$Y_{11}(\text{gca}) = \frac{r_a + (\mu + 1)R}{r_aQ + QR + Rr_a} \quad .. \quad .. \quad (11.2)$$

Rotating the triode anticlockwise (transformation **gca**) and carrying out the corresponding **α, β** transformations **($\alpha_1\alpha_2\alpha_3$, 1)** on the α, β elements gives:—

$$G_{12}(\text{cag}) = \frac{R}{Q + R} \quad .. \quad .. \quad .. \quad (12.1)$$

$$Y_{11}(\text{cag}) = \frac{r_a + (\mu + 1)Q + R}{r_a(Q + R)} \quad .. \quad (12.2)$$

Continuing in this manner one obtains in turn:—

$$G_{12}(\text{agc}) = \frac{(r_a + Q)R}{r_a(Q + R) + (\mu + 1)QR} \quad .. \quad (13.1)$$

$$Y_{11}(\text{acg}) = \frac{(r_a + Q)}{r_a(Q + R) + (\mu + 1)QR} \quad .. \quad (13.2)$$

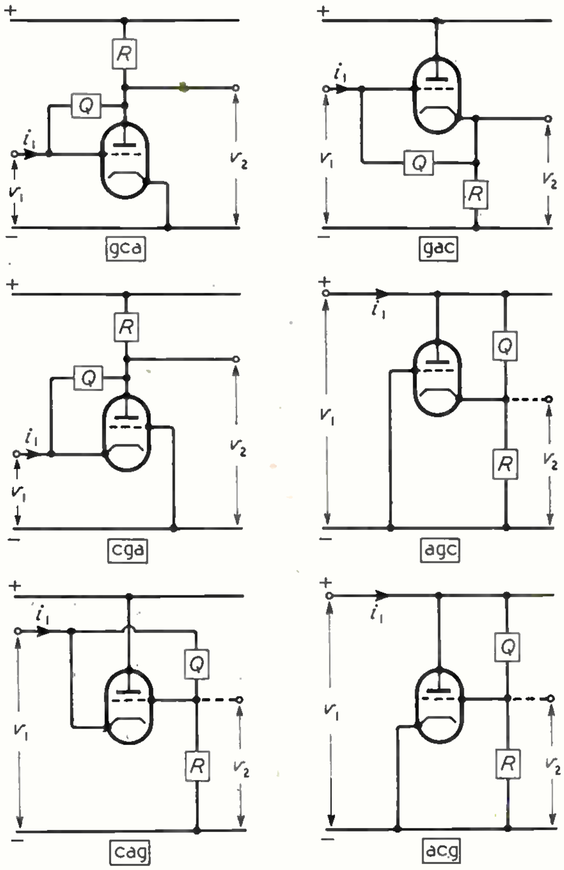

Fig. 6. *Set of six circuits obtainable from the basic configuration shown in Fig. 5.*

$$G_{12}(\mathrm{acg}) = \frac{R}{Q + R} \qquad \ldots \quad \ldots \quad \ldots \quad (14.1)$$

$$Y_{11}(\mathrm{acg}) = \frac{r_a + Q + (\mu + 1)R}{r_a(Q + R)} \qquad \ldots \quad (14.2)$$

$$G_{12}(\mathrm{cga}) = \frac{\{r_a + (\mu + 1)Q\}R}{r_a(Q + R) + QR} \qquad \ldots \quad (15.1)$$

$$Y_{11}(\mathrm{cga}) = \frac{r_a + (\mu + 1)Q}{r_a(Q + R) + QR} \quad \ldots \quad \ldots \quad (15.2)$$

$$G_{12}(\mathrm{gac}) = \frac{(r_a + \mu Q)R}{r_a(Q + R) + (\mu + 1)QR} \qquad (16.1)$$

$$Y_{11}(\mathrm{gac}) = \frac{r_a + R}{r_a(Q + R) + (\mu + 1)QR} \qquad (16.2)$$

The set of six circuits are shown in skeleton form in Fig. 6.

Since the transformations used in this process are characteristic only of the triode and are independent of its associated network they may be applied regardless of the complexity of the circuit. The method is useful not only as a means of establishing new results but also for checking calculations obtained by more conventional methods of analysis.

5. Conclusion

It has been shown that the electrical characteristics of the six distinct circuits obtainable from any particular triode network configuration by connection of the triode to the remainder of the network in all possible ways are interrelated in such a manner that, given the characteristics of any one connection, those of the remainder are obtainable by a regular routine transformation process.

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MATCHING DISCONTINUITIES IN WAVEGUIDES

By J. C. Parr, B.Sc., Graduate I.E.E.

Introduction

In general, the accuracy of an experimental measurement will not be greater than that of the least accurate piece of measuring equipment; in particular, exact measurement of the standing-wave ratio in an unmatched length of waveguide cannot be made unless the inherent standing-wave ratio of the measuring equipment is negligible in comparison with the standing-wave ratio to be determined. Further, it is usually impossible to reduce the standing-wave ratio of an unmatched component to zero (i.e., to match the component) if imperfect measuring equipment is used.

Except for the few special cases described later, it is necessary to match the measuring equipment before it can be used to match other components and, if the measuring equipment is perfectly matched, it is still necessary to match these other components. The most common matching technique is that of inserting a number of tuning screws at more or less regular intervals along the waveguide, and adjusting these screws by trial until the required degree of matching has been obtained. Unless the tuning screws are inserted at certain definite intervals and are adjusted systematically, this matching process is laborious and may even be impossible.

This paper describes a simple systematic method of matching both measuring equipment and other unmatched sections of waveguide, all possible practicable methods being dealt with. Those cases in which it is possible to match the unmatched section without previously matching the measuring equipment are also considered.

Basic Symbols and Equations

The symbols that will be required in those sections which contain mathematical analyses are:

x = distance of an arbitrary point from the generator;
Z = impedance at a distance x from the generator;
ρ = reflection coefficient at a distance x from the generator;
$|E|$ = magnitude of electric field at a distance x from the generator;
β = phase constant of the waveguide considered;
λ_g = guide wavelength in the waveguide considered;
s = standing-wave ratio corresponding to ρ;
Z_0 = characteristic impedance of the waveguide.

The suffix "1" (e.g., for Z_1, ρ_1) denotes quantities measured at the generator, and the suffix "2" denotes quantities measured at a point distant L from the generator ($L \geqslant x$).

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The following equations will be required:

$$\beta = 2\pi/\lambda_g \quad \ldots \quad (1)$$

$$\rho = R.e^{j\phi} \quad \ldots \quad (2)$$

$$\rho = (Z - Z_0)/(Z + Z_0) \quad \ldots \quad (3)$$

$$\left.\begin{array}{l} \rho = \rho_1 . e^{j2\beta x} \\ \rho = \rho_2 . e^{-j2\beta(L-x)} \end{array}\right\} \quad \ldots \quad (4)$$

$$s = (1 - R)/(1 + R) \quad \ldots \quad (5)$$

$$s = |E|_{min}/|E|_{max} \quad \ldots \quad (6)$$

$$|E| = |E|_{min} \text{ at } \phi + 2\beta x = \pm\pi, \pm 3\pi, \pm 5\pi \quad (7a)$$

$$|E| = |E|_{max} \text{ at } \phi + 2\beta x = 0, \pm 2\pi, \pm 4\pi \quad \ldots \quad (7b)$$

$$\frac{Z}{Z_0} = \frac{Z_1 - jZ_0 \tan(\beta x)}{Z_0 - jZ_1 \tan(\beta x)} \quad \ldots \quad (8a)$$

$$\frac{Z}{Z_0} = \frac{Z_1 + jZ_0 \tan(\beta y)}{Z_0 + jZ_2 \tan(\beta y)} \quad \ldots \quad (8b)$$

where $y = L - x$.

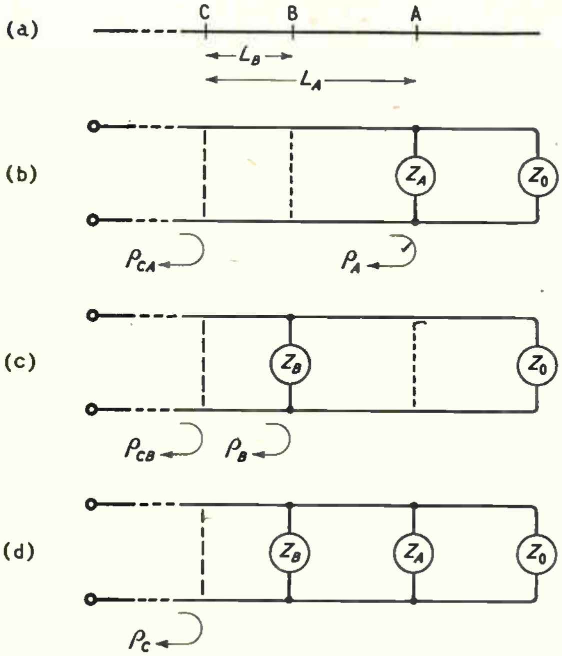

Fig. 1. *Length of waveguide (a) with discontinuities at A and B. The individual equivalent circuits are (b) and (c) and the combined one for both discontinuities is (d).*

Combination of Reflection Coefficients

Suppose that a single shunt impedance Z_A is the only discontinuity in an otherwise matched length of waveguide (Fig. 1). The total impedance at the point A is that formed by Z_A and the termination impedance Z_0 transformed to A [by equation (8b)], or, since the transformed impedance is still Z_0, the total impedance at A is $Z_A Z_0/(Z_A + Z_0)$. This impedance defines the reflection coefficient at A, ρ_A, by equation (3):

$$\rho_A = -Z_0/(2Z_A + Z_0).$$

Finally, the reflection coefficient at C is given in terms of ρ_A and L_A (the distance from A to C) as

$$\rho_{CA} = \rho_A . e^{-j2\beta L_A}$$

Similarly, if the only discontinuity in the waveguide is the shunt impedance Z_B at B, we may find the equivalent reflection coefficient at C in the same way as when Z_A was the only shunt impedance:

$$\rho_{CB} = \rho_B . e^{-j2\beta L_B},$$

$$\rho_B = -Z_0/(2Z_B + Z_0).$$

If we now consider both Z_A and Z_B, acting together, we may find the total impedance at A, as before; then transform this to B, from equation (8b); then add Z_B in parallel; then find the reflection coefficient at B; and finally we may find the equivalent reflection coefficient at C, ρ_C. After carrying out these operations, we find that we can express ρ_C in terms of ρ_{CA} and ρ_{CB} as

$$\rho_C = \frac{\rho_{CA} + \rho_{CB} + 2.\rho_{CA}.\rho_{CB}.e^{j2\beta L_A}}{1 - \rho_{CA}.\rho_{CB}.e^{j4\beta L_B}}$$

(Note that this expression is valid only when A, B, C are taken in that order towards the generator.)

The last equation gives the reflection coefficient at any point in a waveguide produced by two shunt impedances at two definite points, in terms of the reflection coefficients produced at the point by each impedance acting without the other. The only application of this equation with which we shall be concerned is that derived by assuming that the total reflection coefficient is small compared with unity, when, to a first approximation, we may assume that

$$\rho_C = \rho_{CA} + \rho_{CB} \quad \ldots \quad (9)$$

i.e., we may assume that small reflection coefficients at an arbitrary point are additive.

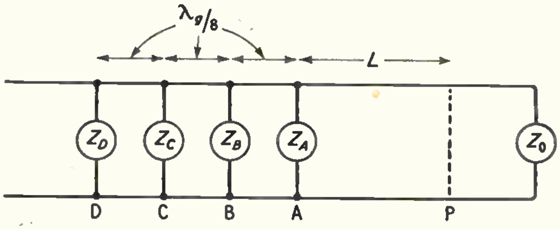

Fig. 2. *This diagram shows the equivalent shunt impedances of a four-screw matching section.*

Theory of Four-Screw Tuning

Suppose that in an otherwise matched length of waveguide there are four shunt impedances Z_A, Z_B, Z_C, Z_D, situated at the points A, B, C, D (Fig. 2). Suppose also that these impedances are all large and are all purely capacitive, when we may write

$$Z_A = Z_0/ja, \quad Z_B = Z_0/jb,$$

$$Z_C = Z_0/jc, \quad Z_D = Z_0/jd,$$

where a, b, c, d are real numbers which are small compared with unity. Finally, suppose that the distance between consecutive impedances is one-eighth of a guide wavelength.

Ignoring for the moment Z_B, Z_C, and Z_D, the total impedance at A is that produced by Z_A and Z_0 in parallel, or

$$Z_{AT} = Z_0/(1 + ja)$$

and the corresponding reflection coefficient at A is

$$\rho_A = -j\tfrac{1}{2}a/(1 + j\tfrac{1}{2}a)$$

from equation (3). Similarly, neglecting Z_A, Z_O, and Z_D, the reflection coefficient at B produced by Z_B is

$$\rho_B = -j\tfrac{1}{2}b/(1 + j\tfrac{1}{2}b)$$

and we can write down the expressions for ρ_C and ρ_D.

Since a, b, c, d are all small compared with unity, we can write, to a first approximation,

$$\rho_A = -j\tfrac{1}{2}a, \quad \rho_B = -j\tfrac{1}{2}b,$$
$$\rho_C = -j\tfrac{1}{2}c, \quad \rho_D = -j\tfrac{1}{2}d.$$

Now suppose that we transform all these reflection coefficients to the point P, distant L from A in the direction of the termination. From (4), the corresponding reflection coefficients at P will be:

$$\rho_{PA} = (-j\tfrac{1}{2}a).e^{j2\beta L}$$
$$\rho_{PB} = (-j\tfrac{1}{2}b).e^{j2\beta L}.e^{j2\beta(\lambda_g/8)}$$
$$\rho_{PO} = (-j\tfrac{1}{2}c).e^{j2\beta L}.e^{j2\beta(2\lambda_g/8)}$$
$$\rho_{PD} = (-j\tfrac{1}{2}d).e^{j2\beta L}.e^{j2\beta(3\lambda_g/8)}$$

From (1) we see that

$$e^{j2\beta(\lambda_g/8)} = e^{j2\beta(\pi/4\beta)} = e^{j\pi/2} = j,$$

and, from (9), the total reflection coefficient at P is

$$\rho_P = \tfrac{1}{2}(-ja + b + jc - d).e^{j2\beta L} \quad .. \quad (10)$$

It can be seen that, by suitable choice of a, b, c, and d, ρ_P can be made to have any desired complex value for any given value of L. All four variables a, b, c, d are necessary, since none of them can be negative. (A tuning screw can be inductive under certain circumstances, but this need not concern us here.)

The practical meaning of the equation above is that we are able to produce a reflection coefficient of any arbitrary complex value at any point in a matched waveguide, by inserting certain definite shunt capacitances at certain definite intervals. It might be pointed out that although $\tfrac{1}{8}$-wavelength was used as the spacing between impedances in the analysis, the same conclusions are true if the spacing is $\tfrac{3}{8}$, $\tfrac{5}{8}$, $\tfrac{7}{8}$, etc., of a wavelength. It is explained below that the larger spacings are inconvenient under usual conditions, but the smallest spacing may be equally inconvenient if the waveguide dimensions are small.

Components to be Matched

The section of waveguide to be matched, whether or not it is part of the measuring equipment, can be either of two possible types: it may be a terminating section, such as a crystal holder, imperfect 'matched load,' or thermocouple; or it may be an 'insertion device', such as a corner, bend, taper, twist, attenuator, directional coupler, or T-junction. For the first type there is no choice of termination, but for the second type the waveguide may be terminated in a short-circuiting plunger, a fixed matched load, or a movable matched load; and, of course, the 'matched load' itself may not be perfect.

There is one further possibility; there are certain applications of waveguide technique when it is necessary to match the generator, and the only practicable method of doing this is by the use of a short-circuit termination.

All these different possibilities will be considered in turn, and the simplest matching technique will be described.

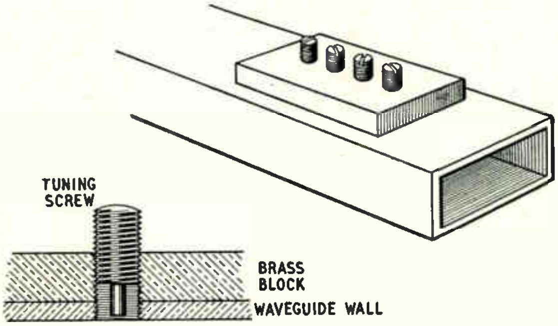

Fig. 3. *This sketch shows the form of a four-screw section and the detail of one of the screws.*

Waveguide Equipment Required

Before describing the actual matching techniques, it is perhaps advisable to describe the components required. The necessary equipment of the waveguide test bench is, in order from the generator, an attenuator, a standing-wave indicator, and a termination. The termination depends on the component to be matched and on the equipment available, and will therefore not be considered here; the standing-wave indicator is, of course, the device with which all measurements are made, and no description of it will be given. The attenuator is necessary for those cases in which the termination is a short circuit, in order to reduce 'frequency pulling' of the generator at certain positions of the plunger.

If the mismatched component is an insertion device, it is inserted between the standing-wave indicator and the termination, and the four tuning screws are either attached directly to the insertion device or attached to a separate length of waveguide coupled directly to the mismatched section. If the mismatched component is a fixed termination the screws may be attached to the termination or to a separate section of waveguide; but if the mismatched component is a movable termination, the screws must be attached to it in such a way that they move along the waveguide with the movable part of the termination. If the generator is to be matched the tuning screws must be inserted at some point between the attenuator and the standing-wave indicator, and a separate length of waveguide is almost invariably needed.

A convenient method of attaching the screws to a waveguide is sketched in Fig. 3.

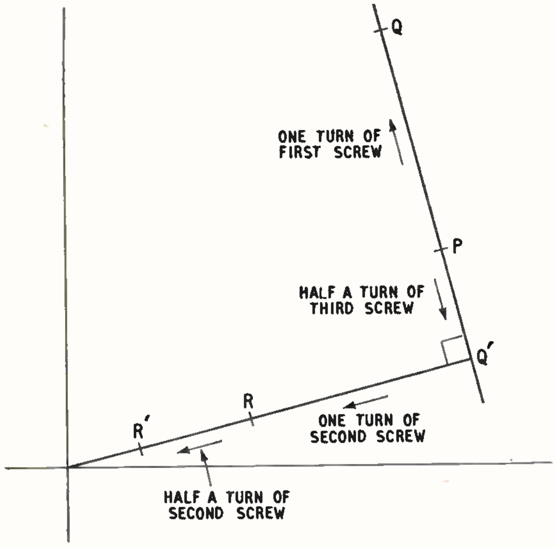

Fig. 4. *Argand diagram; QQ′ represents the locus of the reflection coefficient controlled by the first and third screws while R′Q′ represent that controlled by the second and fourth.*

Determination of Reflection Coefficients

By means of the standing-wave indicator the standing-wave ratio [(minimum voltage)/(maximum voltage)] may be found, and the position of a voltage maximum relative to the zero on the standing-wave indicator scale may be determined[1]. If the standing-wave ratio is s and the distance of a maximum from the scale zero is h, then, from (5),

$$R = (1 - s)/(1 + s),$$

and, from (7b), since h is measured towards the generator,

$$\phi = 2\beta h \pm 2n\pi. \ (n = 0, 1, 2, \ldots)$$

From R and ϕ the reflection coefficient, ρ, is found from (2), and this reflection coefficient is that *at the scale zero*. It will be found necessary to plot reflection coefficients, and this is most conveniently done in polar co-ordinates (R, ϕ), in the complex Argand plane.

Mismatched Fixed Termination

(a) Other Components Perfectly Matched

The mismatched termination may be regarded, if the mismatch is not very large, as being a combination of a perfect matched termination and an unknown shunt impedance. As the position of this impedance is unknown, there is no loss of generality if we transfer it to the zero of the standing-wave indicator scale according to equation (8b).

The reflection coefficient at the scale zero is determined as described in the preceding section, and is plotted on the Argand plane (point P, Fig. 4). One tuning screw is now inserted a known amount into the waveguide (one turn of the screw is usually convenient) and the reflection coefficient is redetermined and replotted (point Q). The line PQ, produced through Q, is the locus of the reflection coefficients which may be obtained at the scale zero by inserting the first screw only.

As can be seen from equation (10), insertion of the third screw has the opposite effect to that produced by inserting the first screw, and therefore any point on the line PQ will correspond to a particular arrangement of the first and third screws. (Points on the same side of P as Q correspond to the penetration of the first screw being greater than that of the third, and conversely.) It can now be decided, by inspection of the Argand plane diagram, what penetrations the first and third screws should have in order that the reflection coefficient at the scale zero should have minimum magnitude (point Q′: first screw withdrawn, third screw about $\frac{1}{2}$-turn inserted). For a reason that will be considered later, it is desirable to introduce the minimum possible number of screws rather than to cancel the effect of one by inserting another.

After making these adjustments, the second screw should be inserted a definite amount (again, say, one turn) and the new reflection coefficient should be plotted (point R). It is then apparent from the diagram that the second screw, if inserted about $\frac{1}{2}$-turn further, will reduce the reflection coefficient at the scale zero to a very small quantity. (Point R′.)

If the whole technique is now repeated, plotting reflection coefficients on a much larger scale, it is possible to reduce the total reflection coefficient at the scale zero to as small a value as required (neglecting the limitations of galvanometer accuracy and scale accuracy). What, in effect, has been done is to introduce a reflection coefficient due to the insertion of the tuning screws which is exactly equal and opposite to the reflection coefficient produced by the mismatch of the termination, and, provided that the screws and termination are always maintained in the same relative positions, the termination will remain perfectly matched.

This method has been described at length as it is the basic method of matching for all possible mismatched sections, even though the details differ in particular cases.

(*b*) *Other Components Imperfectly Matched*

If the measuring equipment is itself imperfectly matched, the method of matching outlined above will match the combined mismatches of the termination and the measuring equipment. After matching, both will have the same inherent standing-wave ratio, equal to the original standing-wave ratio of the measuring equipment. If the termination is movable, it is possible to match it perfectly irrespective of the accuracy of the measuring equipment, as will be shown in the following section.

Mismatched Movable Termination

(*a*) *Other Components Perfectly Matched*

There are two possible procedures for matching an imperfectly-matched movable termination: either use the method described above, remembering that it is valid for only one position of the termination, or attach the screws to the movable part of the termination so that it will be matched in all positions.

(*b*) *Other Components Imperfectly Matched*

In this case there are two separate independent shunt impedances to be considered, one corresponding to the mismatch of the termination, and one corresponding to the mismatch of the measuring equipment. We distinguish between them by noting that one impedance is stationary and that the other moves along the waveguide with the termination.

From an analysis similar to that used for combining reflection coefficients, it is possible to show that the locus of the total reflection coefficient at the scale zero as the load moves along the waveguide is a circle in the Argand plane. The *centre* of this circle corresponds to the reflection coefficient of the measuring equipment, and the *radius* of the circle corresponds to the modulus value of the reflection coefficient of the load.

By determining the total reflection coefficient for three or more positions of the load, three or more points on the circle can be found, and hence the centre and radius of the circle can be measured. As before, these reflection coefficients should be plotted on the Argand plane. The usual matching technique is employed, but the object is to produce a total reflection coefficient coinciding with the centre of the circle, and not at the origin, as before. It is not necessary to move the load during the matching process, but it is a check on the method if the position of the centre of the circle is redetermined before making the final adjustments to the screws.

Mismatched Insertion Device

(*a*) *Fixed Matched Termination*

The method of matching is the same as that used for matching a mismatched fixed termination with perfect measuring equipment, and the remarks made under that heading are also applicable here; the shunt impedance producing the mismatch is associated with the insertion device instead of the termination.

(*b*) *Fixed Mismatched Termination*

The method of matching is the same as that used for matching a mismatched fixed termination with imperfect measuring equipment, and the termination cannot be matched more accurately than the equipment. If possible, the equipment should be perfectly matched first, using a movable termination, as described in (*d*) below.

(*c*) *Movable Matched Termination*

As the position of the termination is immaterial if it is perfectly matched, the movable termination may be treated as a perfectly-matched fixed termination, using (*a*) above.

(*d*) *Movable Mismatched Termination*

The method here corresponds to that used for matching a mismatched movable termination with imperfect measuring equipment, but the object is to produce a final reflection coefficient lying on a circle centred on the origin. Apart from this, the remarks made in part (*b*) of the previous section are applicable.

(*e*) *Short-Circuit Termination*

The assumption that the shunt impedances involved are all large is no longer valid if the termination used is a short-circulating plunger, and the mathematical basis for the matching process must be derived.

Suppose that the mismatch of the insertion device is produced by a shunt impedance Z_0/m at the zero of the standing-wave indicator scale, where m is a complex number small compared with unity. Suppose also that the distance of the short-circuiting plunger from the scale zero is d. If we put $Z_2 = 0$, $y = d$ in equation (8b) we obtain the impedance of the short-circuit transformed to the scale zero as $jZ_0.\tan(\beta d)$. In parallel with Z_0/m, this gives:

$$Z_T = jZ_0t/(1 + jmt)$$

where $t = \tan(\beta d)$. The corresponding reflection coefficient is

$$\rho = -\frac{1 - jt + jmt}{1 + jt + jmt}$$

and, neglecting all terms in m other than those of the first degree, we can simplify this to

$$\rho = -e^{-j2\beta d}[1 - 2m.\sin^2(\beta d)]$$

when the magnitude of ρ is given by

$$R = |1 - 2m.\sin^2(\beta d)|$$

If the generator voltage remains constant (i.e., if the effect of frequency pulling is negligible) the magnitude of the measured voltage at a voltage minimum is proportional to $(1 - R)$. The technique is, therefore, to measure the minimum voltage for various positions of the plunger, when the *maximum* minimum voltage has a value proportional to $(1 - |1 - 2m|)$. The position of the plunger to give this value is noted, and the reflection coefficient is measured and plotted for this position of the plunger. The plunger is then moved through a quarter of a guide wavelength and the *negative* of the reflection coefficient is plotted (to allow for the change of sign of $e^{-j2\beta d}$ in the expression for ρ).

The plunger is then returned to the position where the measured minimum voltage has a maximum value, and the tuning screws are adjusted as before to produce a final reflection coefficient coincident with the second point plotted. It is worth while pointing out that this method of matching is that most liable to errors, as the calculations involve quantities differing from unity by very small amounts. The purpose of plotting the first point on the Argand diagram is to give an indication of the magnitude of the reflection coefficient so that more or less care may be exercised in the measurements.

Mismatched Generator

For the purpose of matching a mismatched generator, some stipulations regarding the waveguide equipment must be made. First, the only practicable method of matching requires a short-circuit termination; secondly, the matching screws must be on the same side of the standing-wave indicator as the generator; thirdly, there must be sufficient attenuation between the generator and the standing-wave indicator to minimize the effect of frequency pulling, and, unless the attenuator is very well designed, it must be left at the same setting to maintain the matching; lastly, the method of measuring the standing-wave ratio must be modified.

In order not to complicate matters, it will be assumed that the measuring equipment has no inherent errors except for that producing the mismatch of the generator; this mismatch will be assumed to be due to a shunt impedance Z_0/m at some point between the generator and the standing-wave indicator. Let the distance between the shunt impedance and the short-circuit be d, as before. The reflection coefficient at the position of the shunt impedance, from the preceding analysis, is given by[2]

$$\rho = -e^{-j2\beta d}[1 - 2m.\sin^2(\beta d)].$$

It can be shown (using certain equations not given in this paper) that the measured voltage at a point *an odd number of quarter-wavelengths from the plunger* is proportional to

$$|1 + jm.\sin(\beta d).e^{-j\beta d}|$$

and this expression is exact. If, then, the probe of the standing-wave indicator is always an odd number of quarter-wavelengths from the short-circuit, the graph of measured voltage against plunger (or probe) position can be plotted; the variation will be approximately sinusoidal, and the standing-wave ratio can be determined. From the standing-wave ratio and the position of the maximum the reflection coefficient can be found, and plotted on the Argand plane. The usual matching technique can then be employed to reduce the reflection coefficient to zero.

Phase-Shifting by Tuning Screws

Although this is not a matching process, it is included here as being relevant to the use of tuning screws. There are occasions when it is required to delay or advance the phase of the voltage wave in a length of waveguide, without introducing any appreciable mismatch, and two tuning screws can be used for this purpose. In order that the screws may have no effect on the matching of the guide it is necessary that they have exactly equal capacitance, and it is also necessary that the distance between them is given by

$$d = \lambda_g.\tan^{-1}(2/k)/2\pi$$

where k is the ratio of the characteristic impedance of the waveguide to the reactance of each screw, and can be calculated from the reflection coefficient of one screw by an analysis similar to that given under the theory of four-screw tuning.

If these two conditions are fulfilled, then the total phase *advance* produced by the screws will be

$$\phi = 2.\tan^{-1}(k/2)$$

$$\text{or } \phi = \pi - 4\pi d/\lambda_g.$$

If the phase-shift required is small, when k is very small, d is very nearly a quarter of a wavelength.

Modifications for Larger Discontinuities

Nearly all the matching techniques described above have had as a condition for their validity that the mismatch involved is small. This assumption cannot be taken as true if the mismatch is large, and we will now consider the situation when first-order approximations are inaccurate.

In general, any mismatch or combination of mismatches may be regarded as a T- or π-section network inserted in the waveguide, and this network will effectively shorten (or lengthen) the waveguide. As the matching process continues, and the total mismatch becomes smaller, the approximate formulae become valid; hence the redetermination of the reflection coefficient or its locus after some alterations have been made. It may be assumed, for practical purposes, that the approximate formulae may be used when the standing-wave ratio is 0·95 or greater with a 'matched' termination, and when the standing-wave ratio is 0·1 or less with a short-circuit termination.

When using two screws for phase-shifting, the two screws introduce equal and opposite impedances, which cancel, but they introduce equal phase shifts which *add*; if, then, a large mismatch is cancelled by four tuning screws, the phase-shift of the screws will be equal to that of the mismatch, and the effective length of the waveguide may be altered by an appreciable amount. Matching processes which involve accurate measurements of distance may therefore be slightly inaccurate, but, again, this inaccuracy will become unimportant when the waveguide is almost completely matched. As an idea of the orders of magnitude involved, a mismatch producing a standing-wave ratio of about 0·875 will, combined with the tuning screws, produce a phase-shift corresponding to about 1/50 of a guide wavelength when matched.

Inherent Mismatches in Equipment

Of the many possible types of discontinuity which may be encountered in waveguide work, we shall consider only two: the mismatch produced by the probe of the standing-wave indicator, and the mismatches introduced by the waveguide couplings. One convenient method of determining whether or not the probe of the standing-wave indicator is introducing an appreciable mismatch is to measure the standing-wave ratio in a length of waveguide which is *almost* perfectly matched with two different probe penetrations; if there is no difference between the results, the probe has no measurable effect.[3]

Mismatches introduced by waveguide couplings may be of any magnitude whatever, depending on the size and type of waveguide, the type of coupling, and the finishing, or machining, of the couplings. It would appear that the most serious inaccuracies occur when the dimensional tolerances of the two pieces of waveguide meeting at the couplings are of opposite kinds; in 1-in. × $\frac{1}{2}$-in. rectangular waveguide the difference between the internal dimensions may be as much as 1/80 in., and, in particularly bad cases, may be even 1/50 in. (i.e., 2%). In general, if no 'step' can be seen at the junction of two lengths of waveguide, the mismatch at the coupling will not be appreciable.

Other Practical Considerations

It has been pointed out that the tuning screws may be spaced at distances of any odd number of eighth-wavelengths; if possible, the shortest convenient distance should be used. It is sometimes necessary to make measurements over a relatively large bandwidth (usually about 5% of the mean frequency), and the distance between adjacent screws will be the designed amount at one frequency only. The difference (in wavelengths) between adjacent screws will increase as the frequency increases and, for a given increase in frequency, will increase as the actual distance between the screws increases. For this reason the actual distance between the screws should be as small as possible and, for the same reason, the screws should be placed as close as possible to the mismatch which they are designed to correct. If the waveguide is to be used over a large bandwidth, the matching process should be carried out at the centre of the frequency band.

Another important point is that, if a crystal rectifier or other non-linear detector is matched for a certain crystal current in a certain load resistance, then, in general, the crystal will not be matched if the current or load resistance is changed. The effect on the matching of alterations in crystal current is very marked. Similarly, if a variable attenuator is included in a matched section of waveguide the waveguide will be matched for one particular setting of the attenuator unless the attenuator has been very well designed.

If the waveguide system contains two different sizes of waveguide, as will be the case if a taper or transformer is used, it must be remembered that the guide wavelength and the phase constant will be different in the different sections of waveguide.

Acknowledgments

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MICROWAVE WIDE-ANGLE SCANNER

By J. Brown, M.A., A.M.I.E.E.

SUMMARY.—The theory of lenses in which the refractive index is a continuously-varying function of position is studied and results are derived for a lens which permits a microwave beam to be scanned through 360°.

Introduction

WIDE-ANGLE scanners using media in which the refractive index is a function of position have recently been used at microwave frequencies. Most of these have been based on a theoretical result derived by R. K. Luneberg[1] relating to a lens system consisting of a sphere of radius a, for which the refractive index, $n(r)$, at any point distant r from the centre O is given by

$$n(r) = [2 - (r/a)^2]^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad (1)$$

Luneberg has shown that if a point radiator is placed at any point A on the surface of such a sphere then a beam will be radiated in the direction AO. This result is derived from the theory of geometrical optics, and so only holds exactly when the wavelength of the radiation (λ) is vanishingly small compared with the diameter of the sphere. The radiated beam is then a parallel-sided one whose cross-section is a circle having the same diameter as the sphere. When the ratio of the wavelength to the diameter becomes finite, diffraction of the beam occurs and it will have a half-power beam-width of the order $\lambda/2a$ radians. Since the lens is symmetrical about its centre, a movement of the point radiator on the surface of the sphere changes only the direction of the radiated beam and leaves its shape unaltered. If the point radiator of feed is free to move anywhere over the surface of the sphere then the beam may be radiated in any required direction. A complete coverage may thus be obtained by moving only the feed.

Practical applications of the principle described above have so far been limited to equivalent two-dimensional lenses.[2,3,4] A section through the centre of the sphere is constructed so that it has the same variation of refractive index as in equation (1). The scanning performance is then restricted to moving the feed around the circumference of this section and the maximum of the radiated beam lies in the plane of the section. In this plane the beam has a half-power beam-width of about $\lambda/2a$ radians and in the plane normal to this the beam-width is governed by the thickness of the lens. Lenses of this type have performed satisfactorily, but a disadvantage is that the feed has to move round the circumference. Since this may be as much as 6 ft the mechanical problems involved in moving the feed are considerable. These problems are greatly simplified if a lens can be designed for which the feed moves around the circumference of a circle whose diameter is less than that of the complete lens. The design of such a lens is considered in this paper.

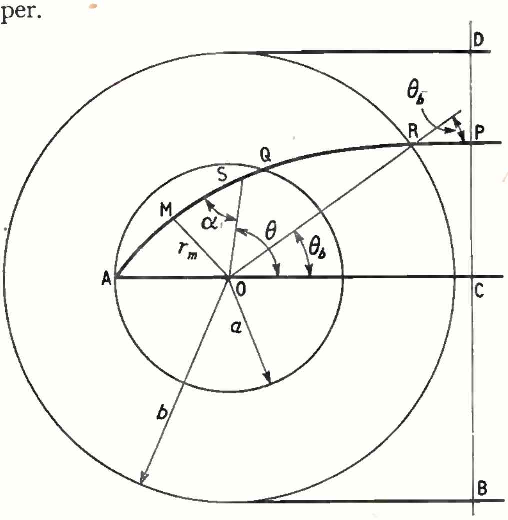

Fig. 1. *Ray paths in wide-angle scanner.* OA $= a$, OC $= b$, OS $= r$, OM $= r_m$, CP $= x$.

General Theory

The theory is most simply presented if only a two-dimensional system analogous to that discussed in the previous paragraph is considered. Such a system is outlined in Fig. 1. It is obvious that if an all-round scan is to be possible the lens must be symmetrical about its centre O; i.e., the refractive index within the lens must be a function of only the radial distance r. The general properties of such a system have been studied by D. Gabor in unpublished work and he has shown that the refractive index may be specified more or less arbitrarily for values of r lying between a and b. The refractive index in the inner region (i.e., $0 \leqslant r \leqslant a$) may then be calculated to ensure that radiation from a point such as A is collimated to give a beam in the direction AOC. This result is derived below.

The radiated beam has its maximum in the direction AOC provided that in the free-space region outside the lens the phase of the radiation is constant on lines such as BCD at right-angles to AOC. This in turn requires that the optical

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path length, (i.e., $\int n\,ds$) should be constant for any ray path from A to a point P on the line CD. Consider a typical ray path AQRP: then

$$\int_A^R n\,ds + \mathrm{RP} = L \qquad \ldots \quad \ldots \quad \ldots \qquad (2)$$

where L is independent of the ray path selected. By Fermat's theorem, the optical path length between two points is stationary for small variations in the path and this applies to the integral in equation (2). Let r, θ be the polar co-ordinates of a point on the ray path. Then

$$\int n\,ds = \int n(r)\,[1 + r^2\theta'^2]^{\frac{1}{2}} dr \qquad \ldots \quad \ldots \qquad (3)$$

where $\theta' = d\theta/dr$ and the restriction that the refractive index depends only on r is indicated explicitly. The condition that the integral is stationary for small variations in the path is found from the Calculus of Variations with the result

$$r\theta' = C/[r^2n^2(r) - C^2]^{\frac{1}{2}} \qquad \ldots \quad \ldots \qquad (4)$$

where C is a parameter whose value depends on the ray path selected. Equation (4) is a differential equation, which when integrated, gives the equation of the ray path. It may be converted to the equivalent relation

$$r\,n(r)\sin\alpha = C \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (5)$$

by using the geometrical result

$$r\theta' = \tan\alpha \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (6)$$

where α is the angle between the tangent to the ray path and the radius vector, as in Fig. 1. Equation (5) must be satisfied for the whole ray path, whether within the lens or in the free-space region outside. It may be noted that at a discontinuity in the refractive index, equation (5) is equivalent to Snell's Law.

So far the lens has been regarded as a radiator. It is more convenient in the rest of the analysis to consider it as a receiving aerial: so that any rays in free-space which are parallel to AOC are brought to a focus at A. This means that any ray path through the lens must pass through the point (a, π) if a parallel beam is incident on the lens. This condition will be used to calculate the values of refractive index in the inner portion of the lens.

Integration of equation (4) gives for the ray path within the lens

$$\theta_b - \theta = \int_r^b \frac{C\,dr}{r[r^2n^2(r) - C^2]^{\frac{1}{2}}} \qquad \ldots \quad \ldots \qquad (7)$$

where θ_b is the value of θ when r is equal to b. From equation (5),

$$C = b\sin\theta_b \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (8)$$

since θ_b is also the angle between the radius vector and the tangent to the ray path in free-space at the point R; i.e., when $r = b$. The refractive index of free-space has of course the value unity.

Before the condition that $\theta = \pi$ when $r = a$ can be inserted in equation (7) it is necessary to consider the general shape of the ray path. It has been shown by Gabor that a solution can only be obtained if the ray penetrates the inner region as shown in Fig. 1 for a typical ray path. The minimum distance of this ray path from the origin, r_m, is obtained from the condition that $\alpha = \pi/2$ at M and so

$$r_m n(r_m) = C \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (9)$$

To avoid confusion in the analysis it is convenient to denote the refractive index in the region $a \leqslant r \leqslant b$ by $n_1(r)$. Applying the condition that $\theta = \pi$ when $r = a$, now gives

$$\begin{aligned}\theta_b - \pi = &-\underset{\mathrm{R}\rightarrow\mathrm{Q}}{\int_a^b \frac{C\,dr}{r[r^2n^2{}_1(r) - C^2]^{\frac{1}{2}}}}\\ &-\underset{\mathrm{Q}\rightarrow\mathrm{M}}{\int_{r_m}^a \frac{C\,dr}{r[r^2n^2(r) - C^2]^{\frac{1}{2}}}}\\ &-\underset{\mathrm{M}\rightarrow\mathrm{A}}{\int_{r_m}^a \frac{C\,dr}{r[r^2n^2(r) - C^2]^{\frac{1}{2}}}} \qquad \ldots \qquad (10)\end{aligned}$$

where the appropriate sign to attach to each integral has been selected by reference to Fig. 1. The origin of each integral is indicated underneath it. Equation (10) may be rewritten

$$\begin{aligned}2\int_{r_m}^a \frac{C\,dr}{r[r^2n^2(r) - C^2)^{\frac{1}{2}}} = \pi - \sin^{-1}\left(\frac{C}{b}\right)\\ - \int_a^b \frac{C\,dr}{r[r^2n^2{}_1(r) - C^2]^{\frac{1}{2}}} \qquad (11)\end{aligned}$$

by using equation (8). This equation has to be satisfied for all values of C lying between 0 and 1 and is therefore an integral equation for the, as yet, unknown function $n(r)$. The solution is obtained by making a number of changes of variable to reduce the equation to a standard type.

$$\text{Let} \quad \xi = rn(r) \qquad \ldots \quad \ldots \quad \ldots \qquad (12)$$

$$\rho = \log_e r \qquad \ldots \quad \ldots \quad \ldots \qquad (13)$$

$$\xi_m = r_m n(r_m) = C \qquad \ldots \quad \ldots \qquad (14)$$

$$\text{and} \quad \xi_a = an(a) \qquad \ldots \quad \ldots \quad \ldots \qquad (15)$$

Equation (11) may now be written

$$\xi_m \int_{\xi_m}^{\xi_a} \frac{(d\rho/d\xi)d\xi}{(\xi^2 - \xi_m{}^2)^{\frac{1}{2}}} = f(\xi_m) \qquad \ldots \quad \ldots \qquad (16)$$

$$\begin{aligned}\text{where} \quad f(\xi_m) = \frac{1}{2}\Bigg[\pi - \sin^{-1}\frac{\xi_m}{b}\\ - \int_a^b \frac{\xi_m\,dr}{r\{r^2n^2{}_1(r) - \xi_m{}^2\}^{\frac{1}{2}}}\Bigg] \qquad \ldots \quad \ldots \qquad (17)\end{aligned}$$

If $d\rho/d\xi$ is regarded as the unknown function, then equation (16) is of Abel's type[5] and a solution, if one exists, is obtained formally by multiplying both sides of equation (16) by $1/(\xi_m{}^2 - z^2)^{\frac{1}{2}}$, where z is less than ξ_m, and integrating with respect to ξ_m from z to ξ_a. This gives

$$\int_z^{\xi_a} \int_{\xi_m}^{\xi_a} \frac{\xi_m(d\rho/d\xi)d\xi d\xi_m}{[(\xi^2 - \xi_m{}^2)(\xi_m{}^2 - z^2)]^{\frac{1}{2}}} = \int_z^{\xi_a} \frac{f(\xi_m)d\xi_m}{(\xi_m{}^2 - z^2)^{\frac{1}{2}}} \quad \ldots \quad \ldots \quad (18)$$

The integral on the right may be simplified by reversing the order of integration and using the result

$$\int_z^{\xi} \frac{\xi_m d\xi_m}{[(\xi^2 - \xi_m{}^2)(\xi_m{}^2 - z^2)]^{\frac{1}{2}}} = \pi \quad \ldots \quad (19)$$

whence

$$\pi \int_z^{\xi_a} \frac{d\rho}{d\xi} d\xi = \int_z^{\xi_a} \frac{f(\xi_m)d\xi_m}{(\xi_m{}^2 - z^2)^{\frac{1}{2}}} \quad \ldots \quad \ldots \quad (20)$$

i.e.,

$$\pi\,[\rho(\xi_a) - \rho(z)] = \int_z^{\xi_a} \frac{f(\xi_m)d\xi_m}{(\xi_m{}^2 - z^2)^{\frac{1}{2}}} \quad \ldots \quad (21)$$

where $\rho(\xi_a)$ is the value of ρ when $\xi = \xi_a$; i.e., from equations (13) and (15) $\rho(\xi_a) = \log_e a$. Similarly, if $z = rn(r)$ then $\rho(z) = \log_e r$. Equation (21) may therefore be written

$$\log_e a/r = \frac{1}{\pi} \int_{rn(r)}^{an(a)} \frac{f(x)\,dx}{\{x^2 - r^2n^2(r)\}^{\frac{1}{2}}} \quad \ldots \quad (22)$$

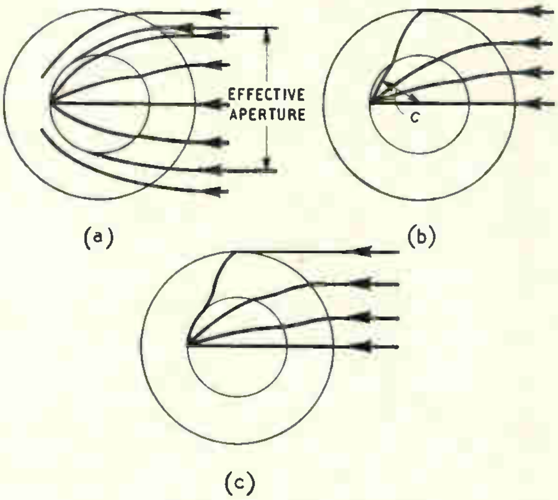
in which the variable under the integration sign has been changed to x. This equation provides the formal solution to the problem and shows, as stated earlier, that the refractive index may be specified in the region $a \leqslant r \leqslant b$. A solution will, however, only exist provided that $f(x)$, and hence the function $n_1(r)$, satisfy certain conditions which will be considered in a later paragraph. When $f(x)$, as given by equation (17), is substituted in equation (21) and some of the integrations carried out, the following result is obtained:

$$\log_e\left(\frac{a}{r}\right) = \log_e\left[\frac{an(a)}{rn(r)} + \left\{\frac{a^2n^2(a)}{r^2n^2(r)} - 1\right\}^{\frac{1}{2}}\right] - \frac{1}{\pi}\int_{an(a)}^{rn(r)} \frac{\sin^{-1}(x/b)dx}{[x^2 - r^2n^2(r)]^{\frac{1}{2}}} - \frac{1}{\pi}\int_a^b \tan^{-1}\left[\frac{a^2n^2(a) - r^2n^2(r)}{y^2n_1{}^2(y) - a^2n^2(a)}\right]\frac{dy}{y} \quad \ldots \quad \ldots \quad (23)$$

In this expression r is only used to denote radial distance in the inner region of the lens. To avoid confusion, the radial distance in the outer region is denoted by y, so that $n_1(y)$ is the known refractive index at a point, whose distance, y, from the centre lies between a and b. Before particular solutions resulting from the specification of $n_1(y)$ are considered, some general properties of the lens will be investigated.

General Properties of the System

A number of important properties of the lens may be deduced from the nature of the ray paths, and from a study of these it is possible to decide on suitable forms for the function $n_1(y)$. The first point is whether or not it is possible for the effective aperture of the lens to equal its diameter; i.e., whether the radiation from the lens appears to come from an aperture of width $2b$. Once again this is most easily studied when the lens is acting as a receiver. A comparison of the diagrams in Fig. 2 shows that there are three possibilities. In Fig. 2(a), certain of the ray paths never intersect the circle of radius a and the effective aperture of the lens is then less than $2b$. In Fig. 2(b), all the ray paths intersect the circle and the effective aperture will be $2b$. The intermediate case is shown in Fig. 2(c) and the outermost ray is just tangential to the circle of radius a. Since the effective aperture must be as large as possible, the first case is of no interest. It is found that a solution can be obtained for the second case only if the refractive index is also specified in the region $c \leqslant r \leqslant a$, where c is the minimum distance of the outermost ray path from the centre of the lens. The analysis then becomes more complicated than for the intermediate case shown in Fig. 2(c) and since the latter also gives the maximum effective aperture it is the only type to be examined here.



(a) (b)

(c)

Fig. 2. *Possible types of lens.*

The condition that the lens should be of the third type is a very simple one. The outermost ray path is the one for which the constant C has the value b (see equation (8)) and is tangential to the circle $r = a$ if the angle α equals $\pi/2$ for this value of r. From equation (5) this requires,

$$an(a) = C = b \qquad \ldots \quad \ldots \quad \ldots \qquad (24)$$

The lens is thus of the third type if $n(a) = b/a$. One result of trying to reduce the feed circle is apparent: the greater the reduction the larger are the values of refractive index which are required.

It has so far been tacitly assumed that the refractive index is continuous at the junction between the inner and outer regions. The condition for this to be so may be obtained from the solution given in equation (23), and it may be seen that $n(r)$ is continuous at $r = a$ if the last integral in this equation vanishes when $r = a$. At first sight it would seem that this is always true because of the factor $[a^2n^2(a) - r^2n^2(r)]^{\frac{1}{2}}$ in the numerator of the $\tan^{-1}$ term. It must be remembered however that integrals for optical path lengths must be taken in the Lesbesque sense and so if the denominator $[y^2n_1{}^2(y) - a^2n^2(a)]^{\frac{1}{2}}$, of the $\tan^{-1}$ term vanishes for a set of points of zero measure in the region $a \leqslant y \leqslant b$, the integral does not vanish. It may be concluded that the refractive index is continuous at $r = a$ provided $yn_1(y)$ does not equal $an(a)$ except possibly at a set of points of zero measure. Further, $yn_1(y)$ cannot be less than $an(a)$ for then the integral would become complex. The function $n_1(y)$ must, therefore, satisfy the condition

$$yn_1(y) \leqslant an(a) \qquad \ldots \quad \ldots \quad \ldots \qquad (25)$$

the equality sign being permissible only for a set of points of zero measure.

If the condition that $an(a) = b$ is substituted in the general solution (23), the first integral may be evaluated for

$$\frac{1}{\pi}\int_{rn(r)}^{b} \frac{\sin^{-1}(x/b)dx}{[x^2 - r^2n^2(r)]^{\frac{1}{2}}}$$

$$= \frac{1}{2}\log_e\left[1 + \left\{1 - \frac{r^2n^2(r)}{b^2}\right\}^{\frac{1}{2}}\right] \qquad \ldots \qquad (26)$$

as has been shown by Luneberg.[1] The general solution may now be written more concisely as

$$\log_e \frac{an(r)}{b} = \tfrac{1}{2}\log_e(1 + P) - I(P) \qquad \ldots \qquad (27)$$

where $$P = \left[1 - \frac{r^2n^2(r)}{b^2}\right]^{\frac{1}{2}} \qquad \ldots \quad \ldots \qquad (28)$$

and $$I(P) = \frac{1}{\pi}\int_a^b \tan^{-1}\left[P\Big/\left\{\frac{y^2n_1{}^2(y)}{b^2} - 1\right\}^{\frac{1}{2}}\right]\frac{dy}{y}$$

$$\ldots \quad \ldots \quad \ldots \quad \ldots \qquad (29)$$

The method used to obtain the solution is only valid if the function $\xi = rn(r)$ is a monotonic increasing one in the range $0 \leqslant r \leqslant a$. It follows that P must decrease monotonically in the same range and the function $n_1(y)$ must be such that the solution satisfies this condition. The derivative dP/dr must therefore be negative in the above range for r. Differentiation of equation (27) gives

$$-\frac{2}{r} - \frac{1}{1 - P}\frac{dP}{dr} = -2I'(P)\frac{dP}{dr} \qquad \ldots \qquad (30)$$

where $I'(P) = dI/dP$ and $n^2(r)$ has been replaced by $b^2(1 - P^2)/r^2$. Hence

$$\frac{dP}{dr} = -\frac{2}{r}\,\frac{1}{\dfrac{1}{1 - P} - 2I'(P)} \qquad \ldots \quad \ldots \qquad (31)$$

Since dP/dr must be negative

$$\frac{1}{1 - P} > 2I'(P) \qquad \begin{array}{l}\text{for } 0 \leqslant r \leqslant a \\ \text{i.e., for } 0 \leqslant P \leqslant 1\end{array} \qquad \ldots \qquad (32)$$

The function $n_1(r)$ must therefore be selected so that this condition is satisfied.

Investigation of Particular Cases

(i) *Refractive Index Constant in Outer Annulus*

A particularly simple choice for the function $n_1(y)$ is that it should be constant for $a \leqslant y \leqslant b$. The ray paths in the annulus are then straight lines and condition (24) requires that the path through the edge of the aperture is tangential to the feed circle. From this condition it follows that the value n_1 of the refractive index in the annulus is b/a. Substitution of this in equation (27) gives

$$\log_e \frac{an(r)}{b} = \frac{1}{2}\log_e(1 + P)$$

$$-\frac{1}{\pi}\int_a^b \tan^{-1}\left\{\frac{P}{(y^2/a^2 - 1)^{\frac{1}{2}}}\right\}\frac{dy}{y} \qquad (33)$$

and it may be shown that this solution satisfies condition (32). The integral may be evaluated numerically and a relation between $n(r)$ and $rn(r)$ obtained. From this $n(r)$ may be calculated as a function of r. If b is equal to $2a$, $n(r)$ has a maximum value of 2·34 at the centre of the lens, and drops monotonically to the value 2 when $r = a$. The refractive index variation is one which could be achieved in practice with the aid of artificial dielectrics such as have been described by Kock.[6] An undesirable feature of the design is the comparatively large value of refractive index at the lens circumference. The abrupt change in refractive index at the circumference would cause excessive reflection and lead to difficulties in matching the feed to the lens. A method of avoiding this reflection is described in the next paragraph.

(ii) *Variable Refractive Index in Annulus*

The reflection mentioned at the end of the last paragraph may be avoided if the refractive index at the lens circumference is unity; i.e., if $n_1(b) =$ unity. This suggests that the function $yn_1(y)$ should be made equal to b throughout the range $a \leqslant y \leqslant b$ for then $n_1(a) = b/a$ and $n_1(b) =$ unity. This is not a possible choice for $n_1(y)$, however, since it does not satisfy the condition given in (25). There is, of course, a very wide choice for the function $n_1(y)$ even subject to the required conditions, but as yet only one has been investigated. This is

$$y^2 n_1^2(y) = b^2[1 + \gamma(b - y)(y - a)] \quad .. \quad (34)$$

which satisfies all the required conditions except (32), if the constant γ is positive. The refractive index in the inner region is calculated by inserting this value of $n_1(y)$ in the general solution (27). The evaluation of the integrals is carried out in the Appendix with the result:

$$\frac{4a}{b} n^2(r) = \frac{(1 + P)[\sqrt{\gamma}(a + b) + \{4P^2 + \gamma(b - a)^2\}^{\frac{1}{2}}]^2}{(P + \sqrt{\gamma ab})^2} \quad .. \quad .. \quad (35)$$

The constant γ must be selected so that condition (32) is satisfied and it is shown in the Appendix that γ must therefore exceed $4/ab$. Any value greater than $4/ab$ gives a solution and it remains to make a suitable choice. It may be seen from equation (35) that as γ increases the maximum value of refractive index required in the lens increases. Since the construction of the lens is simplified by keeping the refractive index as low as possible, it is advantageous to let γ have its minimum value $4/ab$. Numerical values of the refractive index have been calculated for a lens in which b is twice a. The results are plotted in Fig. 3.

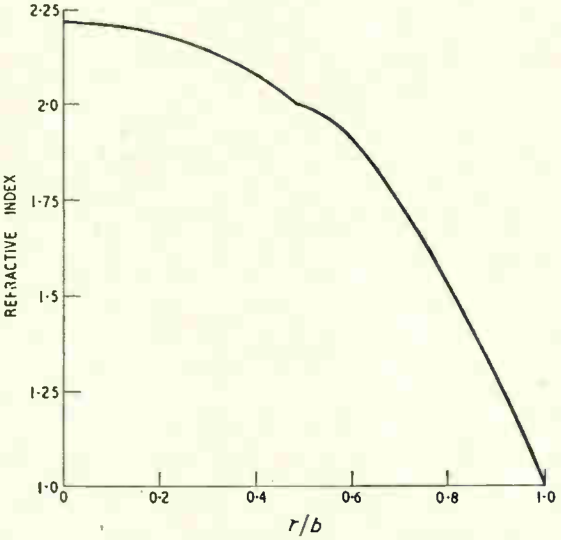


Fig. 3. *Refractive index for scanner in which $b = 2a$.*

The Shape of the Radiation Pattern

The shape of the radiated beam may be calculated from the aperture field distribution using the Fourier transform relation. For such a calculation the aperture is most conveniently taken as the straight line BD in Fig. 1. The field distribution on this line may be calculated from the radiation pattern of the feed and the shape of the ray paths in the lens. Let α_0 be the value of the angle α for a ray path at the feed position A, and let $P(\alpha_0)d\alpha_0$ be the power radiated from the feed in the sector bounded by the angles α_0 and $\alpha_0 + d\alpha_0$. The angle θ_b for the ray path which leaves the feed at the angle α_0 is given from equations (5) and (8) by

$$an(a) \sin \alpha_0 = b \sin \theta_b \quad .. \quad .. \quad (36)$$

and since $an(a) = b$ by equation (24), α_0 must equal θ_b. If the position of a point P in the aperture is specified by the distance x from C, then

$$x = b \sin \theta_b = b \sin \alpha_0 \quad .. \quad .. \quad (37)$$

Let $A(x)$ be the amplitude distribution in the aperture so that the power radiated between x and $x + dx$ is $kA^2(x)dx$, where k is a proportionality constant. If x is related to α_0 by equation (37) and

$$dx = b \cos \alpha_0 \, d\alpha_0 \quad .. \quad .. \quad .. \quad (38)$$

then $\quad kA^2(x)dx = P(\alpha_0)d\alpha_0$

so that

$$A(x) = \left[\frac{P(\alpha_0)}{k}\frac{d\alpha_0}{dx}\right]^{\frac{1}{2}} = \left[\frac{P(\alpha_0)}{kb \cos \alpha_0}\right]^{\frac{1}{2}} \quad (39)$$

The amplitude distribution may therefore be calculated by inserting the power radiation pattern of the feed in equation (39) and using equation (37) to relate α_0 and x. A suitable feed for a lens such as the one discussed in this paper is a small horn and since for such a feed the power radiated falls off more rapidly than $\cos \alpha_0$, the aperture distribution will be sufficiently tapered to give a reasonable radiation pattern.

Acknowledgment

The author is indebted to Dr. D. Gabor of the Imperial College of Science and Technology for making available the results of unpublished work, and for helpful suggestions during the preparation of this paper.

APPENDIX

Evaluation of $I(P)$ for the refractive index given by equation (34).

From equation (34) and the definition of $I(P)$,

$$I(P) = \frac{1}{\pi}\int_a^b \tan^{-1}\left\{\frac{P}{[\gamma(b - y)(y - a)]^{\frac{1}{2}}}\right\}\frac{dy}{y}$$

and so $$I'(P) = \frac{1}{\pi}\int_a^b \frac{[\gamma(b - y)(y - a)]^{\frac{1}{2}}dy}{y[P^2 + \gamma(b - y)(y - a)]}$$

Let $\quad y = (a + bt^2)/(1 + t^2)$

Then $$I'(P) = \frac{2\sqrt{\gamma}}{\pi}\int_0^\infty \frac{(b - a)^2 t^2 \, dt}{[P^2(1 + t^2)^2 + \gamma(b - a)^2 t^2](a + bt^2)}$$

The integration may be carried out by factorizing the denominator with the result

$$I'(P) = \sqrt{\gamma}\,\frac{\left[\sqrt{ab}\,\{4P^2 + \gamma\,(b - a)^2\}^{\frac{1}{2}} - (a + b)P\right]}{(\gamma ab - P^2)\;\{4P^2 + \gamma\,(b - a)^2\}^{\frac{1}{2}}}$$

This may be integrated with respect to P by using the substitution

$$4P^2 + \gamma(b - a)^2 = x^2$$

and it is found that

$$I(P) = \frac{1}{2}\log_e \frac{4b}{a}\cdot\frac{(\sqrt{\gamma ab} + P)^2}{\left[\sqrt{\gamma}\,(a + b) + \{4P^2 + \gamma\,(b - a)^2\}^{\frac{1}{2}}\right]^2}$$

The solution must satisfy condition (32) and so

$$\frac{1}{1 - P} \geqslant \frac{2\sqrt{\gamma}\left[\sqrt{ab}\,\{4P^2 + \gamma\,(b - a)^2\}^{\frac{1}{2}} - (a + b)P\right]}{(\gamma ab - P^2)\;\{4P^2 + \gamma(b - a)^2\}^{\frac{1}{2}}}$$

for $0 \leqslant P \leqslant 1$

Consider first the condition when P is zero. It becomes

$1 \geqslant 2/\sqrt{\gamma ab}$; i.e., $\gamma ab \geqslant 4$.

Since γab is greater than unity, it is always greater than P^2 and so $\{4P^2 + \gamma(b - a)^2\}^{\frac{1}{2}} < \sqrt{\gamma}(a + b)$

The value of $2\,I'(P)$ is therefore less than

$$\frac{2\sqrt{\gamma ab}}{\gamma ab - P^2} - \frac{2P}{\gamma ab - P^2} = \frac{2}{\sqrt{\gamma ab} + P}$$

which in turn is always less than or equal to $1/(1 - P)$ in the range $0 \leqslant P \leqslant 1$. If γ exceeds $4/ab$ condition (32) is therefore satisfied throughout the required range of P.

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20TH NATIONAL RADIO EXHIBITION

A Predominantly Television Show

HELD at Earls Court from 2nd to 12th September, this year's exhibition was primarily one of broadcast receiving apparatus, among which television was predominant. This is a natural consequence of the fact that the exhibition is one for the non-technical public. In fact, it did cover more than broadcasting, for sound recording, commercial apparatus and 'electronic equipment' were all exhibited, while the Services had a big display of radar and communications apparatus.

It was, however, only in broadcasting that the exhibition could be considered in any sense complete and representative of the industry. However great the interest and importance of individual exhibits in other sections, they remained individual exhibits. Only in broadcasting could the exhibition be regarded as an opportunity for assessing the present state of the art.

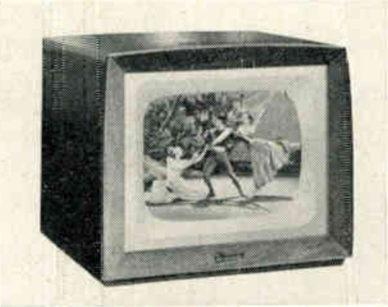
Within this field, television is of major technical interest, for designers have not yet reached the same measure of general agreement on the best solutions to the problems of design and production as they have in the case of sound broadcasting. It has been apparent for some years that the technical development of sound broadcast receivers has reached, temporarily at any rate, a large measure of stability. No major changes in design have occurred. New models use well-tried forms of circuit and, apart from the exterior styling, differ from earlier ones mainly in giving an equivalent performance in a somewhat more economical way, or a somewhat better performance for the same cost, or greater reliability. This last is a matter which it is impossible to judge by circuitry or appearance, of course, for reliability cannot be seen, only experienced. Intensive development directed towards the improvement of reliability is taking place in other branches of radio and, although it would be uneconomic and unnecessary to apply many of its results to the relatively mild conditions of broadcasting, so much has been learned about the factors which influence reliability that appreciable improvement can be expected. Improvements do not always demand more expensive components and materials; they can result from knowing what not to do as much as from knowing what to do.

It was evident last year that television had not reached the same stability of design as sound broadcast receivers, nor has it yet. It has, however, progressed some way towards it, for in broad outline the circuits of the majority of television receivers are very similar. It is in their details that they differ from one another, and this they do very much. The superheterodyne is practically universal; the straight r.f. amplifier, which was so popular when British television had only one transmitter, has been killed by the difficulty of making it adjustable for reception on the different frequencies now in use.

Bush TV32 with 14-in. rectangular tube and moulded bakelite case.

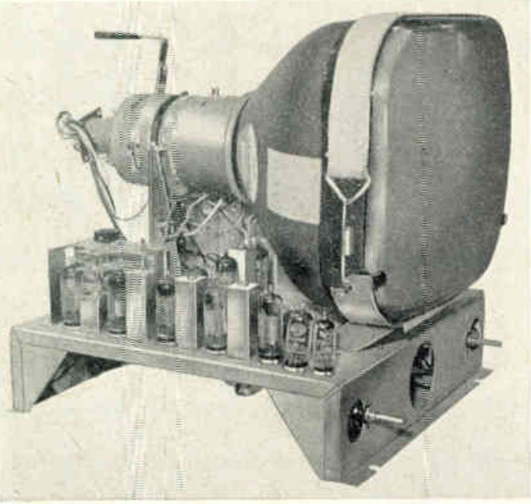
The typical television receiver of to-day has one r.f. stage followed by a frequency-changer and two vision-channel and two sound-channel i.f. stages. Diode detectors and diode noise limiters are used in each channel One video stage is used to feed the cathode of the c.r. tube and, in the sound channel, there are sometimes one and sometimes two a.f. stages.

Sync separation is effected by a pentode; but an auxiliary diode separator is usually employed as well to separate the frame sync pulses from the line. The frame time-base has two valves, the line usually three plus the e.h.t. rectifier. The tube operates at from 10 to 17 kV, the supply being derived from the line-flyback pulse. The h.t. supply for the line-scan output valve is augmented by an auxiliary supply (boost) which derives some 200 V at 100 mA from the energy stored in the deflector coils. The main power supply comes from the mains via a half-wave rectifier, and a.c./d.c. technique with a 'live' chassis is practically universal. Only about half the sets are suitable for operation from a d.c. supply, however, for the other half include an auto-transformer.



Marconiphone VT62DA with a 15-in. tube. The controls are grouped on a side panel.

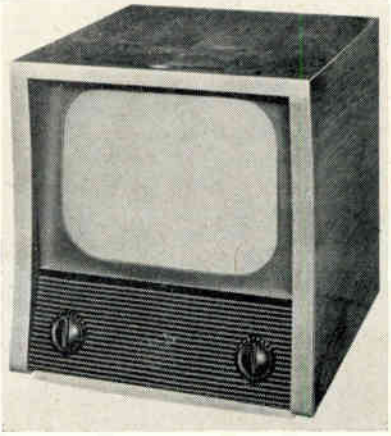
Tube sizes range from 9 in. to 27 in.! The most common sizes are the 12-in., 14-in., and 17-in. and these accounted for 39%, 29% and 20% respectively of the television sets sold during the first half of 1953. Apparently only about 3% of sales are now of the 9-in. and 10-in. tube sizes and only two firms, Bush and Pye, showed 9-in. models. The latter firm, as well as having one of the smallest types also showed the largest with a 27-in. diagonal directly-viewed tube having a total deflection angle of 90°. The tube operates at 16·5 kV. H.M.V. exhibited a set with a 21-in. tube, as they have done previously.



Peto-Scott television set with a rectangular tube.

In spite of the present popularity of the 12-in. tube, it seems likely that it will soon be ousted by the 14-in. which has now appeared in table models at attractive prices. The rectangular screen is much favoured for 14-in. and 17-in. tubes since it enables an appreciable reduction to be made in the size of the set. For the same reason, all but the smallest tubes are now of the wide-angle type, the beam being deflected through a total angle of about 70°. Ion traps and aluminized screens are commonly used. The dark screen, grey filter, call it what you will, has obviously come to stay for many tubes are now made with the face itself of grey glass. It is usual to employ glass with a 60% light-transmission factor, which gives an appreciable improvement of contrast in the picture when viewing under conditions of high ambient lighting, without too much brightness being lost.

There is a tendency to mount the tube so that its screen is leaning forward slightly. Sometimes the safety glass (implosion guard) leans forward rather more. This is done to reduce the reflection of room lighting, which is usually above the set and which would otherwise be reflected straight into the eyes of the viewer.

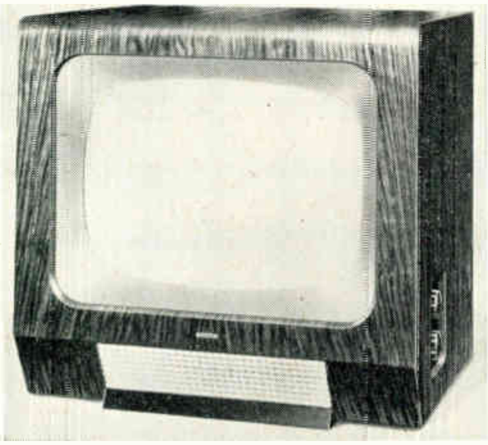


Kolster-Brandes KV35 with 14-in. tube and tilted glass screen to reduce reflections.

Ferguson have adopted a scheme which they call Halolight, and which has been applied in the cinema. This is to light the surround of the tube by means of fluorescent lamps, the amount of illumination being adjustable by the viewer.

One of the most noticeable features of to-day's sets is their relatively small size. The overall dimensions of a set producing a picture of 17-in. diagonal are much the same as those of older models giving only a 12-in. picture. This comes about partly because of the general use of miniature valves and partly because of the adoption of the rectangular tube with the 70° deflection angle. Further reduction of size is foreshadowed by the G.E.C. 12-in. tube with a 90° deflection angle. This is still in the development stage but is an indication of what may be expected in the future.

In their internal details, the different sets vary very much. The efficiency line-scan system using an auto-transformer with a low-loss core for coupling the driving pentode to the deflector coils and efficiency diode is now practically universal. An overwinding steps up the fly-back pulse to 10–17 kV for the e.h.t. rectifier. The variations in this part of the circuit are mainly in the methods of achieving a linear scan. The well-known resonant auxiliary transformer is still quite widely used, but the saturated-coil method which was introduced two years ago is now a serious rival to it. This coil is included in series with the deflector coil and the degree of saturation of its core is adjustable as a linearity control by means of a movable permanent magnet. The aim of the system is to make the voltage-drop across the combination of deflector coil and linearity coil fall, instead of rising, with increasing current, so that the current–voltage characteristic will fit that of the valves.


Cossor 17-in. tube model type 933.

The driving valve is a pentode and it is driven at its grid either by a saw-tooth or by a pulse or some combination of the two. The pulse drive is well suited to operation with a limited h.t. supply voltage, for the valve is driven below the knee in its characteristic to a point where it is of low resistance, and so the maximum anode-voltage swing is achieved. Both pentode and the efficiency-diode characteristics must be taken into account in designing the linearity circuit, therefore. With a saw-tooth drive at the grid, operation is restricted to the region above the knee and the permissible anode swing is reduced. However, only the diode characteristic need be considered in the linearity circuit and there is greater freedom of design here.

With pulse drive, the pulse is usually obtained with the aid of an extra triode which forms a multivibrator with the pentode. However, sometimes it is taken from a winding on the anode-circuit transformer, so that the driving valve is self-oscillating, as in the Ambassador TV10 and the Ferguson 998T receivers. With a saw-tooth drive, a blocking-oscillator saw-tooth generator is the favourite but the multivibrator type is sometimes employed, the saw-tooth being obtained by using a shunt capacitance in the pentode grid circuit.

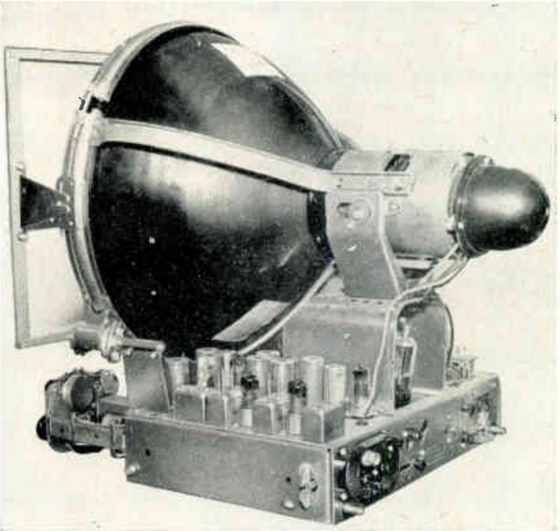
The blocking-oscillator seems to be reviving in popularity and is very widely used, not only as a line saw-tooth generator, but also as a frame generator. The valve is commonly a triode and the saw-tooth voltage which appears across the grid-circuit capacitor is used.

In the frame circuit, the use of a pentode amplifier after the saw-tooth generator is usual and this generally feeds the deflector coils through a transformer, although occasionally an auto-transformer is used and, in a few instances, resistance-capacitance (Cossor) or choke-capacitance (Ambassador) feed to a high-inductance deflector coil. The transformer draws a large magnetizing current and considerable pre-distortion of the waveform is needed to counter its effect. There are three basic ways of producing this and all are in use. The first, and probably the most popular, is the Blumlein linearity circuit, or some modification of it, which is a negative-feedback circuit having an RC correcting circuit in the feedback path. The second is the use of a passive RC correcting circuit between the saw-tooth generator and the output valve. The third is the utilization of the curvature of the output-valve characteristic. In the first two methods, some elements of the RC networks are made adjustable for linearity control. In the third, the grid bias of the valve is adjustable.

This last method is very simple and is popular in the U.S.A. but has not found much application in this country up to now. This year, however, its use is more common.

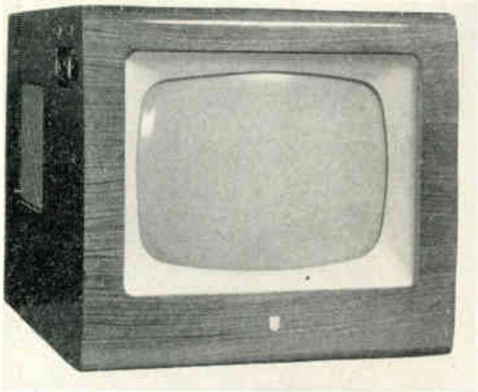
Synchronizing methods are, in the main, unchanged. It is usual to employ a pentode separator, the video signal being fed to its grid with positive-going sync pulses and d.c. restored there. The output, of sync pulses only, is fed to the line time-base through a simple differentiator. It also often feeds an integrator, the output of which is applied via a diode to the frame time-base, the diode removing the line pulses. Other methods are used, however.

Last year, it appeared as if fly-wheel sync methods were coming to the fore, since several receivers using them appeared. They are universally employed in the U.S.A., and it was thought that they might this year show signs


Murphy television receiver chassis. The deflector-coil assembly and the focus magnet are enclosed in a metal case which forms part of the tube mounting.

of becoming general. This is not so, however, and there are few more examples than last year. It is retained in the Ferguson set introduced last year, but is not employed in this firm's latest model. The former, however, is a set in the 'fringe-area' category whereas the new one is not. Murphy use it in the V214A and V216CA receivers but not in the V214 nor in the V216C. The two former are special 'fringe-area' models of the latter. It is clear that the present general tendency is to use flywheel sync only in fringe-area sets and to retain the normal pulse-locking methods for the other models.

Depending as it does upon the cumulative effect of a large number of sync pulses, flywheel sync is affected relatively little by noise and interference. It is therefore of most benefit in fringe areas where the signal-to-noise ratio is likely to be poor.



Philips 1726U table model with 17-in. tube. A fringe-area model (1726UF) is available.

There are three or four different forms of flywheel sync but all depend on a comparison between the phase of the line sync pulses and that of a waveform derived from some part of the line time-base. A phase discriminator is used to produce a d.c. output of magnitude dependent on the magnitude of the phase error and of polarity dependent on the direction of the error. An integrating circuit of long time constant is used, so that the response is virtually only to mean values. The circuit is essentially a negative-feedback one and the usual stability problems of such systems arise.

The problem of interference is tackled in a different way in the Philips and Stella sets. In these, the normal pulse lock is retained but an endeavour is made to clear the pulses of interference by using much more severe limiting than usual. Instead of using merely a differentiator between the main sync separator and the line time-base, two triode limiter stages are added, one before and one after the differentiator. The aim is to select and amplify only a very narrow slice through the sync pulse and so make the circuit immune to interference except that which occurs on the actual selected edge of the pulse.

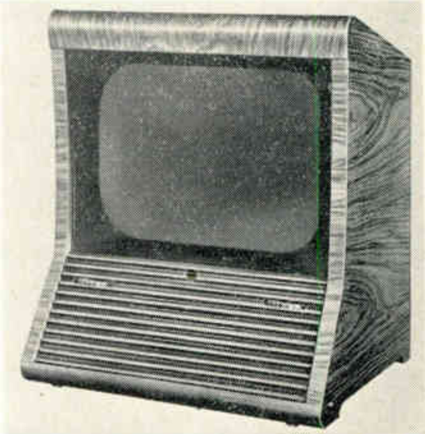
The practice of feeding the video signal to the cathode of the c.r. tube by a direct connection is now standard and there are very few sets which have any other arrangement. The video stage itself is nearly always a pentode with often, but not always, inductive compensation for the unavoidable capacitance in the anode circuit. It is becoming quite common to supplement this by a small capacitance in shunt with the cathode-bias resistor. In terms of frequency response, this gives a characteristic rising at the higher frequencies. In one case, at least (Murphy) there is an adjustable RC circuit here to give a user control of picture quality.

An exception to the general rule of one video stage occurs in the case of some Pye sets for in these a cathode-follower is interposed between the video pentode and the c.r. tube. This is done mainly because the sets have an a.g.c. system which requires to be fed with the video signal from a low-impedance source. The a.g.c. system, which is termed automatic picture control by the makers, is a true black-level type.[1]

A pulse is derived from the line-scan circuit and delayed slightly so that it occurs during the back porch of the video signal; that is, during the 6·5-μsec interval of black level which follows each 10-μsec sync pulse. The pulses, one for each line scan, are applied to the d.c. restorer which clamps one level of the pulses to the bias point of the d.c. restorer. This bias is actually the video signal from the cathode follower and the bias voltage is the black level of the signal, since the pulses occur during the black-level intervals only.

Although the pulses are all of the same amplitude they are, in effect, superimposed on the black-level amplitude of the signal, and the combined peak amplitude varies with the black level. A peak diode rectifier and integrator produce an output proportional to black level and this is applied as bias to an i.f. stage. The circuit tends to keep the black level at the tube constant and to function to prevent any change of signal strength from affecting the output, while leaving unaffected the normal picture changes.

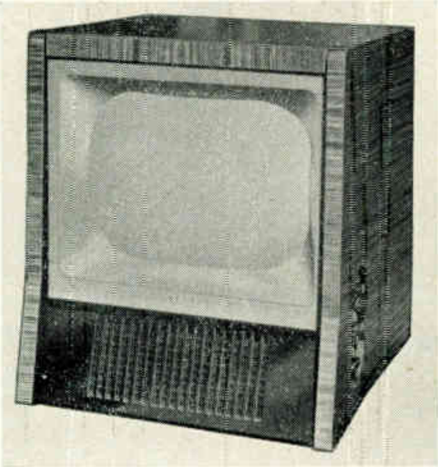
Most television sets have a.g.c. only on the sound channel. There is now a tendency to fit it to the video channel also. True black-level operated schemes, such as that just described, tend to be complex and some people are fitting simpler arrangements which are operated by the mean brightness of the picture. Provided that the mean brightness required is constant, they can control the signal equally well. However, if the mean brightness varies intentionally, as between a daylight



Pye V4 and V7 models with 14-in. and 17-in. tubes have the same external appearance with a forward-tilted screen.

and a night scene in a picture, this form of a.g.c. system operates to remove the intentional change.

Most pictures transmitted do appear to have a mean level which is roughly constant, and it is quite rare for there to be marked variations from it. In spite of the theoretical objections to it, therefore, an a.g.c. system operating on the mean-brightness level can be quite satisfactory on most pictures.



Ekco model T205 with 14-in. tube and vision-channel a.g.c.

One very simple example of this form of a.g.c. is to be found in the Ekco T207. A voltage is normally developed on the grid of the sync separator by the d.c.-restoring action of its grid circuit and a portion of this is applied through an integrator to the grid of an i.f. valve as bias. A manual gain control is associated with it and a diode is included to prevent the i.f. grid from being driven positive. By its normal d.c.-restoring action the mean grid voltage of the sync separator depends on both the total amplitude of the signal and on the mean brightness of the picture.

Turning now to the signal chain, it is the usual practice to employ two i.f. stages in the video channel and one or two in the sound. Diode detectors are used in both. The sound channel is usually given a bandwidth of 50–100 kc/s, partly to permit some drift of the oscillator frequency and partly because such a bandwidth is necessary to obtain good action of the noise limiter. If this is to be effective, it is essential for noise pulses to be kept of as short a duration as possible.

On the video side, a bandwidth of some 2·75 Mc/s is desirable but an attenuation of at least 40 db at only 0·75 Mc/s from one edge of the band and 50 db at 1·5 Mc/s from the other edge is necessary if a high-quality picture is to be obtained with freedom from interference by the sound channels of the same and an adjacent channel. Bandwidths actually obtained range from 2 Mc/s to 2·75 Mc/s.

The high attenuation at the sound channels is usually obtained with the aid of wavetraps and up to four such circuits are used, but two is a more common number. They are generally coupled in some way to the intervalve couplings, but are sometimes instead inserted in the cathode lead of an i.f. valve. Pairs of damped coupled tuned circuits are often used as the intervalve couplings and most sets have at least one such pair. They are not used exclusively, however, and a mixture of coupled pairs and single tuned circuits is very general.

Separation of the vision and sound signals usually takes place after the frequency-changer, but sometimes the first i.f. stage is used as a common amplifier. Frequency-changing itself is often done by a single-valve mixer-oscillator, the oscillator tuned circuit being connected between the control and screen grids in Hartley or Colpitts fashion and the signal from the r.f. stage being fed to a tapping at the null point on it. A close runner-up, however, is the two-valve frequency-changer using a triode pentode or a double triode.

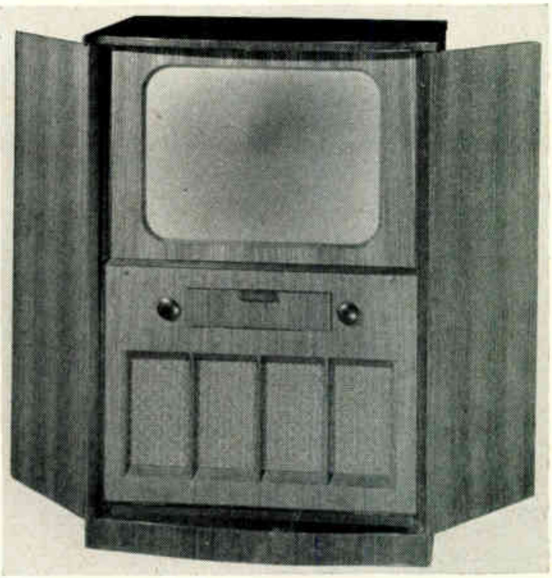
One r.f. stage is usual, but sometimes two are employed, notably in the Murphy sets. The number of tuned circuits at signal frequency varies from two to four.

Station-selection methods vary quite a bit, but there is undoubtedly a trend towards providing continuous tuning over the television band. Composite dust-iron and metal slugs are used and usually operated by separate knobs at the rear of the set. The oscillator is often provided with a rough tuning scale. Sometimes the controls are ganged.

Philips retain the system of plug-in coils, which they have now employed for some time. In general, there has been little change in selection methods, but some increase in the use of continuous tuning.

It is noticeable that, in the interests of safety, greater care is being taken over the aerial circuit. It is usual to have an earth terminal by which the aerial and feeder can be connected directly to a local earth, while the feeder itself is isolated from the set by capacitors and/or a double-wound input transformer. This is made necessary by the use of a.c./d.c. technique with which the chassis is live to the supply mains.

This technique is practically universal although not all the sets are suitable for d.c. supplies. Many include an auto-transformer for the h.t. supply with separate windings for one or two 'difficult' valve heaters. In



Ferranti projection set, model 22K3, giving a 22-in. picture.

the true a.c./d.c. set, a thermistor is invariably included in the heater chain to safeguard the c.r. tube, which has a heater of very different characteristics from those of the valves. G.E.C., indeed, go a step further and have revived the barretter, which they use as well as a thermistor.

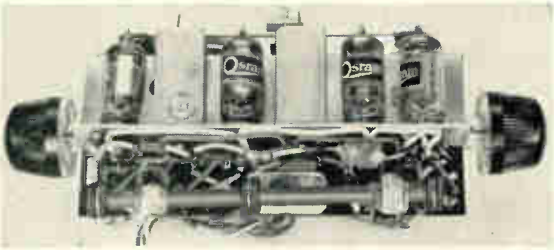
Projection television sets are still to be found, although in rather fewer numbers than some years back. In their general form they have not changed much and are still based on a 2½-in. tube operating at 25 kV and have a Schmidt optical system. The electrical part does not differ greatly from the directly-viewed sets; the time-bases are a little simpler, tube-protection circuits are included, and the e.h.t. comes from a separate generator using a ringing-choke system. The improvements in projection television are mainly in detail; e.h.t. regulation has been improved and viewing screens are better.

Television aerials are still basically of the same form, the H and X types being the commonest. For fringe areas three- and four-element types are used. Considerable improvement in detail has taken place, however, notably in making them more weather and corrosion proof.

A feature of some Antiference models is the avoidance of the possibility of a bad connection to an aerial rod by making the connection in the form of a capacitance coupling to the rod itself.

A certain amount of test equipment for television was shown. Telequipment have an r.f. generator which is modulated by a square wave having a 0·05-μsec rise time. It is intended for the production testing of the transient response of television receivers and the display is on a 3½-in. tube provided with marker pips at 0·05-μsec intervals. This same firm also showed a television field-strength meter which comprises a five-channel receiver, the output signal being displayed on a 1-in. c.r. tube. It has a sensitivity of 10 μV, and there are calibrated gain controls; it is intended to assist dealers to determine what form of aerial to recommend.

In sound-broadcast receivers an interesting development is the use of 'r.f. iron' as a core for a frame aerial. The use of dust-iron cores for such aerials is well known in aircraft d.f. systems[2]; ferrite cores are now being used in this broadcast application. One example of this practice is the Cossor model 512, a four-valve a.c./d.c. transportable. The aerial comprises a long ferrite rod carrying a winding of small diameter and, therefore, quite unlike the conventional frame aerial.



Chassis of the Decca Deccette. The 'frame aerial' is the long ferrite rod at the bottom carrying several short coils.

The mains/battery receiver is well known. Some of the types now made are primarily battery-operated portable sets with provision for mains operation when required. In the Decca 'Deccette', which also has a ferrite-cored frame aerial, this takes the form of a separate power-supply unit which acts as a base for the set when it is used with mains supply. The Bush BAC31, on the other hand, has the mains equipment built in.

One unusual piece of apparatus is the H.M.V. 1507 portable radio-gramophone. For a.c. operation, it includes a 5-valve receiver covering long, medium and short waves in three bands and a three-speed gramophone; it weighs only 19½ lb.

Magnetic tape recorder incorporating the Truvox Mark III tape mechanism.

One feature of this year's exhibition was the great interest in recordings. The apparatus falls into two groups; that concerned with the reproduction of disc records, and that for magnetic tape. In the first category, it is noteworthy that record-changers now handle any of the three types of disc—the 78, 45 and 33⅓ r.p.m.—not as a mixture, it is true, but any one at will, by means of a simple adjustment. A development in pick-up design is illustrated by the H.M.V. model which includes a ceramic piezoelectric element.

In the second group, magnetic tape, the apparatus comprises recorder and reproducer in one unit. One of these, the 'Reflectograph' [Rudman, Darlington (Electronics), Ltd.] is unusual in having a continuously-variable tape speed. A synchronous motor is used and the variation is obtained in the reduction drive. It is provided so that the tape speed and, therefore, the playing time of a reel of tape, may be adjusted to suit the length of programme and also to provide a precise adjustment of pitch on replay.

More normal tape recorders, such as the Wright & Weaire Tape-Deck, have two tape speeds (7½ in./sec and 3¾ in./sec), and give playing times of 45 and 90 minutes with a 1,750-ft reel. Rewind and erasing facilities are included. The tape is ¼ in. wide and can take two parallel sound tracks each 0·1 in. wide.

A special long-duration instrument is the Simon LDT7 Tape Monitor which gives a duration of 4 hours on one 2,400-ft reel of ¼-in. tape.

REFERENCES

[1] "Vision A.G.C.", *Wireless World*, April 1953.

[2] "Iron-Cored Loop Receiving Aerial", by R. E. Burgess, B.Sc., *Wireless Engineer*, July 1946, p. 172.

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.

Bandwidth Limitation of Junction Transistors

Sir,—With the aid of Steele's paper[1] it is possible to calculate a fundamental limit to the wideband figure-of-merit g_m/C of junction transistors, if the output capacitance is not taken into account.

In valves, a current $i_g = v_g j\omega C_i$ is applied to the grid and it will give an anode current $i_a = g_m v_g$. So we can write:

$$\frac{g_m}{C_i} = \frac{i_a}{i_g} j\omega.$$

In junction transistors the base can be considered as a control grid and the collector as an anode, so

$$\frac{g}{C_i} = \frac{i_c}{i_b} j\omega = \frac{\alpha j\omega}{1 - \alpha} = \frac{2j\omega}{(1 + j\omega\tau_B)w^2/L^2{}_B}$$

where $\alpha = \frac{i_c}{i_c + i_b}$, τ_B is the lifetime of the minority carriers in the base, L_B is their diffusion length and w is the width between emitter and collector.

As generally $\omega\tau_B \gg 1$:

$$\frac{g}{C_i} = \frac{2}{\tau_B}\left(\frac{L_B}{w}\right)^2 = \frac{2D}{w^2} = \omega_0 \text{ [formula (25) of Steele's paper]}$$

where D is the diffusion constant of minority carriers.

This wideband figure-of-merit appears to be equal to the frequency cut-off ω_0 of α. The series resistance of the base and the in-phase component of the term $i_b/(i_c + i_b)$ will give a departure from this rule but several methods exist to make both influences small. The direct collector current I_c does not appear in the formulae. This follows from the fact that both the slope and the base-emitter capacitance are proportional to I_c.

As for germanium the diffusion constant of holes $D_p = 45$ cm²/sec and the diffusion constant of electrons $D_n = 97$ cm²/sec the necessary width w between emitter and collector to attain a figure-of-merit g/C_i of 1 mA/V per picofarad is 3 microns for a p-n-p transistor and 4·4 microns for an n-p-n transistor.

A. J. W. M. van Overbeek.
F. H. Stieltjes.

Philips Research Laboratories,
Eindhoven, Holland.
25th August 1953.

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[1] Earl L. Steele. "Theory of Alpha for p-n-p Diffused-Junction Transistors", *Proc. Inst. Radio Engrs*, Vol. 40, p. 1424, November 1952.

Nomenclature for E.M. Wave Modes

Sir,—Professor Barlow's suggestions in the September issue for the standardization of nomenclature in waveguide modes fail in one respect, where there has always been confusion, in defining unequivocally the order of the subscripts of modes in rectangular guides.

In addition to following the accepted sequence of the co-ordinates—and I think most would agree with the Editor's comment that xyz is the accepted sequence rather than yzx—it is essential also to specify the orientation of the co-ordinate axes relative to the waveguide. It is normal practice, probably for artistic reasons, to draw a rectangular waveguide with its broad side horizontal. Following Professor Barlow and using the x-direction as the direction of propagation (sequence yzx) we can have either of the arrangements shown in Fig. 1 (a, b) which, with subscripts following the co-ordinate sequence, give rise to the notations H_{01} and H_{10} respectively for the principal wave. Both are possible, although when the waveguide is drawn in cross-section, as in Fig. 1 (c, d), the former arrangement appears awkward. I suggest it is more convenient to use the normal xyz sequence (with propagation along the z axis), which gives rise to the satisfactory diagrams of Fig. 2 and the notation H_{10} for the principal mode.

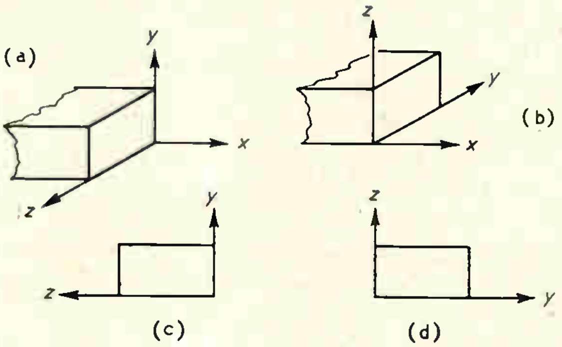

Fig. 1

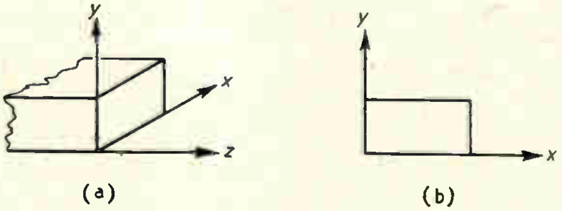

Fig. 2

A British standard does exist which defines the notation for the modes in rectangular and circular guides (Supplement No. 1 (1948) to B.S.204 : 1943). For the H_{mn} mode in a rectangular guide the definition reads "the subscripts m and n denote the number of half-period variations in the electric field parallel to the short and long sides respectively of the guide." This definition avoids the possible ambiguities of a notation based on a co-ordinate system and for that reason I consider it preferable. However, the choice of subscript order, which is arbitrary, has been such as to make H_{01} the notation for the principal wave. This is at variance with what I have suggested above as the natural choice and is, moreover, at variance with the more popular present day usage, H_{10}. (In an examination of 20 text books the preference was found to be 4 : 1 in favour of H_{10}.) This is a matter in which the British Standards Committee might consider a revision before the present supplements are incorporated in B.S.204.

H. R. L. Lamont.

Pinner, Middx.
11th September 1953.

Harmonic Distortion and Negative Feedback

Sir,—In reply to Mr. Kerr's letter in the September issue, referring to my article "Harmonic Distortion and Negative Feedback," I will not quibble over the point raised in the first paragraph of his letter: it would have been better if I had written " . . . deviate greatly from linearity . . .".

The theory which he expounds in the remainder of his letter, however, is rendered invalid, owing to a grievous error which he makes early on, namely, in the statement that "the gain A is a function of the input voltage." This is quite true for an amplifier without feedback, but where feedback occurs, the gain A is a function of the grid-cathode voltage. The expression for the output voltage with feedback which he writes as

$$V_0' = \frac{V_0}{1 + A\beta} = \frac{V_0}{1 + \beta f(V_i)} = \frac{V_i f(V_i)}{1 + \beta f(V_i)}$$

should have been written

$$V_0' = \frac{V_i f(V_i - \beta V_0')}{1 + \beta f(V_i - \beta V_0')}$$

from which V_0' cannot be extrapolated in the general case.

R. O. ROWLANDS.

Evesham, Worcs.
15th September 1953.

NEW PUBLICATIONS

Radio Research, 1952

(Including the Report of the Radio Research Board and the Report of the Director of Radio Research.) Pp. 51 + iv. Published for the Department of Scientific & Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 2s.

Fundamental Processes of Electrical Contact Phenomena

By F. LLEWELLYN JONES, M.A., D.Phil., F.Inst.P. Radio Research Special Report No. 24. Pp. 66 + vi. Published for the Department of Scientific & Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 3s.

British Standard for Measurement of Frequency Variation in Sound Recording and Reproduction (B.S. 1988 : 1953)

The British Standards Institution, Sales Branch, 2 Park Street, London, W.1. Price 1s.

Permanent Magnets

Pp. 64. Mullard, Ltd., Century House, Shaftesbury Avenue, London, W.C.2.

This booklet contains information on the design and performance of magnetic circuits, data on magnetic materials, comparative data of British and American permanent magnets and details of the Mullard standard range of Ticonal, Magnadur and Reco magnets.

Guide to Broadcasting Stations (7th Edition)

Pp. 104. Published for *Wireless World* by Iliffe & Sons, Dorset House, Stamford Street, London, S.E.1. Price 2s. (Postage 2d.)

This booklet lists all European long- and medium-wave broadcasting stations and over 1,700 short-wave broadcasting stations, as well as television and other v.h.f. stations. The data is corrected to 1st August 1953.

BRIT.I.R.E. PRESIDENT

For the second year in succession, William E. Miller, M.A.(Cantab.), Editor of the *Wireless & Electrical Trader*, has been elected by the Council of the British Institution of Radio Engineers as President.

During his 20 years' membership of the Institution, Mr. Miller has served on six of its standing committees and on the General Council and he was elected Vice-President in 1948.

OCTOBER MEETINGS

I.E.E.

8th October. President's Inaugural Address, by H. Bishop, C.B.E., B.Sc.(Eng.).

14th October. Radio Section Chairman's Address, by J. A. Smale, C.B.E., B.Sc.

19th October. "Television." Discussion to be opened by H. Bishop, C.B.E., B.Sc.(Eng.).

26th October. "Long-Playing Disc Records compared with Magnetic Tape for Sound Reproduction in the Home." Discussion to be opened by H. F. Smith

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

Brit.I.R.E.

21st October. Annual General Meeting at 6 p.m. to be followed at 7 o'clock by "Printed Circuits: Some Principles and Applications of the Foil Technique", by P. Eisler, Dr. Ing. To be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for August 1953

Date 1953 August	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	— 1·1	— 1	N.M.
2	— 1·2	— 1	N.M.
3	— 1·1	— 1	N.M.
4	— 1·1	— 1	+ 38·9
5	— 1·1	— 1	+ 38·8
6	— 1·2	— 1	+ 38·4
7	N.M.	0	+ 37·5
8	— 1·1	— 1	N.M.
9	— 1·0	— 1	N.M.
10	— 1·0	— 1	+ 36·3
11	— 1·0	0	+ 35·7
12	— 1·0	0	+ 34·3
13	— 1·0	0	+ 33·5
14	— 0·9	0	+ 33·0
15	N.M.	0	N.M.
16	N.M.	N.M.	N.M.
17	— 0·9	+ 1	+ 28·4
18	— 0·9	+ 1	+ 26·4
19	— 0·9	0	+ 24·8
20	— 0·9	+ 1	+ 23·0
21	— 0·8	+ 2	+ 21·3
22	— 0·8	+ 1	+ 19·9
23	— 0·8	N.M.	+ 18·1
24	— 0·8	+ 1	+ 16·4
25	— 0·8	+ 2	+ 14·7
26	— 0·7	+ 1	+ 13·3
27	— 0·7	+ 1	+ 11·3
28	— 0·7	+ 1	N.M.
29	— 0·7	+ 3	+ 8·7
30	N.M.	N.M.	+ 7·3
31	— 0·7	+ 3	+ 5·8

The values are based on the astronomical data available on 1st September 1953.
N.M. = Not measured.

ABSTRACTS and REFERENCES

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

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

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ACOUSTICS AND AUDIO FREQUENCIES

534 : 061.3 **2855**

Trends in Electro-acoustics.—(*Wireless World*, Aug. 1953, Vol. 59, No. 8, pp. 352–354.) Report of the proceedings at the International Electro-acoustics Congress held in the Netherlands, June 1953. The subjects covered were (*a*) sound recording, (*b*) public-address systems, (*c*) acoustic measurements, (*d*) hearing aids and audiometers, (*e*) electro-acoustics in ultrasonics, (*f*) electro-acoustics applied to musical instruments, (*g*) sound insulation of light-weight structures.

534.01 **2856**

Response of Linear Time-Dependent Systems to Random Inputs.—D. B. Duncan. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 609–611.)

534.21 **2857**

On a Variational Principle in Acoustics.—O. K. Mawardi. (*Acustica*, 1953, Vol. 3, No. 3, pp. 187–191.) A generalized theory is presented for the propagation of sound waves in a tube of rectangular cross-section with staggered openings in thin partitions placed at periodic intervals. A variational principle is applied to determine the wave-propagation constant for the labyrinth, and the phase velocity and attenuation for the simpler type of waves propagated are evaluated.

534.213.4-14 **2858**

Propagation of Acoustic Waves in a Liquid-Filled Cylindrical Hole surrounded by an Elastic Solid.—E. V. Somers. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 515–521.) The theory of propagation in systems in which discontinuities in the physical parameters occur along coaxial cylindrical surfaces is developed for an annular impulsive source. The field problem is solved for a basic model consisting of a cylindrical liquid column surrounded by an elastic solid.

534.22 : 534.321.9-14 **2859**

A New Precision Method for the Measurement of Ultrasonic Velocities in Liquids.—B. R. Rao & K. S. Rao. (*Nature, Lond.*, 13th June 1953, Vol. 171, No. 4363, pp. 1077–1078.) Description of a simple method based on diffraction effects in a resonant liquid column. It is suitable only for transparent liquids and cannot be used at very high frequencies. Results obtained at frequencies in the range 1–3 Mc/s are quoted for water and some organic liquids.

534.231 **2860**

Remarks on the Concept of Acoustic Energy.—A. Schoch. (*Acustica*, 1953, Vol. 3, No. 3, pp. 181–184.) The conventional expressions for acoustic energy are shown to have characteristic properties which justify their use in acoustics. See also 2194 of August (Markham).

534.24 + [538.566 : 535.3 **2861**

The Scattering of Waves at an Uneven Surface.—Brekhovskikh. (See 2968.)

534.24 + [538.566 : 535.3 **2862**

Scattering of Waves by a Statistically Irregular Surface.—Isakovich. (See 2969.)

534.321.9 : 534.23 **2863**

Transmission of Ultrasonic Waves through a Plane Plate made of Viscoelastic Material Immersed in a Liquid Medium.—Y. Torikai. (*J. phys. Soc. Japan*, March/April 1953, Vol. 8, No. 2, pp. 234–242.) Formulae relative to transmission through a nonabsorptive plate in a liquid medium are quoted and the corresponding formulae for the case of an absorptive plate are derived. The expression for the ratio of the incident and transmitted energy is of the same form as for the nonabsorptive plate, but the propagation constants are complex quantities. When the absorption coefficient of the plate is appreciable, a simple approximate formula can be applied which gives results in good agreement with experiment.

534.321.9-14 **2864**

Demodulation of Ultrasonic Waves in Liquids.—K. Altenburg & S. Kästner. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 161–165.) The relations between the ultrasonic-wave energy density, l.f. sound amplitude and the depth of modulation, for a modulated ultrasonic wave, are investigated theoretically and experimentally. In water, the l.f. amplitude increases linearly with the depth of modulation and is also proportional to the energy density of the carrier wave. The second harmonic of the l.f. sound is proportional to the square of the depth of modulation. The experimental method may be used to determine the energy density of ultrasonic waves of intensity $> 0{\cdot}25\ \mu$bar.

534.522.1 **2865**

The Application of the Phase Method in Visualizing Ultrasonic Waves.—Y. Torikai & K. Negishi. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 119–124.) Theory of the phase microscope is extended to the visualization of ultrasonic waves and confirmed experimentally. Comparison is made with the schlieren method.

534.614 + 534.64 **2866**

The Properties of Water-Filled Tubes for Acoustic Impedance and Velocity Measurements.—W. Kuhl. (*Akust. Beihefte*, 1953, No. 1, pp. 111–123.)

534.64 **2867**

The Practical Representation of Standing Waves in an Acoustic Impedance Tube.—W. K. R. Lippert. (*Acustica*, 1953, Vol. 3, No. 3, pp. 153–160.) The pressure/distance envelope function is considered for highly reflecting to normally absorbing samples and for highly absorbing samples. Measurement of the s.w.r. is described, and the detection of disturbances of the equipment, by comparison of envelope curves derived from theory with those obtained by experiment, is illustrated by examples. A definition of the resolving power of an impedance tube is proposed and discussed.

534.641 **2868**

Acoustic-Impedance Measurement by the Transmission-Characteristic Method.—A. F. B. Nickson & R. W. Muncey. (*Acustica*, 1953, Vol. 3, No. 3, pp. 192–198.) The transmission-characteristic method used by Harris (3577 of 1945) is applied to the determination of acoustic impedance of materials at various frequencies and angles of incidence. The most suitable dimensions are determined for the reverberation chambers used, and their construction is described. Allowances can be made for the impedance of the bare walls, so that for comparison measurements over a range of frequencies, wooden boxes lined with sheet metal are satisfactory.

534.75 **2869**

Contribution to a Scientific Theory of Single-Channel Transmission of Sound: Part 3.—P. Burkowitz. (*Funk. u. Ton*, May 1953, Vol. 7, No. 5, pp. 241–249.) Data required in measurement and calculation of sound intensity and transit time are considered. Reverberation effects are discussed in relation to the theory. Part 2: 1890 of July.

534.75 : 534.322.1 **2870**

The Influence of Nonlinear Distortions on the Audibility of Frequency Vibrato.—E. Zwicker & W. Spindler. (*Akust. Beihefte*, 1953, No. 1, pp. 100–104.) Investigations were carried out on musical intervals with f.m. in the range 0·8–6·0 c/s, the loudness level at the receiving headphones being adjusted over a 2:1 range and nonlinear distortion introduced. The more consonant the musical interval, the sooner additional beats appear when the applied distortion is increased. This effect is partly due to the nonlinearity of the characteristic of the ear.

534.839 **2871**

The Measurement of the Attenuation of Impact Sound using Sinusoidal Excitation.—T. Lange. (*Acustica*, 1953, Vol. 3, No. 3, pp. 161–168. In German.) Sinusoidal excitation, instead of the standard hammer impact, is used to determine the attenuation of sound by ceilings. The two methods are compared.

534.843 **2872**

The Effect of Wall Shape on the Decay of Sound in an Enclosure.—J. W. Head. (*Acustica*, 1953, Vol. 3, No. 3, pp. 174–180.) The method of Feshbach (4026 of 1944) is applied to determine the effect of regularly spaced symmetrical projections on the natural frequencies and normal modes of a two-dimensional room. An explanation is given of the superior subjective properties of rooms with rectangular coffering, as shown experimentally by Somerville & Ward (14 of 1952).

621.395.623.7 : 534.62 **2873**

Direct Measurement of the Efficiency of Loudspeakers by Use of a Reverberation Room.—H. C. Hardy, H. H. Hall & L. G. Ramer. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 99–107.) A relatively simple method is described for integrating the power output from a loudspeaker excited by a warble tone in a large reverberation room. Typical measurement results are presented and correlated with results calculated from free-field measurements.

621.395.623.7 : 681.142 **2874**

An Analogue for Use in Loudspeaker Design.—Baruch & Lang. (See 3050.)

621.395.625.2 **2875**

The Calibration of Disc Recordings by Light-Pattern Measurements.—P. E. Axon & W. K. E. Geddes. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 217–227.) Analysis of the light-pattern formation suggests that more accurate results are obtained if the light pattern is observed in the focal plane instead of on the disk surface. In the apparatus developed, the pattern width is measured as the distance between centres of twin images after they have been adjusted to lie edge to edge. Other simplifications enable calibration accuracy to be attained with compact equipment involving relatively inexpensive optical items. Experimental results confirm theoretical expectations.

621.395.625.3 **2876**

Interference Effects in Magnetic Recording Heads.—A. H. Mankin. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 108–112.)

621.395.625.3.001.4 **2877**

Noise Test for Magnetic Recording Media.—R. C. Curtis. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 216–222.) D.c. bias current sufficient to saturate the medium to be tested is supplied from a filtered d.c. source through a resistance high compared with that of the head. Variations of voltage across the head are fed, after amplification, to a valve voltmeter. This measures total noise. With a suitable filter or harmonic wave analyser inserted the noise contribution in various parts of the spectrum can be measured. Typical test curves are shown and discussed.

AERIALS AND TRANSMISSION LINES

621.315.212 **2878**

Noise-Free Instrument Cable.—(*Commun. Engng*, March/April 1953, Vol. 13, No. 2, pp. 22..39.) Spurious electrical signals caused by mechanical shock and vibration arise in instrument cables due to the separation of static charges at the surfaces of the dielectric. Covering the inner and outer surfaces of the dielectric by a well-bonded conductive coating suppresses these unwanted signals. Details of laboratory procedure for making short lengths of noise-free cable are given. See also *Electronic Engng*, July 1952, Vol. 25, No. 305, p. 293.

621.392.029.63/.64 **2879**

The Use of Short Transmission Lines in Ultra-High Frequency Techniques.—Rohan. (See 2917.)

621.392.21 : 512.831 **2880**

Synthesis of Uniform Transmission Line.—D. W. C. Shen. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 287–289.) The application of matrix theory to network synthesis is illustrated. The uniform line is treated as an infinite number of quadripole networks and the limiting values of the associated matrices are evaluated.

621.392.26 **2881**

Attenuation and Surface Roughness of Electroplated Waveguides.—F. A. Benson. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 213–216.) Examination of the surfaces of copper- and silver-plated waveguides shows that the plated surfaces are generally rougher than and bear no relation to the original drawn-tube surface. After 19 weeks' exposure of waveguides to weathering, the attenuation at 9·375 kMc/s showed increases of up to 22%.

621.392.26 **2882**

Field in a Rectangular Waveguide with Conducting Membrane.—H. L. Knudsen. (*Onde élect.*, April 1953, Vol. 33, No. 313, pp. 217–234.) General theory is given of the field in a waveguide divided into several homogeneous sections by longitudinal conducting membranes parallel to one of the walls. The only possible waves in such waveguides are the longitudinal H and E types. Calculation of the components of these waves involves the solution of a transcendental equation whose roots are, in general, complex. A detailed study is made of the TE_{mo} waves in a rectangular waveguide with a single membrane. The dependence of the types of wave on the position of the membrane, the resistance per unit of surface, and the frequency, is shown by a curve in the complex plane for each type of wave. From these curves the attenuation and phase constants and the field configuration for the different types of wave can easily be found. Such calculations are made for typical cases. With suitable choice of the position of the membrane and of its unit-surface resistance, a value of the attenuation or phase can be obtained, for a given type of wave, that is fairly independent of frequency over a wide band. Theory of attenuators in which the conducting membrane does not extend from one waveguide wall to the opposite one is being developed.

621.392.26 **2883**

Electromagnetic Propagation through Waveguides of Rhombic Cross-Section.—W. B. Swift & T. J. Higgins. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 274–283.) Analysis is presented for waveguides of rhombic cross-section which permits determination of the effects produced in rectangular waveguides by deformation. The frequency of the lowest possible mode is calculated for waveguides (*a*) with sides of equal width, (*b*) with sides of width 2*a* and *a*, the smaller angle of the cross-section ranging from 60° to 90°. 36 references.

621.392.26 **2884**

The Theory of a Waveguide containing a Spiral, Partly Filled with a Dielectric.—V. P. Shestopalov. (*Zh. tekh. Fiz.*, March 1952, Vol. 22, No. 3, pp. 414–425.) Mathematical theory is developed which enables determination to be made of the structure of the field in the waveguide, the phase velocity and dispersion, the energy flow inside and outside the spiral and in the dielectric, and the losses in the spiral and in the waveguide wall. The advantages and disadvantages of this type of waveguide are enumerated.

621.392.26 **2885**

Duo-Dielectric Coaxial Waveguide.—R. E. Beam & D. A. Dobson. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 301–312.) Analysis is presented for a cylindrical waveguide with coaxial polystyrene rod. The guide wavelength is determined and curves are given showing its variation with frequency for the E_{00} E_{01}, H_{01} and HE_{11} modes. Representative field-distribution curves are also given for the various modes. The characteristic impedance for the principal mode E_{00} is determined. This mode has an axial component of electric field which vanishes at zero frequency. At high frequencies the field is concentrated mainly within the polystyrene rod and the radius of the outer conductor has little effect on the propagation in the waveguide.

621.392.26 **2886**

Cut-Off Frequency for Circular Waveguides containing Two Coaxial Dielectrics.—R. D. Teasdale & G. N. Crawford. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 296–300.) A conditional equation is derived for a cylindrical waveguide with a coaxial dielectric rod. The solution of this equation determines the cut-off frequency of the structure in the TM_{01} mode. A graphical method of solving the equation is described and curves are given which show how the cut-off frequency of the composite waveguide compares with that of the same waveguide wthout the dielectric rod.

621.392.26 **2887**

Electromagnetic Transients in Waveguides.—G. I. Cohn. (*Proc. nat. Electronics Conf.*, *Chicago*, 1952, Vol. 8, pp. 284–295.) Transient effects in cylindrical waveguides with arbitrary driving functions are determined in terms of Fourier integrals, the form of which gives some information on the general characteristics of the propagation of transients in waveguides. The Fourier integral is put into a form suitable for computation and a solution is obtained for the case of an applied sine-wave voltage with a unit-step-function envelope.

621.396.26 : 538.614 **2888**

Modes in Waveguides containing Ferrites.—M. L. Kales. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 604–608.) Formal solutions are obtained of the equations for wave propagation in waveguides of circular section containing a ferrite material, and with an axial magnetic field. TE, TM and TEM modes are shown to be non-existent. Variation of the relative phase of the two circularly polarized wave components with distance along the waveguide is demonstrated, corresponding to the Faraday effect. Similar results have been obtained by Suhl & Walker (2710 of 1952).

621.392.26 : 538.614 **2889**

The Faraday Rotation of Waves in a Circular Waveguide.—H. Gamo. (*J. phys. Soc. Japan*, March/April 1953, Vol. 8, No. 2, pp. 176–182.) Exact solutions for the magnetic rotation of waves in a circular waveguide of infinite length are obtained, the rotational terms in both magnetization and electric polarization being considered. The effect of the applied field is to produce two partial waves, with right-hand and left-hand circular polarization, which are neither pure TE nor TM modes, but reduce to these for zero field. Curves showing the frequency dependence of the propagation constants for both partial waves are given for the quasi-TE_{11} and quasi-TM_{11} modes for particular values of the coefficients of the rotational terms. The dependence of the cut-off frequencies on these coefficients is determined. For TM modes, the cut-off frequencies of both partial waves are identical.

621.392.26 : 538.614 **2890**

Experiments on the Faraday Rotation of Guided Waves.—A.A.T.M. van Trier. (*Appl. sci. Res. B*, 1953, Vol. 3, No. 2, pp. 142–144.) A cylindrical waveguide, with a Ni-ferrite rod mounted axially and a movable plunger sliding on the rod, was used to determine the Faraday rotation at 24 kMc/s. The two resonance lengths for each magnitude of the axial magnetic field were measured and the specific rotation calculated.

621.392.26 (083.75) **2891**

Rigid-Waveguide Specifications.—(*Commun. Engng*, May/June 1953, Vol. 13, No. 3, p. 32.) A table and graphs show the frequency range, attenuation, dimensions and maximum power-handling capacity of American standard sizes of rigid waveguide.

621.392.26.016.3 **2892**

High-Powered Microwave Dummy Loads.—T. N. Anderson. (*Tele-Tech*, May 1953, Vol. 12, No. 5, pp. 92–94..164.) The development of a series of dry dummy loads for waveguides is described. The loads are designed for operation, without auxiliary cooling, at 2-kW average power dissipation at frequencies in the range 2·6–12·4 kMc/s.

621.396.67 **2893**

An Experimental Investigation on Models of Aerials with Passive Directors.—D. M. Visokovski. (*Zh. tekh. Fiz.*, Sept. 1952, Vol. 22, No. 9, pp. 1477–1482; German translation, *Nachr Tech.*, Feb. 1953, Vol. 3, No. 2, pp. 59–62.) Measurements are reported on reduced-scale models of systems comprising a radiating dipole with a somewhat longer reflector and several shorter directors. The radiation patterns obtained differ considerably from those published by Uda and Yagi and reproduced in many text-books. The discrepancy is attributed to the unsuitability of their methods. The influence of the number of directors on the aerial gain and beam width is discussed.

621.396.67 **2894**

The Absorption Gain and Back-Scattering Cross-Section of the Cylindrical Antenna.—S. K. Dike & D. D. King. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 926–934.) Discussion on 2716 of 1952.

621.396.67 : 621.396.828 **2895**

New Antenna Fittings reduce P-Static Interference.—C. de Vore. (*Tele-Tech*, May 1953, Vol. 12, No. 5, pp. 77–79..143.) Illustrated description of insulated aerial fittings, developed at the Naval Research Laboratory, which reduce precipitation interference.

621.396.67.029.64 **2896**

Recent Investigations on U.H.F. Dielectric Aerials.—H. Aberdam. (*Électronique, Paris*, April 1953, No. 77, pp. 27–34.) The design and performance of various dielectric aerials are discussed and compared, including an omnidirectional type described by Ducot (*La Recherche Aéronautique*, May/June 1949) which takes the form of a thin disk with central aperture. Compared with a biconical aerial this has the advantage of reduced size. It is particularly useful for frequencies of the order of 10 kMc/s, whereas the type described by Mueller (1205 of 1952) is preferred for use at 25 kMc/s.

621.396.676 : 621.396.933 **2897**

F.M. Altimeter Slot Radiators.—L. E. Raburn. (*Tele-Tech*, April 1953, Vol. 12, No. 4, pp. 73–75..180.) The use of single-slot and double-slot radiators, in conjunction with a 440-Mc/s f.m. altimeter (1863 of 1946), is described and the best location and type of radiator for conventional aircraft are indicated. An outline of the method of measurement of feed-through between the transmitter and receiver is given, and radiation patterns of dipole single-slot and double-slot radiators are shown.

621.396.677 **2898**

Free-Space Radiation Impedance of Rhombic Antenna.—J. G. Chaney. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 536–540.) The expression for the driving point impedance of a generalized electric circuit is partially integrated, and the physical significance of certain terms is discussed in connection with their application to aerial problems. Assuming an unattenuated travelling wave as a first approximation to the current along a terminated rhombic or V aerial, a formula is derived for the free-space radiation impedance of each. The resistive component of the impedance of the rhombic aerial is in agreement with the radiation resistance given by Lewin.

621.396.677 **2899**

Radiation Field of a Conical Helix.—J. S. Chatterjee. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 550–559.) If a conical helix is used instead of a cylindrical helix, the axial mode of radiation can be maintained over a much wider band of frequencies. The radiation pattern of a conical helix (base diameter 60 cm, linear taper to 20 cm at top in 10 turns, height 112 cm, ground plane 100 cm in diameter) was studied experimentally. The axial mode of radiation was maintained from 150 to 450 Mc/s. The bandwidth can be increased by increasing the number of turns. Radiation diagrams are given for various feed arrangements.

621.396.677 **2900**

U.H.F.-TV Radiators using Slot Arrays.—R. J. Stegen. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 152–155.) Slot arrays consisting of longitudinal slots offset from the centre line of the broad face of a waveguide are considered. Graphs, scale-model results and typical azimuth patterns indicate methods of design and range of application. The advantages include simplicity of feed arrangements, low cost, low transmission-line power loss, improved aperture efficiency and ease in obtaining shaped vertical-plane patterns.

621.396.677 **2901**

Experimental Investigation and Calculation of End-fire Aerial Arrays.—D. M. Visokovski. (*C. R. Acad. Sci. U.R.S.S.*, 1st March 1953, Vol. 89, No. 1, pp. 41–44. In Russian.) A short account of an investigation of the dependence of the radiation characteristics on the number of radiator dipoles and on the length of the director elements.

621.396.677 : 537.226 **2902**

The Properties of Artificial Dielectrics comprising Arrays of Conducting Elements.—M. M. Z. Kharadly & W. Jackson. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 199–212.) Analysis is made of the effective permittivity of rectangular arrays of perfectly conducting elements of simple geometric shapes, such as circular-section rods, thin strips, spheres and disks, and results of experimental investigations at 1 kc/s are given. Representation of the elements as simple dipoles is inadequate at high concentration; higher-order multipole interaction between them must be taken into account. In cases where this interaction can be evaluated, experimental and theoretical permittivity values are in close agreement over the whole range of concentrations investigated. The same method was applied in calculations of the dielectric properties of a cubical array of imperfectly conducting spheres. Agreement with experiment is good except where the conducting elements occupy more than about 20% of the total volume.

621.396.677 : 621.392.26 **2903**

Waveguide Arrays with Symmetrical Conductance Functions.—H. D. Griffiths. (*Canad. J. Phys.*, May 1953, Vol. 31, No. 4, pp. 520–528.) Formulae for the efficiency and for the radiation pattern of nonresonant slotted-waveguide arrays are derived and methods are described for calculating the patterns from the asymmetrical near-field distributions produced by symmetrical arrays. Experimental results for two arrays with squint angles of 1·75° and 5·25° are shown graphically and are in satisfactory agreement with theory.

621.396.677.012.12 : 621.317.755 **2904**

Aerial Radiation Patterns. Apparatus for Cathode-Ray-Tube Display.—Ling. (See 3069.)

621.396.677.5 **2905**

Radiation Resistance of a Small Circular Loop in the presence of a Conducting Ground.—J. R. Wait. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 646–649.) The total power flow from an oscillating vertical magnetic dipole above a flat homogeneous conducting earth is evaluated. The result is used to derive an expression for the radiation resistance of a small horizontal wire loop. For the case of earth with finite conductivity, the radiation resistance is very great when the height of the loop is a small fraction of λ.

621.396.677.6 **2906**

Null Characteristics of the Rotating Adcock Antenna System.—J. G. Holbrook. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 530–532.) Mathematical analysis indicates that the optimum spacing of the dipoles for the most clearly defined nonmultiple nulls approaches the value of the shortest wavelength for which the system is designed, whereas the usual practice is to adopt a $\lambda/2$ spacing.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21 : 517.75 **2907**

Geometrical Transformation of Impedance Diagrams.—H. Briner & W. Graffunder. (*Helv. phys. Acta*, 15th Sept. 1952, Vol. 25, No. 5, pp. 487–488. In German.) Note on the use of graphical methods and of Riemann's number sphere in impedance calculations.

621.3.018.41 **2908**

The Concept of Instantaneous Frequency.—P. Poincelot. (*Onde élect.*, April 1953, Vol. 33, No. 313, pp. 214–216.) Discussion shows that this concept can properly be applied to the cases of demodulation by a linear discriminator and f.m. of a self-oscillator by variation of a parameter of the circuit, but its use in other cases, such as calculation of the harmonic distortion of a simple tuned circuit, leads to erroneous results. This is demonstrated by rigorous calculation of the distortion introduced by an *LGC* circuit in the anode circuit of a pentode amplifying stage.

621.314.3† **2909**

Theory of the Magnetic Amplifier with Free Magnetization.—L. Kühn. (*Frequenz*, April 1953, Vol. 7, No. 4, pp. 89–94.)

621.314.3† **2910**

Comparison of Methods of Analysis of Magnetic Amplifiers.—L. A. Finzi & G. F. Pittman, Jr. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 144–157.)

621.314.3† **2911**

Types of Magnetic Amplifiers — Survey.—J. G. Miles. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 229–238.) A scheme for classifying magnetic amplifiers under three main headings, with subdivisions by number and letter, is proposed and discussed.

621.314.3† **2912**

The Figure of Merit of Magnetic Amplifiers.—J. T. Carleton & W. F. Horton. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 239–244.)

621.316.54 : 512 **2913**

Sketch for an Algebra of Switchable Networks.—J. Shekel. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 913–921.) A combination of complex-number and Boolean algebra is outlined which can be applied in the analysis and synthesis of networks containing switches.

621.318.572 : 621.314.7 **2914**

A Method of Designing Transistor Trigger Circuits.—F. C. Williams & G. B. B. Chaplin. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 228–244. Discussion, pp. 245–248.) A fairly general technique was developed after the war for designing circuits that would operate with valves having characteristics which varied within fairly wide tolerances, without component adjustments being necessary when valves were changed. A similar technique is outlined for point-contact transistors; this is based on valve-circuit technique, and particularly on the analogy between the characteristics of transistors and pentodes. Basic pulse circuits considered include 2-state devices, timing circuits, counters and relaxation oscillators. Their application in digital computers is reviewed in detail.

621.319.4 **2915**

Nonlinear Condensers.—B. G. Lewis. (*Radio & Telev. News, Radio-Electronic Engng Section*, May 1953, Vol. 49, No. 5, pp. 3–5 . . 25.) The properties and applications of high-permittivity materials containing TiO_2 are discussed, with particular reference to the dependence of the dielectric constant on temperature and on the voltage applied. Applications of the variation of capacitance with the d.c. bias voltage in a modulator and in a capacitor-type amplifier are described.

621.319.4 : 621.315.612.4 **2916**

The Nonlinearity of Titanate Capacitors.—M. Kornetzki. (*Frequenz*, May 1953, Vol. 7, No. 5, pp. 121–127.) Ferroelectric and ferromagnetic materials are compared and expressions for the 'specific nonlinearity' are derived for both. These expressions give the hysteresis loss/cm^3 referred to unity circular frequency and a reactive power of 1 W/cm^3. The specific nonlinearities of (*a*) $(Ba,Sr)TiO_3$ material with a dielectric constant of 2 000 and (*b*) (Mn,Zn) ferrite with an initial permeability of 1 100 are of the same order of magnitude.

621.392.029.63/.64 **2917**

The Use of Short Transmission Lines in Ultra-High-Frequency Techniques.—P. Rohan. (*Philips tech. Commun., Aust.*, 1953, No. 1, pp. 3–12.) Lines of length $<\lambda$ are considered; their uses as inductances, capacitances and resonant circuits are reviewed. Numerical examples are given. Applications as filters and impedance transformers are described.

621.392.21.011.21 **2918**

New H.F. Proximity-Effect Formula.—A. C. Sim. (*Wireless Engr*, Aug. 1953, Vol. 30, No. 8, pp. 204–207.) Arnold's results (1017 of 1947) are somewhat complex. A new asymptotic formula was therefore developed which is valid whenever the proximity effect is so great that the resistance of the pair of parallel conductors is more than doubled. A combination of the Maxwell-equation and integral-equation approaches is used in deriving the new impedance formula, which is rigorous and very simple.

621.392.5 + 621.396.615 **2919**

The Equivalent Q of RC Networks.—D. A. H. Brown. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 294–298.) The effective Q factor of a RC phase-shift network is defined as one half the rate of change of phase with frequency at the operating point. For networks commonly used as frequency-determining elements the effective Q value is about unity, but a capacitance type of bridged-T network may have a Q value of 10.

621.392.5 **2920**

Design of Low-Frequency Constant-Time-Delay Lines.—C. M. Wallis. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 135–140.) Full paper. See 1603 of June.

621.392.5 : 534.321.9 **2921**

Ultrasonic Delay Lines.—(*Electronics*, July 1953, Vol. 26, No. 7, pp. 210..216.) Survey of the characteristics and operation of delay lines using fused-quartz or glass rods with quartz-crystal transducers.

621.392.5.029.64 **2922**

A New Non-reciprocal Waveguide Medium using Ferrites.—E. H. Turner. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, p. 937.) A narrow ferrite strip is attached along the wall of a circular-section waveguide and subjected to a transverse magnetic field. A difference in phase constant exists for the two directions of transmission along the guide. For suitable ferrite dimensions, a nonreciprocal rotation of the plane of polarization of an e.m. wave takes place. A more complete theoretical explanation is to be published. The case of an axial ferrite rod with an applied axial field has been considered by Hogan (1233 of 1952).

621.392.52 **2923**

Semifinite-Terminated Electric Wave Filters.—S. S. L. Chang. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 149–156.) Intended for use either as channel filters or to work between a finite resistance at one end and a very high or very low impedance at the other. Two design methods are presented, one based on the equivalent image-parameter filter, the other on Norton and Darlington's insertion-loss procedure. The minimum ratio of terminating resistances lies between 5 and 50 for most applications.

621.392.52 **2924**

A New Approach to Optimum Filtering.—E. W. Pike. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 407–418.) In filter problems, if enough is known about the signal and the noise to describe them accurately by their power spectra, conventional methods based on Wiener's original method will provide the most effective filters. An alternative procedure for computing optimum filters, based on description of the signal in terms of a partial sum of orthogonal functions, is presented. This will specify good filters with far less labour than is necessary in the Wiener method, and it has a wider range of practical application. A table of summation-orthogonal analogues of the Laguerre functions, which are useful in this procedure, is appended.

621.392.52 **2925**

Synthesis of a Dynamically Variable Electronic Filter.—J. G. Truxal & J. N. Warfield. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 419–426.) The principles of network synthesis are applied to the design of a variable-bandwidth filter with an upper cut-off frequency which can be varied dynamically over the frequency range 1–10 kc/s. The bandwidth variation is automatically controlled by voltages applied to the control grids of pentodes, slight changes occurring in selectivity and mid-band gain.

621.396.6 **2926**

Printed and Potted Electronic Circuits.—G. W. A. Dummer & D. L. Johnston. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 177–186. Discussion, pp. 187–191.) Survey and assessment of production techniques, and description of some applications.

621.396.611 (083.72) (47) **2927**

Russian Circuit Notations.—G. F. Schultz. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, p. 936.) The methods adopted for indicating the values of resistors and capacitors on Russian circuit diagrams are explained.

621.396.611.1 **2928**

A Relation between Susceptance Slope and Selectivity for Oscillator Design.—W. A. Edson & R. D. Teasdale. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 427–434.) Two quantities, *(a)* the slope of the susceptance curve, computed at an antiresonance frequency, *(b)* the dimensionless selectivity factor Q, are significant in specifying the performance of oscillatory circuits. The relation between the two quantities for practical resonant circuits is discussed and general conclusions are drawn relating to resonant-circuit design. These conclusions are applied to the comparison of the Clapp and Colpitts types of oscillator and to the study of the equivalent circuit of a quartz crystal operating near resonance.

621.396.611.1 **2929**

Subharmonics in a Series Nonlinear Circuit as Influenced by Initial Capacitor Charge.—W. J. McKune & M. F. Brust. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 200–205.)

621.396.611.1 **2930**

Note on 'The Response of an LCR Circuit'.—S. V. Soanes. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, p. 935.) Comment on 3041 of 1952 (Marique).

621.396.611.3 : 621.392.26 **2931**

The Transvar Directional Coupler.—K. Tomiyasu & S. B. Cohn. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 922–926.) The coupling element consists of n closely spaced identical apertures, separated by $(n-1)$ grid wires, in the common narrow wall between two waveguides. By changing the relative longitudinal position of the two waveguides, the number of exposed apertures, and hence the effective coupling length, can be changed. Power transfer is variable between 0 and 100 %. Maximum power transfer is available at one frequency only, and hence bandwidth depends on the tolerable decrease in power transfer. Two models, with $n = 22$, and differing only in grid-wire diameter, have maximum power transfer at about 9 and 10 kMc/s respectively. Agreement between theory and performance is close.

621.396.611.4 **2932**

External Influences on the Circuit Properties of U.H.F. Resonators.—G. Megla. (*NachrTech.*, Nov. 1952, Vol. 2, No. 11, pp. 332–336.) The Q of a cavity resonator depends not only on its physical dimensions and the material of which it is made, but also on the surface finish. The influence of the material and its finish on the fundamental damping and resonance resistance is investigated for resonators of various types. An examination is made of the degree of roughness permissible, with reference to standard specifications on surface finishes. Effects due to variations of atmospheric temperature, pressure and humidity, and methods of eliminating these effects, are also discussed.

621.396.611.4 **2933**

Interaction of Modes in a Microwave Cavity Resonator.—S. K. Chatterjee. (*J. Indian Inst. Sci.*, Section B, April 1953, Vol. 35, No. 2, pp. 59–69.) The minimum energy differences between the desired and undesired modes in a 9-kMc/s cylindrical resonator are calculated by considering the e.m. field inside the resonator as due to an infinite number of harmonic oscillators. Expressions for the mutual energy and the coefficient of coupling between the companion modes TE_{01} and TM_{11} are derived. The significance of the total energy of a microwave cavity resonator is discussed.

621.396.615 **2934**

Power Spectrum of a Nonlinear Oscillator perturbed by Noise.—A. Blaquière. (*Ann. Radioélect.*, April 1953, Vol. 8, No. 32, pp. 153–179.) Analysis of the effect of background noise in producing amplitude fluctuations. If the steady state of an oscillator is disturbed, recovery takes place exponentially with a time constant τ which is the principal factor affecting the noise-power spectrum. Simple relations between τ and the circuit parameters are derived which show that for minimum noise power oscillations of large amplitude should be used, without introducing too many harmonics, and the phase of the feedback should vary only slowly with frequency. See also 968 of April and back references.

621.396.615 **2935**

Two *RC* Oscillator Circuits.—H. L. Armstrong. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 234–238.) Cathode-coupled and transitron types of *RC* oscillator are described which operate at frequencies of about 400 kc/s and 20 kc/s respectively.

621.396.615 **2936**

Low-Power Blocking Oscillators.—J. R. Clark. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 231–235.) Information of use in final adjustment of blocking oscillators, particularly as regards requirements for short rise-times and narrow pulses, is reviewed. A linearized equivalent circuit is used to derive an explicit expression for pulse duration which gives results in agreement with experiment, particularly when the grid coupling capacitor is small.

621.396.615 : 621-526 **2937**

Use of Servo Techniques in the Design of Amplitude-Stabilized Oscillators.—A. W. Dickson. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 166–176.) Expressions for the oscillation amplitude in terms of the stabilizing-signal voltage are derived and the amplitude-stabilizing system is discussed as a negative-feedback loop. Experimental and theoretical results for a high-*Q* oscillator are compared.

621.396.615 : 621.314.7 **2938**

F.M. Transistor Oscillator.—H. H. Wieder & N. Cass. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 198–202.) A titanate capacitor is coupled to the resonant-circuit inductor of the transistor oscillator, the transformation ratio being chosen at will. Varying the potential applied to this capacitor alters its effective capacitance and hence the resonance frequency. A polarization potential is applied from a Zamboni pile to keep the capacitance variations within the linear range. A 2·5-Mc/s oscillator constructed used a CK722 junction transistor and had a power output of 1 mW when supplied from a 30-V battery. The titanate capacitor consisted of two series-connected 100-pF units. A 3·5-Mc/s oscillator using a Type-1698 point-contact transistor gave similar results. Some form of a.f.c. is desirable in view of the temperature variations of the components.

621.396.615 : 621.316.722.1 **2939**

An Oscillator with Constant Output Voltage.—L. Ensing & H. J. J. van Eyndhoven. (*Philips tech. Rev.*, April 1953, Vol. 14, No. 10, pp. 304–312.) The theory and design of an output-voltage stabilizer circuit are given. By a suitable choice of the values of two resistors and of the constant reference voltage, the oscillator output voltage is made virtually independent of valve characteristics and circuit impedance. An application to the calibration of valve voltmeters is described.

621.396.615 : 621.316.729 **2940**

Mechanism of the Synchronization of *LC* Oscillators.—J. van Slooten. (*Philips tech. Rev.*, April 1953, Vol. 14, No. 10, pp. 292–298.) The pulling effect on the phase and frequency of an *LC* oscillator due to interference is considered. The phase change is calculated for the case of a single interfering pulse, a series of equidistant pulses, and a sinusoidal voltage. Interfering pulses occurring at an oscillator voltage peak affect the amplitude, those occurring at zero voltage cause a change in phase only. These phase changes are permanent. A differential equation, relating oscillator-voltage phase deviation, time, peak oscillator voltage and the magnitude of the interference current, is solved and the conditions for synchronism are obtained.

621.396.615 : 621.396.611.21 **2941**

Some Aspects of Mixer-Crystal Performance.—P. D. Strum. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 875–889.) A method of calculating conversion loss and conductances of a crystal mixer is presented, and their dependence on terminations at the image frequency and at the sum frequency is considered. Experimental data from various sources support the calculations. Graphs showing (*a*) excess crystal noise versus output frequency, and (*b*) mixer noise temperature versus crystal noise temperature for varying conversion loss, are derived and applied, together with the conversion-loss data, to practical calculations of optimum mixer configuration and optimum i.f. Two examples are worked out. Experimental confirmation of results was obtained.

621.396.615.142.2 : 621.316.726.078 **2942**

Automatic Frequency Control of High-Power Klystrons.—R. F. Denton, T. A. Wilson & A. R. Margolin. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 56–63.) A method of control is described in which a cavity resonator and two microwave crystals are used as a standing-wave discriminator in a system for automatic frequency search and lock for a pulsed klystron oscillator. Frequency stability to within 1 part in 10^4 is achieved.

621.396.615.142.2 : 621.396.822 **2943**

Length of Coherent Microwaves Generated by an Electronic Oscillator.—K. Shimoda. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 131–132.) Investigation of the noise associated with the generation of oscillations, of frequency 8 kMc/s, by a reflex klystron showed a bandwidth of about 3 kc/s. This is attributed to fluctuation of the electron beam.

621.396.615.17 **2944**

The Square-Wave Generator.—E. Piepgras. (*Funk u. Ton*, May 1953, Vol. 7, No. 5, pp. 230–240.) The circuit of a multivibrator type of generator with stabilized input is shown and a complete analysis based on the characteristics of the double triode, Type ECC40, is given.

621.396.615.17 **2945**

A Reliable Locked-Oscillator Pulse Timer.—P. G. Sulzer. (*Tele-Tech*, April 1953, Vol. 12, No. 4, pp. 68–69, 173.) Description, with complete circuit diagram and method of adjustment, of a highly stable locked-oscillator pulse timer providing pulses with recurrence frequency 25/sec from a 100-kc/s input. The frequency divider has previously been described (944 of 1952). The unit has an overall phase stability to within $\sim 0 \cdot 1$ µs per volt of line-voltage change.

621.396.615.17 **2946**

An Analogue Reciprocal-Function Unit for Use with Pulsed Signals.—P. A. V. Thomas. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 302–304.) A monostable multivibrator operates so that the width of the pulse generated is approximately inversely proportional to the peak voltage of an input pulse. The pulse-width variation is converted to a corresponding amplitude variation of sawtooth output pulses by means of a constant-velocity timebase circuit.

621.396.645 : 621.313.2 **2947**

Calculations for a Power Amplifier for [driving] a D.C. Motor.—J. Zakheim. (*Onde élect.*, April 1953, Vol. 33, No. 313, pp. 235–239.) Analysis of a symmetrical class-A amplifier supplying energy to a motor connected in the cathode circuit of the output stage. See also 687 of March.

621.396.645.029.3 **2948**

High-Power Audio Amplifiers.—L. F. Deise & H. J. Morrison. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 81–88.) Recent applications for a.f. power up to 10 kW include vibration work, servomechanism studies, pulsed service, and variable-frequency power sources, all of which require response, distortion and noise-level characteristics similar to those for broadcasting service. In high-power amplifiers with class-B push-pull output, the output transformer is the largest and most expensive item in the equipment and may weigh some hundreds of pounds. A new equivalent circuit is used in the design of such transformers. Harmonic distortion caused by the transformer is analysed and methods of improving the performance of class-B power amplifiers are described. A few details are given of a 5/10-kW amplifier with a response curve flat to within 1 db from 10 c/s to 30 kc/s.

621.396.645.029.62/.63 **2949**

General Design Considerations of a Cavity-Type Power Amplifier.—W. S. Elliott. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 64–72.) Details are given of the construction of a single-stage amplifier using a Type-2C39A triode valve. Constant-current curves for class-C grounded-grid operation were used to determine the operating parameters of the triode. A 3-stage amplifier using Type-2C39A and Type-5893 triodes is described, with performance details.

621.396.645.3.029.424 **2950**

A Simple Low-Frequency Amplifier.—J. R. Beattie & G. K. T. Conn. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 299–301.) Description of a tuned-amplifier circuit for the range 1–3 c/s, incorporating a damped phase-shift oscillator with independent control of resonance frequency and Q-factor.

621.396.822 : 621.3.014.1.025 **2951**

Differential Law for the Intensities of Electrical Fluctuations, and Influence of Skin Effect.—M. A. Leontovich & S. M. Rytov. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 3, pp. 246–252.) An investigation is made of the form of the correlation function of the external field which expresses the thermal electrical fluctuations in a conductor, and of the 'differential law' on which Nyquist's integral formula (1) is based.

621.396.822 : [621.316.86 + 621.314.63 **2952**

On the Temperature Dependence of Contact Noise.—Y. Inuishi. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, p. 128.) Results of noise measurements on Se and Cu_2O rectifiers and on carbon resistors over different temperature ranges between −40° and 130°C are shown graphically.

GENERAL PHYSICS

534.01 **2953**

Subharmonic Oscillations in Nonlinear Systems.—C. Hayashi. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 521–529.) The order of the subharmonics in systems with nonlinear restoring force is closely related to the form of the nonlinearity. A detailed analysis is presented for the subharmonic oscillation of order $\frac{1}{3}$, for the cases in which the nonlinearity is expressed by a cubic or a quintic function. Stability is discussed by means of the criterion previously given (2361 of August).

535.12 : 535.31 **2954**

Step-By-Step Transition from Wave Optics to Ray Optics in Inhomogeneous Absorbing Media: Part 1 — Equations for Wave Normal, Refractive Index and Polarization.—K. Suchy. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 113–130.) The transition is made by successive approximations. Each approximation is considered from the physical point of view and the physical significance of the equations derived is considered.

535.37 **2955**

Luminescent Layers and Films with Polyvinyl Alcohol as Protective Colloid and Binder.—F. Eckart. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 169–174.)

535.37 : 537.228 : 621.32 **2956**

Electroluminescence — Electrical and Optical Properties.—J. F. Waymouth, C. W. Jerome & W. C. Gungle. (*Sylvania Technologist*, July 1952, Vol. 5, No. 3, pp. 54–59.) Electro-optical measurements are reported on electroluminescent lamps operating on the principle described by Payne et al. (1341 of 1951), and fundamental properties of electroluminescent phosphors are studied.

537.12 **2957**

Method for Measurement of the Charge of Fast Electrons.—R. Fleischmann & R. Kollath. (*Z. Phys.*, 17th April 1953, Vol. 134, No. 5, pp. 526–529.) A method is outlined for determining whether the variation of m/e with the electron velocity depends entirely on m.

537.12 **2958**

Measurements of Charge on Electrons in Motion.—R. Kollath & D. Menzel. (*Z. Phys.*, 17th April 1953, Vol. 134, No. 5, pp. 530–539.) No variation in the electron charge was detected in the range 2·5–30 keV. The method used was that outlined by Fleischmann & Kollath (2957 above).

537.525.5 : 621.327.43 : 538.561 **2959**

Oscillatory Phenomena of Arc in Hot-Cathode Discharge Tube: Part 1 — Experimental (Characteristics of Oscillation).—H. Yosimoto. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 59–68.)

537.525.5 : 621.327.43 : 538.561 **2960**

Oscillatory Phenomena of Arc in Hot-Cathode Discharge Tube: Part 2 — Theoretical (Mechanism of Oscillation).—H. Yosimoto. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 69–76.)

537.533 : 537.525.92 **2961**

Space-Charge Spread of Reflected Electron Beams studied by a Photographic Method.—J. T. Wallmark. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 590–597.)

537.533.7 : 537.241 : 530.145.61 **2962**

The Self-Charge Spreading of a Focused Electron Beam.—H. Grümm. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 131–144.) Simple formulae, derived from wave-mechanical considerations, are obtained for the upper limits of the self-charge effects, i.e., spreading of beam, and displacements of image plane and Fraunhofer plane.

537.582 **2963**

Energy Distribution of Thermo-electrons in Electron-Beam Devices.—H. Boersch. (*Naturwissenschaften*, May 1953, Vol. 40, No. 9, pp. 267–268.) Experiments are described which show that the energy distribution of the thermo-electrons in oscillograph tubes, electron microscopes, etc., is not of the Maxwell type. The deviations from the Maxwell distribution, particularly in the low-energy range, are attributed to reciprocal dynamic effects.

538.114 **2964**

On the Theory of the Ising Model of Ferromagnetism.—G. F. Newell & E. W. Montroll. (*Rev. mod. Phys.*, April 1953, Vol. 25, No. 2, pp. 353–389.)

538.122 + 537.212 **2965**

The Two-dimensional Magnetic or Electric Field above and below an Infinite Corrugated Sheet.—N. H. Langton & N. Davy. (*Brit. J. appl. Phys.*, May 1953, Vol. 4, No. 5, pp. 134–137.)

538.311 **2966**

A Conducting Permeable Sphere in the presence of a Coil carrying an Oscillating Current.—J. R. Wait. (*Canad. J. Phys.*, May 1953, Vol. 31, No. 4, pp. 670–678.) The secondary magnetic fields due to the sphere are calculated for frequencies <500 c/s.

538.311 **2967**

Current Element near the Edge of a Conducting Half-Plane.—R. F. Harrington. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 547–550.) "The two-dimensional problem of the Maxwellian field from a line source of current adjacent to a conducting half-plane is treated by a transform method of solution. Expressions in the form of contour integrals are given for the current induced in the conductor and for the radiated field."

538.566 : 535.3] + 534.24 **2968**

The Scattering of Waves at an Uneven Surface.—L. M. Brekhovskikh. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 3, pp. 275–304.) An approximate method is proposed for calculating the field of acoustic and electromagnetic waves scattered by an uneven surface, the irregularities being large in comparison with the wavelength. Particular cases considered include that of sinusoidal corrugations.

538.566 : 535.3] + 534.24 **2969**

Scattering of Waves by a Statistically Irregular Surface.—M. A. Isakovich. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 3, pp. 305–314.) A method is proposed for calculating the field of plane waves scattered by a perfectly reflecting, statistically irregular surface, the irregularities being large in comparison with the wavelength. The method can be used for acoustic as well as electromagnetic waves.

538.566 : 537.562 **2970**

Quenching of Afterglow in Gaseous Discharge Plasmas by Low-Power Microwaves.—L. Goldstein, J. M. Anderson & G. L. Clark. (*Phys. Rev.*, 1st May 1953, Vol. 90, No. 3, pp. 486–487.) Continuation of work noted in 2637 of September. Results are reported for neon of the effect on the plasma decay of applying microwave pulses at various time intervals after the production of the plasma. The state of decay was determined by observing the afterglow by means of photocells, and displaying the amplified photocurrent on an oscilloscope. Decrease of light intensity observed during the application of the pulse is most probably due to reduction of the recombination rate. Differences in the traces observed at pressures of 20 mm Hg and 1 mm Hg are discussed. The effect may be useful for detecting e.m. energy.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72 : 621.396.822 **2971**

On Some Features of Noise Storms.—T. Hatanaka & F. Moriyama. (*Rep. Ionosphere Res. Japan*, June 1952, Vol. 6, No. 2, pp. 99–109.) Discussion of solar r.f. emission of abnormally high intensity on metre wavelengths. The radiation is circularly polarized, may last for several hours, and is often associated with sudden bursts of nonpolarized radiation. The association of such noise storms with large and magnetically active sunspots is considered.

550.384 **2972**

Time Variation of the Earth's Magnetic Field at the Time of Bay Disturbance.—Y. Kato & J. Ossaka. (*Rep. Ionosphere Res. Japan*, March 1952, Vol. 6, No. 1, pp. 37–41.)

550.385 **2973**

Distribution of S.C.* of Magnetic Storms.—T. Nagata. (*Rep. Ionosphere Res. Japan*, March 1952, Vol. 6, No. 1, pp. 13–30.) The abbreviation S.C.* is used to denote the reverse impulse preceding a sudden commencement. The distribution with respect to local time and latitude was examined from records obtained at five U.S. stations during a period of 3½ years, and also from world-wide records of individual geomagnetic storms. The reverse impulses appear chiefly in the afternoon, with a maximum at about 1600 local time. The pulse magnitude, as well as its ratio to the succeeding main pulse, increases markedly with geomagnetic latitude.

550.385 **2974**

The Recurrence Tendency of Magnetic Storms.—M. Ota. (*Rep. Ionosphere Res. Japan*, Dec. 1952, Vol. 6, No. 4, pp. 212–214.) Data relative to 196 storms observed at Aso Magnetic Observatory during the period September 1949–August 1952 are analysed. 23 storms showed no recurrence tendency. The rest are divided into 33 groups, each containing several storms recurring at intervals of about 27 days.

550.385 : 523.755 **2975**

Correlation of Magnetic M-Storms with the Monochromatic Corona.—R. Müller. (*Observatory*, April 1953, Vol. 73, No. 873, pp. 75–77.) Results obtained at Wendelstein show that a maximum of geomagnetic activity is observed 3·7 days before the central-meridian passage of a region of high coronal activity, or 3·1 days after the appearance of bright coronal emission at the east solar limb. Similar characteristics are found for the minimum epochs. The relation apparently holds good only when no visible or significant sunspot phenomena exist.

550.385 + 551.594.5] : 621.396.9 **2976**

Radio Echoes observed during Aurorae and Terrestrial Magnetic Storms, using 35- and 74-Mc/s Waves Simultaneously.—L. Harang & B. Landmark. (*Nature, Lond.*, 6th June 1953, Vol. 171, No. 4362, pp. 1017–1018.) A report of observations at Tromsö, near the auroral zone, and at Kjeller, near Oslo. The two pulse transmitters at each station had common pulse modulators, so that the shape and energy of the pulses could be made identical. The peak pulse power was 25 kW and repetition frequency 50/sec. Echoes on 35 Mc/s appeared regularly at Tromsö during auroral displays and in a number of cases also on 74 Mc/s. Study of the results led to the conclusion that the echoes were not due to direct scatter from the ionosphere, but could be explained by reflection of backwards scatter from land or sea via the E_s layer, the ionization of which increases greatly during aurorae and geomagnetic storms. In all cases, the echoes on 35 Mc/s showed a smaller range than those on 74 Mc/s, the difference in range usually exceeding 100 km. Differences between the echo patterns on the two frequencies are noted. A full account of the observations will be given in *J. atmos. terr. Phys.*

550.385 (98) **2977**

Constitution of Polar Magnetic Storms.—T. Nagata & N. Fukushima. (*Rep. Ionosphere Res. Japan*, June 1952, Vol. 6, No. 2, pp. 85–97.) Analysis of data obtained in the International Polar Year indicates that polar magnetic storms consist of a number of elementary disturbances which occur successively. The equivalent current system corresponding to these disturbances can be represented approximately by that due to an electric dipole located in the auroral zone.

550.385(98) **2978**

Constitution of Polar Magnetic Storms.—N. Fukushima. (*Rep. Ionosphere Res. Japan*, Dec. 1952, Vol. 6, No. 4, pp. 185–193.) An examination of world-wide data relative to the polar magnetic storm on 30th April 1933 confirms the conclusions noted in 2977 above.

551.510.534 : 551.524.7 **2979**

Annual Variation of Reduced Density and Mean Temperature of Atmospheric Ozone in Afghanistan.—A. Khalek. (*C. R. Acad. Sci., Paris*, 22nd June 1953, Vol. 236, No. 25, pp. 2424–2426.)

551.510.535 + 523.71 **2980**

Abbreviated Codes of European Ursigrams: Part 2.—(*U.R.S.I. Inform. Bull.*, March/April 1953, No. 78, pp. 16–37.) The codes here given are: (*a*) Data on Monochromatic Intensity of the Solar Corona, 'CORON' code; (*b*) Data on E_s Critical Frequency (fE_s), 'ESFRE' code; (*c*) Data on F_2 Critical Frequency (f_0F_2), 'FODEU' code for f_0F_2 hourly values, 'SYMBO' code for descriptive symbols for the values given; (*d*) Terrestrial Magnetism, 'MAGNE' code; (*e*) Ionospheric-Disturbance Warning, 'PERTU' code; (*f*) Radio Solar-Emission Observatories, 'SOLER' code. Part 1: 2650 of September.

551.510.535 + 523.71 **2981**

European Ursigrams.—(*U.R.S.I. Inform. Bull.*, March/April 1953, No. 78, p. 16.) Correction to 2650 of September.

551.510.535 **2982**

Some Characteristics of Ionospheric Storms.—T. Obayashi. (*Rep. Ionosphere Res. Japan*, June 1952, Vol. 6, No. 2, pp. 79–84.) Analysis of data for the F_2 region. The frequency of occurrence of disturbances in this region is closely related to its seasonal height, being greater in summer than in winter.

551.510.535 **2983**

Classification of F_2-Layer Storms with respect to their World-Wide Distribution and Characteristics: Part 1.—H. Uyeda & Y. Arima. (*Rep. Ionosphere Res. Japan*, March 1952, Vol. 6, No. 1, pp. 1–12.) Analysis of available data shows that there is no correlation between F_2-layer storms and sunspot activity, but close correlation with geomagnetic storms. A classification of these storms with respect to their distribution over the geomagnetic latitudes is suggested, based on layer-height or critical-frequency characteristics.

551.510.535 **2984**

On World-Wide Distributions of f_0F_2.—Y. Aono. (*Rep. Ionosphere Res. Japan*, June 1952, Vol. 6, No. 2, pp. 69–78.) Hourly values of f_0F_2 observed at 40 stations, distributed all over the world, during the period 8th–11th May 1948 are analysed, contour lines being used to indicate places with common critical frequencies. Diagrams are also given showing (*a*) the distribution of the difference between f_0F_2 and its monthly mean value at 0700 and at 0800 on 9th May 1948, (*b*) the distribution of f_0F_2 for noon local time on the same date, (*c*) the latitude distribution of f_0F_2 for noon local time on longitudes 135°E and 75°W.

551.510.535 **2985**

On the Cause for Unnatural Distribution of Occurrence of fE_s and $f_{min}F$.—H. Uyeda, K. Miya & T. Kobayashi. (*Rep. Ionosphere Res. Japan*, Dec. 1952, Vol. 6, No. 4, pp. 179–183.) Observations carried out at Kokubunji indicate a correlation between receiver interference and the frequency of occurrence of fE_s and $f_{min}F$.

551.510.535 : 550.384.4 **2986**

A Relation between F_2-Layer Disturbance and Geomagnetic Condition.—N. Fukushima & T. Hayasi. (*Rep. Ionosphere Res. Japan*, Sept. 1952, Vol. 6, No. 3, pp. 133–136.) A correlation is found between the daily mean value of the electron density of the F_2 layer and the storminess of the geomagnetic field. The F_2-layer electron density appears to decrease in local summer and increase in local winter in the middle-latitude regions of both the northern and southern hemispheres.

551.510.535 : 550.385 **2987**

Characteristics of Ionospheric Disturbances during Severe Magnetic Storms.—K. Miya & N. Wakai. (*Rep. Ionosphere Res. Japan*, Sept. 1952, Vol. 6, No. 3, pp. 137–146.) Analysis of world-wide data from 40 ionosphere stations for the period 8th–11th May 1948, including two geomagnetic storms with sudden commencements (s.c.), shows that an ionospheric disturbance coincident with a s.c. in a region of high geomagnetic latitude spreads down to lower latitudes. In high latitudes the disturbances travel at a high velocity, which is about 300 km/hr in the vicinity of Japan. The intensity of the disturbance decreases with latitude. The connection between the time of occurrence of a s.c. and disturbance of radio communication is briefly discussed.

551.510.535 : 551.594 **2988**

Studies on Ionospheric Storms.—H. Uyeda. (*Rep. Ionosphere Res. Japan*, Dec. 1952, Vol. 6, No. 4, pp. 169–177.) A definite relation is found between the types of storm discussed by Appleton & Piggott (891 of 1950) and the seasons in which they occur. The development of ionospheric storms is discussed in some detail. The need for further world-wide ionospheric investigations, particularly in high latitudes and at intervals less than an hour, is stressed.

551.510.535 : 621.3.087.4 **2989**

Modern Ionosphere Sounding Equipment of the Netherlands P.T.T.—P. L. M. van Berkel. (*Tijdschr. ned. Radiogenoot.*, May 1953, Vol. 18, No. 2, pp. 149–165. In English.) Full description, with block diagram, of a 1-kW pulse equipment for the range 1–20 Mc/s. Pulse duration is adjustable from 50 to 100 μs; repetition frequency is 50/sec. A complete frequency sweep can be made in either 10 or 20 sec. Two c.r. tubes in parallel provide monitoring and photographic facilities.

551.510.535 : 621.396.11 **2990**

C.C.I.R.-U.R.S.I. Co-operation. Exchange of Information for Forecasts and Warnings.—(See 3092.)

551.510.535(98) : 537.56 : 550.385 **2991**

Anomalous Ionization in the Upper Atmosphere over the Auroral Zone during Magnetic Storms.—M. Sugiura, M. Tazima & T. Nagata. (*Rep. Ionosphere Res. Japan*, Sept. 1952, Vol. 6, No. 3, pp. 147–154.) The depth of penetration of and the rate of ionization by protons of extraterrestrial origin are estimated. An explanation is given of the rapid variations of the geomagnetic field in and near the auroral zone.

LOCATION AND AIDS TO NAVIGATION

621.396.9 **2992**

Magnetic Modulators for Pulsed Radar.—W. S. Melville. (*B.T.-H. Activ.*, Nov./Dec. 1952, Vol. 23, No. 6, pp. 169–173.) The use of magnetic materials with rectangular B/H characteristics for the generation of pulses is described. The principles of polarization and cascade discharge are outlined and the operation of a marine radar modulator is described. Constructional features of magnetic switches, termed pulsactors, are mentioned. Magnetic and electronic modulators are compared, the principal advantages of the former being low radiated interference, instantaneous operation and the facility with which high-power pulses can be generated.

621.396.9 : 551.508.11 **2993**

Pulse Observations of Wind by Radio 'Echosonde' without Responder.—J. Lugeon. (*C. R. Acad. Sci., Paris*, 22nd June 1953, Vol. 236, No. 25, pp. 2426–2428.) A description is given of a method in which the radiosonde oscillator, quiescent in the intervals between its pulse transmissions, is triggered by a pulse from the ground station. The distance between ground station and radiosonde is found by the usual radar method.

621.396.9 : 551.578.1 **2994**

Vertical Recording of Rain by Radar.—S. K. H. Forsgren & O. F. Perers. (*Chalmers tek. Högsk. Handl.*, 1951, No. 107, 19 pp.) Reprint. See 681 of 1952.

621.396.93 **2995**

Radio Rescue Beacon.—(*Wireless World*, Aug. 1953, Vol. 59, No. 8, pp. 381–382.) Short description of a small battery-operated transmitter, with a r.f. pulse power of about 16 W, designed for attachment to an airman's life-saving jacket. Known as 'Sarah' (Search And Rescue And Homing) the apparatus, when brought into action, continues to transmit omnidirectional pulse signals for about 20 hours. One version of the equipment also provides for two-way speech communication over a limited distance.

621.396.93 **2996**

Radio Navigational Aids.—A. van Weel. (*Tijdschr. ned. Radiogenoot.*, May 1953, Vol. 18, No. 2, pp. 129–148.) Survey of the basic principles and characteristics of (*a*) nondirectional beacons and radio compass, (*b*) radio range, (*c*) instrument landing system, (*d*) directional beacons, (*e*) distance-measuring equipment, (*f*) loran and decca systems.

621.396.932/.933].2 **2997**

Radio Direction Finding. Influence of Buried Conductors on Bearings.—F. Horner. (*Wireless Engr*, Aug. 1953, Vol. 30, No. 8, pp. 187–191.) "The currents induced, at low frequencies, in a buried cable in good contact with the ground may greatly exceed those in a similar cable insulated from the ground. These currents may lead to large errors in a loop direction finder, even when the length of the cable is a small fraction of the wavelength. The errors are likely to be small if the direction finder is near one end of the cable. Formulae are derived for the currents induced in a buried conductor and these lead to calculated errors in reasonable agreement with measured errors, at a frequency of 10 kc/s. The results show that the effect of burying a cable in soil of good conductivity is to increase errors at low frequencies and to reduce errors at high frequencies. The transition frequency depends on the length of the cable and is, for example, about 300 kc/s for a cable 200 metres long. Adock direction finders are less liable to errors due to cables, but there is some risk of errors if a cable is laid in close proximity to an aerial feeder."

621.396.933 **2998**

Continuous-Indicating Loran Navigator.—R. B. Williams, Jr. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 166–169.) A 12-valve unit is added to the standard Sperry Mark-2 indicator to synchronize it with the received signals, maintain correct pulse amplitude as shown on the c.r. tube screen and maintain pulse superposition in height and time. Details are given of the circuits provided for a.g.c., automatic amplitude-balance control, pulse-amplitude sampling with time-sharing relay, a.f.c., automatic time-difference control and pulse-position detection. The time-sharing relay is an e.m. servo-type switch. One unit gives continuous indication of one line of position of the aircraft. Two units synchronized on separate pairs of stations give continuous indication of exact position.

MATERIALS AND SUBSIDIARY TECHNIQUES

53.087.3 : 537.311.33 **2999**

A Micromanipulator for Electrical Investigations of Semiconducting Materials.—A. Kelen. (*Appl. sci. Res. B*, 1953, Vol. 3, No. 2, pp. 125–128.) The construction is described of a simple instrument for the accurate location of 1–5 point contacts.

531.788.7 **3000**

A Cylindrical Magnetron Ionization Gauge.—A. H. Beck & A. D. Brisbane. (*Vacuum*, April 1952, Vol. 2, No. 2, pp. 137–146.) The gauge is essentially a diode with an axial magnetic field, its cylindrical structure distinguishing it from the Penning-type gauge. Its working range is 10^{-4}–10^{-8} mm Hg and when used in a bridge type of leak detector its sensitivity is of the same order as that of the mass spectrometer.

531.788.7 **3001**

Degassable Penning Gauge.—A. Bobenrieth. (*Le Vide*, March 1953, Vol. 8, No. 44, pp. 1302–1304.) Description of a gauge whose electrodes can be degassed by the Joule effect, h.f. heating not being required.

535.215 + 537.323] : 546.36.863 **3002**

Some Experimental Studies of the Conductivity and Thermo-electromotive Force of Cs_3Sb Photo-Cathodes.—T. Sakata. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 125–126.) Experimental results are shown graphically. The thermo-e.m.f. ranges from 1·1 mV/deg at 268° to 0·3 mV/deg at 343° K. The resistivity is given by the formula $\rho = \rho_0 \exp(\epsilon/kT)$, where T is the absolute temperature and ϵ is about 0·2–0·3, *p*-type conductivity being indicated.

535.215.1 : 621.396.822 **3003**

Noise Measurements on Thin Lead-Sulphide Photoelectric Layers.—F. Eckart. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 166–168.) Electrical noise measurements were made on PbS layers treated with oxygen. The absolute value of noise input is proportional to the cell current for both thermal and photoelectric excitation. The dark current and the mean noise input are proportional to the e.s. field strength. The mean life of the charge carriers is estimated to lie between $1{\cdot}3 \times 10^{-3}$ and $3{\cdot}7 \times 10^{-5}$ sec.

535.215.2 **3004**

Photoelectric Emission and Energy Structure of Cs_3Sb.—H. Miyazawa. (*J. phys. Soc. Japan*, March/April 1953, Vol. 8, No. 2, pp. 169–175.) Spectral distributions of the emission and energy distributions of the photoelectrons were determined for Cs_3Sb at room temperature, 195°K and 90°K. A possible energy structure is discussed.

535.371 + 535.377 + 535.215.1] : 546.47-13 **3005**

Photodielectric Effect, Ferroelectricity and Thermoluminescence in Zinc Oxide under Ultraviolet Irradiation.—J. Roux. (*C. R. Acad. Sci., Paris*, 29th June 1953, Vol. 236, No. 26, pp. 2492–2494.) The sample of ZnO considered had a greenish fluorescence under ultraviolet irradiation, was thermoluminescent but not phosphorescent, and had a marked photodielectric effect (permittivity variation) affected by the intensity of an applied e.s. field and also by infrared light during and after the excitation. The material is ferroelectric under excitation at room temperature.

535.377 **3006**

The Characteristics of a Class of Sulphate-Base Phosphors.—M. A. Konstantinova-Shlezinger, N. A. Gorbacheva & E. I. Panasyuk. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 5, pp. 588–592.) Attenuation and thermal luminescence curves are plotted for $PbSO_4$-Sm with different fluxes, and also thermal luminescence curves for $CdSO_4$ activated with Mn, Pb and Mn-Pb and for $PbSO_4$ with the double activator Sm-Ce.

537.226 **3007**

The Properties of Very Small Ferroelectric Particles.—M. Anliker, W. Känzig & M. Peter. (*Helv. phys. Acta*, 15th Sept. 1952, Vol. 25, No. 5, pp. 474–475. In German.) Preliminary report of investigations of anomalies in the ferroelectric state about the Curie point.

537.226 : 537.533.8 **3008**

The Secondary Emission from Dielectrics under Single-Impulse Conditions.—A. R. Shul'man & V. L. Makedonski. (*Zh. tekh. Fiz.*, 1952, Vol. 22, No. 10, pp. 1540–1542.) The coefficient of secondary emission σ varies with time owing to changes taking place in the surface layer of the target. To eliminate these effects in measurements of σ, single pulses may be used. A circuit was developed by the aid of which it is possible to measure the σ of dielectrics with constant primary current, with periodic impulses and with single impulses. The main requirements of this circuit are discussed and a block diagram is given.

537.226 : 546.431.824-31 **3009**

The Dielectric Properties and Optical Anomalies of $BaTiO_3$ Single Crystals.—I. N. Belyaev, N. S. Novosil'tsev, E. G. Fesenko & A. L. Khodakov. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 2, pp. 211–216.) Three groups of $BaTiO_3$ crystals were investigated, of which two were grown from a solution and one was obtained by reaction between $BaCO_3$ and Na_2TiO_3. The temperature and frequency dependence of permittivity and loss angle were studied.

537.226 : 546.431.824-31 **3010**

The Effect of Deviations from the Stoichiometric Composition on the Properties of $BaTiO_3$ Ceramics in Strong Fields.—N. S. Novosil'tsev, A. L. Khodakov & M. S. Shul'man. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 3, pp. 336–339.) An experimental investigation shows that even slight deviations from the stoichiometric composition (from 1·024 to 0·940) seriously affect the dielectric properties of $BaTiO_3$. The dependence of permittivity and loss angle on the composition was investigated at mains frequency and several anomalies in strong fields were observed.

537.226 : 546.431.824-31 **3011**

Theory of the Dielectric Properties of Barium Titanate in Stationary Fields.—A. E. Glauberman & A. F. Lubchenko. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 2, pp. 188–198.) Approximate calculations are made of the effective field in $BaTiO_3$ and of its permittivity below the Curie point. The calculations are based on the concept of the existence of a co-valent bond between the Ti ion and one of the oxygen ions, taking account of the lattice structure. Theory is developed for the qualitative determination of the temperature dependence of permittivity for temperatures below the Curie point, and a method is indicated for calculating the dependence of permittivity on the intensity of the external field and also the dependence of spontaneous polarization on temperature.

537.226.33 : 546.431.824-31 **3012**

Hysteresis Loops of Ceramic Barium Titanate at Higher Frequencies: Part 1.—K. Kambe, I. Nakada & H. Takahasi. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 9–14.) Report of observations, using pulse voltages, of polarization hysteresis loops at frequencies up to 3 Mc/s and temperatures from room temperature up to 150°C.

537.226.33 : 546.431.824-31 **3013**

Hysteresis Loops of Ceramic Barium Titanate at Higher Frequencies: Part 2.—K. Kambe. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 15–20.) Investigations show that with pulse voltage the shape of the hysteresis loops corresponds to the initial shape on applying direct voltage. The loop shape is almost independent of pulse width and shape, and varies only slightly in the range 10–100 kc/s.

537.228.1 : 534.213 **3014**

On the Theory of Sound-Wave Propagation in Piezoelectric Crystals.—R. Meier & K. Schuster. (*Ann. Phys., Lpz.*, 10th Feb. 1953, Vol. 11, No. 8, pp. 397–406.) Formal treatment of the propagation of plane acoustic waves, taking account of electro-mechanical reactions. Calculations are made of the velocity of propagation and oscillation directions in quartz, LSH ($Li_2SO_4H_2O$) and ADP crystals.

537.311.35 **3015**

Theory of Thermal Rectification: Part 1—Thermal Rectification Effect.—K. B. Tolpygo & I. M. Tsidil'kovski. (*Zh. tekh. Fiz.*, Sept. 1952, Vol. 22, No. 9, pp. 1442–1454.) A theoretical discussion of the effect discovered and investigated by Amirkhanov et al. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, 1941, Vol. 5, p. 447.) The voltage/current characteristic is derived for a homogeneous semiconductor with no resistances at the electrodes but with a temperature gradient in the direction of the electric field. See also 2552 and 2602 of 1946 (Amirkhanov).

537.311.33 3016

Electronic Conductivity of Anthracene Single Crystals.—H. Mette & H. Pick. (*Z. Phys.*, 17th April 1953, Vol. 134, No. 5, pp. 566–575.) The preparation of pure anthracene crystals is described. Measurements in the range 85–190°C established an exponential dependence of the conductivity on the temperature. The results show that the crystals are semiconductors.

537.311.33 3017

Cyclotron Resonances, Magnetoresistance, and Brillouin Zones in Semiconductors.—W. Shockley. (*Phys. Rev.*, 1st May 1953, Vol. 90, No. 3, p. 491.)

537.311.33 : 538.632 3018

The Adiabatic Hall Effect in Semiconductors.—V. A. Johnson & F. M. Shipley. (*Phys. Rev.*, 15th May 1953, Vol. 90, No. 4, pp. 523–529.) The relative difference between the adiabatic and isothermal Hall coefficients is calculated for the following cases:—(*a*) 'classical' impurity semiconductor; (*b*) degenerate impurity semiconductor; (*c*) 'classical' semiconductor at high temperature. For the thermal and electrical conductivities characteristic of Si and Ge, the relative differences are of the order of 1% or less. Tables are presented from which the relative difference may be calculated for any semiconductor of known electrical conductivity, thermal conductivity, and variation of carrier density with temperature.

537.311.33 : [546.863.221 + 546.873.221 3019

Electrical Properties of Sb_2S_3 and Bi_2S_3.—G. Galkin, G. Dolgikh & V. Yurkov. (*Zh. tekh. Fiz.*, 1952, Vol. 22, No. 10, pp. 1533–1539.) The temperature dependence of electrical conductivity of Sb_2S_3 and Bi_2S_3 samples was investigated. The value and sign of the temperature coefficients of sulphides as well as of the thermo-e.m.f. of the metal/semiconductor couple depend essentially on the heat treatment of the samples and on the temperature range under investigation. The results obtained are discussed from the standpoint of the band theory of semiconductors.

537.311.33 : 621.314.7 3020

Impurity Diffusion and Space-Charge Layers in 'Fused-Impurity' *p-n* Junctions.—J. S. Saby & W. C. Dunlap, Jr. (*Phys. Rev.*, 15th May 1953, Vol. 90, No. 4, pp. 630–632.) Diffusion calculations for various impurities, heating times and temperatures show that diffusion for even a few seconds at very moderate temperatures results in the introduction of more than 2×10^{12} atoms per cm^2 in the case of the common donors or acceptors. Diffusion of this order ensures the location of the space-charge layer entirely within the base material.

537.311.33 : 621.314.7 3021

Transistors: Theory and Application: Part 5—Point-Contact Transistor Operation.—A. Coblenz & H. L. Owens. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 158–163.) A simple but detailed analysis of current theory of operation.

537.311.33 : 621.396.822 3022

Boundary-Layer Noise in Semiconductors.—W. M. Buttler. (*Ann. Phys., Lpz.*, 16th Jan. 1953, Vol. 11, Nos. 4/7, pp. 362–367.) Starting from Gisolf's formula for current fluctuations (667 of 1950) and considering mean-square field strength, an expression is derived which gives the total noise figure, taking account of both boundary-layer noise and noise originating within the semiconductor.

537.311.33 : 621.396.822 3023

Boundary-Layer Noise in CdS Single Crystals.—W. M. Buttler. (*Ann. Phys., Lpz.*, 16th Jan. 1953, Vol. 11, Nos. 4/7, pp. 368–376.) The expression derived in 3022 above forms the basis of a new theory of boundary-layer noise, which is the greater the less the length of the semiconductor. Experiments to determine the dependence of noise on the applied voltage are described. Results indicate that in CdS the noise originates predominantly in the boundary layer.

538.221 3024

Determination of the Coercive Force of Nonuniformly Magnetized Specimens.—A. I. Bulanova, M. N. Mikhnevich & V. B. Perets. (*Zh. tekh. Fiz.*, Aug. 1952, Vol. 22, No. 8, pp. 1325–1333.) Results are given of an experimental and theoretical investigation of the errors which arise in the determination of coercive force in specimens of complex configuration and magnetically anisotropic structure.

538.221 3025

Ferromagnetism and Order in Nickel/Manganese Alloys.—G. R. Piercy & E. R. Morgan. (*Canad. J. Phys.*, May 1953, Vol. 31, No. 4, pp. 529–536.)

538.221 3026

Theoretical Considerations on the Effect of Ellipsoidal Inclosures, of Dimensions Small compared with the Thickness of the Bloch Wall, on the Ferromagnetic Coercive Force.—E. Schwabe. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 99–112.)

538.221 3027

The Structure of the Ferromagnetic State in αFe and some Fe Alloys.—G. Heber. (*Ann. Phys., Lpz.*, 20th Nov. 1952, Vol. 11, Nos. 2/3, pp. 155–160.)

538.221 : 538.652 3028

On the Theory of Magnetostriction and of Other Even Effects in Strong Magnetic Fields.—G. P. D'yakov. (*Zh. eksp. teor. Fiz.*, 1952, Vol. 23, No. 5, pp. 525–531.) A mathematical theory is developed for multicrystal ferromagnetic materials. Formulae are derived for calculating even effects in non-textural materials, and the following conclusions are made: (i) the law of the approach of even effects to saturation relates all the main characteristics of a ferromagnetic material to one another; (ii) on the basis of this law it is possible, from measurements of magnetostriction in multicrystal samples, to determine all the main constants of ferromagnetic materials; (iii) from the experimental variation of magnetostriction when approaching saturation it is possible to determine the constant of the magnetic energy anisotropy and its sign. The relations obtained are generalized to cover the case when elastic stresses are present in the material, and the following two limiting cases are considered: (*a*) when stresses are nonuniform, (*b*) when stresses are uniform and coincide with the direction of the magnetic field.

538.221 : 621.396.822 3029

The Space Correlation of Noise of Cyclical Magnetic Reversals.—A. A. Grachev. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1952, Vol. 85, No. 4, pp. 741–744. In Russian.) Investigations on permalloy wires.

538.221 : 669.15.782-194 3030

Grain-Oriented Iron-Silicon Alloys.—G. H. Cole. (*Elect. Engng, N.Y.*, May 1953, Vol. 72, No. 5, pp. 411–416.) Chemical compositions and graphs of the magnetic properties of typical alloys are given. The maximum permeability is obtained when the orientation is in the direction of rolling, parallel to the applied field. This gives also the condition for minimum power losses. Applications of such materials are noted.

539.234 : 537.311.31 **3031**

Experimental Study of the Electrical Conductivity of Very Thin Metal Films obtained by Thermal Evaporation.—N. Mostovetch. (*Ann. Phys., Paris*, Jan./Feb. 1953, Vol. 8, pp. 61–125.) Full report and discussion of results of investigations on various metal films of thickness < 100 Å deposited on glass, quartz and mica. Results obtained show (*a*) that the action of absorbed gases causes a decrease in resistance; (*b*) exponential variation of resistance with temperature; (*c*) a linear relation between log R and $F^{\frac{1}{2}}$ where R is resistance and F is the intensity of an applied electric field of between 1 and 50 V/cm. See also 376, 1634 (Mostovetch & Duhautois) and 2192 (Mostovetch et al.) of 1952.

539.234 : 546.289 **3032**

Evaporation of Germanium Films from a Carbon Crucible.—K. Lehovec, J. Rosen, A. MacDonald & J. Broder. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 513–514.) Equipment for evaporation of Ge in vacuo at various measurable rates is described. The activation energy of evaporation is 3·5 eV, a value in close agreement with data recently published by Searcy.

546.28 + 546.289 **3033**

The Solubility of Silicon and Germanium in Gallium and Indium.—P. H. Keck & J. Broder. (*Phys. Rev.*, 15th May 1953, Vol. 90, No. 4, pp. 521–522.) Solubility values for the temperature range of interest in preparing junction transistors are presented in graphs. Si crystals of fair purity have been obtained by slow cooling of saturated solutions.

546.831 : 621.385.032.461 **3034**

Quality Estimation of Powdered Zirconium for use in Valve Manufacture.—H. Figour & H. Bonnel. (*Le Vide*, March 1953, Vol. 8, No. 44, pp. 1305–1306.) A simple method of evaporation in vacuo is suggested for rapid determination of the quality of Zr intended as getter material.

548.7 : [546.732-31 + 546.711.241 **3035**

The Antiferromagnetic Structure Deformations in CoO and MnTe.—S. Greenwald. (*Acta cryst., Camb.*, 10th May 1953, Vol. 6, Part 5, pp. 396–398.)

548.7 : 546.76.86 **3036**

Crystal Structure and Antiferromagnetism of CrSb.—B. T. M. Willis. (*Acta cryst., Camb.*, 10th May 1953, Vol. 6, Part 5, pp. 425–426.)

621.315.615.9 **3037**

Some Fluorinated Liquid Dielectrics.—N. M. Bashara. (*Elect. Engng, N.Y.*, May 1953, Vol. 72, No. 5, p. 424.) Digest only. The remarkable properties of fluorocarbon materials, including stability at high temperatures, high dielectric strength, low power factor, non-inflammability, etc., make them suitable for many electrical applications.

621.318.2.042.15 **3038**

Micropowder Magnets.—(*Elect. J.*, 8th May 1953, Vol. 150, No. 3908, pp. 1645–1646.) Short account of the properties of permanent magnets now manufactured in England. Coercive forces approaching 1 000 oersted can be obtained with pure iron powder, the best results being obtained with a particle size between 0·01 and 0·1 μ. The best micropowder available includes 30% Co and has an energy factor equivalent to that of alnico, but its density of about 4·5 gm/cm³ gives it a great advantage over cobalt steel or similar alloys. Similar accounts in *Engineering, Lond.*, 29th May 1953, Vol. 175, No. 4557, pp. 682–684, and *Engineer, Lond.*, 22nd May 1953, Vol. 195, No. 5078, pp. 740–741.

621.318.23 **3039**

Optimum Design of Permanent Magnets.—H. K. Ziegler. (*Elect. Engng, N.Y.*, May 1953, Vol. 72, No. 5, p. 445.) Digest only. Discussion of the use of the average value of the slope of minor loops of the B/H demagnetization curve, in the region of the maximum energy product, for optimum magnet design. Design curves are given for Alnico V, VI and XII.

621.383.4 : 546.817.221 **3040**

Effect of Oxygen on Conductivity and Photoconductivity of Lead-Sulphide Photoresistors.—S. M. Rivkin & L. N. Malakhov. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1952, Vol. 85, No. 4, pp. 765–768. In Russian.) The variation of conductivity with the degassing time at temperatures in the range 32·5–63°C was investigated. The characteristics of PbS photoresistors are apparently determined solely by the effect of oxygen on the properties of the interstitial layer surrounding the grains of PbS. See also 702 of 1952 (Ehrenberg & Hirsch).

MATHEMATICS

517.63 **3041**

A Contribution to the Clear Interpretation of the Laplace Transformation.—H. F. Schwenkhagen. (*Elektrotech. Z., Edn A*, 1st March 1953, Vol. 74, No. 5, pp. 133–136.) Discussion of analytical methods and the connection between Fourier and Laplace transforms.

681.142 **3042**

Linear-to-Logarithmic Voltage Converter.—R. C. Howard, C. J. Savant & R. S. Neiswander. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 156–157.) For use with analogue computers. The logarithmic output is obtained by adjustment of anode voltage and anode and cathode resistors in a triode circuit with a large grid resistor and a much smaller input-to-cathode resistor. A cathode-follower stage cancels the offset voltage and also isolates the converter from the actual output. Logarithmic bases and output voltages obtainable with various types of valve are listed. The logarithm base can be adjusted, after conversion, by changing the gain.

681.142 **3043**

The Analog Computer.—D. H. Pickens. (*Product Engng*, May 1953, Vol. 24, No. 5, pp. 176–185.) Description of the different types of element used in analogue computers, and their combination for the solution of specific problems.

681.142 **3044**

Analog Computer for the Roots of Algebraic Equations.—L. Löfgren. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 907–913.)

681.142 **3045**

DINA, a Digital Analyzer for Laplace, Poisson, Diffusion, and Wave Equations.—C. Leondes & M. Rubinoff. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 303–308. Discussion, pp. 308–309.)

681.142 **3046**

Basic Features of a Programme for a Chess-playing Computer.—G. Schliebs. (*Funk u. Ton*, May 1953, Vol. 7, No. 5, pp. 257–265.)

681.142 : 519.272.119 **3047**

Simple Computer Automatically plots Correlation Functions.—A. H. Schooley. (*Tele-Tech*, May 1953, Vol. 12, No. 5, pp. 71–73..158.) An analogue computer for use in evaluating the autocorrelation and cross-correlation functions of time functions is described. Loops of 35-mm film, cut to the shape of the time functions, are moved by two sprockets at a uniform speed. Slide-wire resistors are operated by levers resting against the cut edges of the films; the resistor outputs are passed via a RC filter to the amplifier and paper-strip recorder.

681.142 : 621.314.3† **3048**

The Single-Core Magnetic Amplifier as a Computer Element.—R. A. Ramey. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 442–446.)

681.142 : 621.315.612.4 **3049**

Ferroelectric Materials as Storage Elements for Digital Computers and Switching Systems.—J. R. Anderson. (*Trans. Amer. Inst. elect. Engrs*, 1952, Vol. 71, Part I, pp. 395–401.)

681.142 : 621.395.623.7 **3050**

An Analogue for Use in Loudspeaker Design.—J. J. Baruch & H. C. Lang. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 89–93; *Trans. Inst. Radio Engrs*, Jan./Feb. 1953, Vol. AU-1, No. 1, pp. 8–13.) Description of a computer designed to facilitate analysis of loudspeaker performance as a function of frequency.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) : [621.314.7 + 621.396.611.21 **3051**

Precision Transistor Oscillator.—P. G. Sulzer. (*Radio & Telev. News, Radio-Electronic Engng Section*, May 1953, Vol. 49, No. 5, pp. 18, 29.) See 2064 of July.

621.3.087.4 : 551.510.535 **3052**

A 16-kW Panoramic Ionospheric Recorder.—R. Lindquist. (*Chalmers tek. Högsk. Handl.*, 1951, No. 109, 41 pp.) Reprint. See 1425 of 1951.

621.314.7.001.4 **3053**

Methods and Equipment for Transistor Testing.—B. J. O'Neill & A. Gutterman. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 172–175.) Description of c.r.o. equipment for tracing characteristics, and of an accessory circuit for re-forming defective units.

621.317.3 : 538.632 **3054**

On the Geometrical Arrangement in Hall Effect Measurements.—V. Frank. (*Appl. sci. Res. B*, 1953, Vol. 3, No. 2, pp. 129–140.) The short-circuiting effect of the current electrodes in Hall-effect measurements is discussed for the case of an arbitrary geometrical arrangement. For singly connected geometries the correction to be applied is a universal function of a single parameter characteristic of the geometry. This parameter can be determined experimentally. The theory was verified experimentally for a particular geometry by measurements on copper.

621.317.3 : 621.396.822](083.74) **3055**

Standards on Electron Devices: Methods of Measuring Noise.—(*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 890–896.) Standard 53 IRE 7S1.

621.317.331.028.3 **3056**

Measurement of Multimegohm Resistors.—A. H. Scott. (*J. Res. nat. Bur. Stand.*, March 1953, Vol. 50, No. 3, pp. 147–152.) Measurements made by a null method using an electrometer are described. A variable air-dielectric capacitor supplies the charge flowing through the resistor on test. The potential is maintained constant by decreasing capacitance at the correct rate. Protracted measurements on resistors in the range 10^8–10^{11} Ω show that the most stable resistors available are subject to fluctuations of 0·5–1·0% and are generally sensitive to voltage change. For a shorter account, see *Tech. News Bull. nat. Bur. Stand.*, June 1953, Vol. 37, No. 6, pp. 90–92.

621.317.335 **3057**

Use of Classical Circuits for the Measurement of Dielectric Constants and Losses at Frequencies between 80 c/s and 40 Mc/s.—R. Dalbert. (*Rev. gén. Élect.*, May 1953, Vol. 62, No. 5, pp. 237–246.) Analysis is made of three methods using known bridge arrangements in which the required determination is effected by obtaining a balance. The accuracy of these methods is evaluated, and the influence of parasitic circuit elements on the validity of the approximations introduced is studied.

621.317.336 : 621.315.212 **3058**

More on the Sweep-Frequency Response of RG/6U Cable.—D. A. Alsberg. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, p. 936.) Comment on 3168 of 1952 (Blackband).

621.317.35 : 621.3.018.783 **3059**

The Rigorous Basis of an Intermodulation Method of Nonlinear-Distortion Measurement.—H. Söding. (*Frequenz*, May 1953, Vol. 7, No. 5, pp. 127–133.) The mathematical representation of the intermodulation factor of an amplifier having a characteristic of order n is considered. The measurement circuit used is shown schematically and the method of exact determination of the intermodulation factor from two voltage readings is established. See also 2734 (Bloch), 2735 (Maxwell), 2736 (Roys) and 2737 (Scott) of September.

621.317.4 : 538.221 **3060**

Measuring Methods for some Properties of Ferroxcube Materials.—C. M. van der Burgt, M. Gevers & H. P. J. Wijn. (*Philips tech. Rev.*, March 1953, Vol. 14, No. 9, pp. 245–256.) For measurement of permeability and losses in weak fields a resonance method is applied using (*a*) lumped circuit elements for the frequency range 50 kc/s–10 Mc/s, (*b*) a coaxial line for the range 5 Mc/s–3 kMc/s. Sources of error and limitations of the methods are discussed. In strong fields, determination of the magnetization curve replaces permeability measurement. For measurement of losses, a calorimetric or coaxial-line method may be used according to flux density and operating frequency. Distortion is measured by a selective voltmeter or, at frequencies below about 2 kc/s, may be determined from the hysteresis resistance. See also 2823 of 1952 (Went & Gorter).

621.317.41 **3061**

Apparatus for Bulk Tests on Sheet Iron.—E. Wettstein. (*Helv. phys. Acta*, 15th Sept. 1952, Vol. 25, No. 5, pp. 488–490. In German.) Outline description of equipment for rapid measurement of permeability and thickness.

621.317.444 : 621.396.645.35 : 621.317.755 **3062**

An Integrating Amplifier for the Oscillographic Recording of Magnetic Flux.—S. Ekelöf, L. Bengtson, G. Kihlberg & P. Leithammel. (*Chalmers tek. Högsk. Handl.*, 1951, No. 120, 23 pp.) Reprint. See 2276 of 1952.

621.317.7 : 621.396.61/.62 **3063**

Test Equipment for the TD-2.—A. S. May. (*Bell Lab. Rec.*, May 1953, Vol. 31, No. 5, pp. 187–193.) Illustrated descriptions are given of (*a*) transmitter/receiver testing equipment used at all the stations of this radio relay system; this includes i.f. (70 Mc/s ± 25 Mc/s) and r.f. (3·7–4·2 kMc/s) variable-frequency generators, signal-power meters and oscilloscope; (*b*) equipment for terminal stations only, including oscilloscope, f.m. test set and linearity test set.

621.317.72 **3064**

An Isolating Potential Comparator.—T. M. Dauphinee. (*Canad. J. Phys.*, May 1953, Vol. 31, No. 4, pp. 577–591.) A double-pole double-throw chopper, with a capacitor connected between the vibrating contacts, is used with a galvanometer or electronic amplifier to indicate the inequality between two e.m.f.'s such as those of two thermocouples. There is no net current flow between the circuits. Theory of the method is given and applications are suggested.

621.317.73.012.11 **3065**

Automatic Smith-Chart Plotter.—K. S. Packard. (*Tele-Tech*, April 1953, Vol. 12, No. 4, pp. 65–67..183.) An outline description of the Impedance Plotter is given. This instrument is designed to cover 100–400 Mc/s in two ranges, either range being swept at 2 c/s. The measured impedance/frequency curve is traced out on a c.r. screen with a Smith-chart overlay.

621.317.75 : 621.392.26 **3066**

A Direct-Reading Standing-Wave Indicator.—A. C. Grace & J. A. Lane. (*J. sci. Instrum.*, May 1953, Vol. 30, No. 5, pp. 168–169.) A waveguide squeeze-section [2865 of 1948 (Montgomery)] is used together with a c.r.o. to give an instantaneous indication of changes in s.w.r. at wavelengths of the order of 1 cm. One side of the squeeze-section is fixed, while the other is given a small transverse displacement by means of a cam driven at about 10 r.p.s. The test signal, modulated at 10 kc/s, is obtained from a reflex klystron giving about 20 mW and is detected by a Type-VX-4042 crystal. Load matching to within about 2% can be achieved.

621.317.755 **3067**

A Direct-Reading Oscilloscope for 100-kV Pulses.—R. C. Hergenrother & H. G. Rudenberg. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 896–901.) Description of the design, construction and performance of a unit using e.s. deflection, and responding to pulses of duration of a fraction of a microsecond.

621.317.755 : 531.765 **3068**

A Recurrent-Sweep Chronograph.—H. Moss. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 282–286.) The operation and circuit details of a counter-type chronograph designed for ballistics measurements are described. Timing signals are displayed on a c.r.o. so that confusion with spurious noise signals is avoided. In the form of clipped pulses they are traced on a 10-line raster on which a 10-line time scale is superimposed. Line-sweep and frame-sweep generators are controlled by a 1-kc/s tuning fork operating through decade frequency dividers. The maximum period measurable is 10^5 μs; time resolution, ± 15 μs.

621.317.755 : 621.396.677.012.12 **3069**

Aerial Radiation Patterns. Apparatus for Cathode-Ray-Tube Display.—T. T. Ling. (*Wireless Engr*, Aug. 1953, Vol. 30, No. 8, pp. 192–195.) The receiving aerial under investigation is mounted on a turn-table rotatable at 1 to 30 r.p.m. The r.f. power picked up by the aerial from a pulsed magnetron transmitter (3·2 cm λ) is converted to a sine wave whose amplitude is proportional to the power of the received signal. The sine wave is applied to the primary of a magslip, geared to the turn-table, via a cathode follower. The outputs from the two mutually perpendicular secondaries, after amplification, are fed to the *x* and *y* coils of a c.r.o. with long-persistence screen. As the turn-table rotates, a luminous area is swept out on the screen, the contour representing the aerial polar diagram. With suitable blanking arrangements, the contour only is obtained. Typical radiation patterns are illustrated.

621.396.611.21.029.45/.51 **3070**

The Sunbury Portable Frequency Standard for 100 c/s to 100 kc/s.—A. H. Morser. (*Electronic Engng*, July 1953, Vol. 25, No. 305, pp. 290–293.) Description, with complete circuit details, of a tone source designed for calibrating electronic tachometers. It provides an output >25 V r.m.s. and accurate to within 0·1% at 100, 10, 1 or 0·1 kc/s, derived from a vacuum-mounted 100-kc/s quartz crystal. Frequency division is effected by regenerative modulation.

621.396.619.13 : 621.317.755 **3071**

A Method for Establishing the Modulation Index of Periodic F.M. Signals.—A. H. Phillips & M. Cooperstein. (*Sylvania Technologist*, April 1953, Vol. 6, No. 2, pp. 31–34.) The case of modulation by a single sinusoidal voltage is considered. By means of a mixing process the original f.m. signal is used to derive a new f.m. signal whose carrier frequency is less than the f.m. excursion, so that over certain parts of the modulation cycle the instantaneous frequency is 'negative', corresponding to a negative rate of change of phase. The output from the mixer, after passing through a low-pass filter, is displayed on an oscilloscope, and the modulation index is deduced from the observed number of oscillations between the points of stationary phase. The use of high-order side-frequency components for rapid interpretation of large modulation indices is demonstrated.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.522.1 : 539.3 **3072**

Ultrasonic Waves measure the Elastic Properties of Polymers.—G. W. Willard. (*Bell Lab. Rec.*, May 1953, Vol. 31, No. 5, pp. 173–179.) Ultrasonic light-diffraction apparatus operating on the principle previously described (3754 of 1947) is used to measure the velocity of ultrasonic waves in the polymer; the elastic constants of the material can then be determined.

535.82 : 621.397.611.2 **3073**

Flying-Spot Microscope.—F. Roberts, J. Z. Young & D. Causley. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 137–139.) See 1733 of 1951 (Young & Roberts). Other applications of this microscope are briefly mentioned.

537.533 : 535.417 **3074**

Electron Beam Interferometer.—L. Marton, J. A. Simpson & J. A. Suddeth. (*Phys. Rev.*, 1st May 1953, Vol. 90, No. 3, pp. 490–491.) A practical form of the instrument discussed previously [2562 of 1952 (Marton)] uses a three-crystal beam-splitting device.

621.316.7 : 681.142 **3075**

Analysis of Control Systems involving Digital Computers.—W. K. Linvill & J. M. Salzer. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 901–906.)

621.384.611/.612 **3076**

The Synchrocyclotron at Amsterdam: Part 4 — Details of Construction and Ancillary Equipment.—F. A. Heyn & J. J. Burgerjon. (*Philips tech. Rev.*, March 1953, Vol. 14, No. 9, pp. 263–279.) Part 3: 223 of 1952 (Heyn). Please note change in U.D.C. numbers for particle accelerators.

621.384.622.2 **3077**

Axial Motions of Electrons in a Linear Accelerator with $\beta = 1$.—J. Swihart & E. Akeley. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 640–643.)

621.384.622.2 **3078**

Study and Development of a Standing-Wave Linear Accelerator for Electrons.—A. Sarazin. (*Ann. Radioélect.*, April & July 1953, Vol. 8, Nos. 32 & 33, pp. 134–152 & 228–263.) Theoretical and experimental investigations relating to an accelerator giving 1·2-MeV electrons.

621.384.622.2 : 537.291 **3079**

Effect of Variation of the Amplitude of the Accelerating Field on the Motion of Ions in Linear Accelerators.—M. Y. Bernard. (*C. R. Acad. Sci., Paris*, 8th June 1953, Vol. 236, No. 23, pp. 2226–2228.)

621.385.833 **3080**

A Method of Observing Selected Areas in Electron and Optical Microscopes.—J. F. Nankivell. (*Brit. J. appl. Phys.*, May 1953, Vol. 4, No. 5, pp. 141–143.) Brief description of simple apparatus for positioning plastic replicas as desired on electron-microscope specimen grids, to enable direct comparison to be made of electron- and photo-micrographs.

621.385.833 **3081**

Aberrations of Electronic Images of Irregular Emissive Cathodes.—F. Bertein. (*J. Phys. Radium*, April 1953, Vol. 14, No. 4, pp. 235–240.) Investigation of the astigmatism, field curvature and distortion resulting from surface irregularity. The astigmatism can appreciably reduce the resolution.

621.385.833 **3082**

The Field Emitter: Fabrication, Electron Microscopy, and Electric-Field Calculations.—W. P. Dyke, J. K. Trolan, W. W. Dolan & G. Barnes. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 570–576.)

621.385.833 **3083**

Practical Calculation of the Third-Order Aberrations of Centred Electronoptical Systems.—M. Barbier. (*Ann. Radioélect.*, April 1953, Vol. 8, No. 32, pp. 111–133.)

621.385.833 + 621.392.5].001.362 **3084**

Representation of Optical or Electronic Lenses by Quadripoles consisting of Pure Resistances.—F. Bertein. (*C. R. Acad. Sci., Paris*, 29th June 1953, Vol. 236, No. 26, pp. 2494–2496.) The equivalence is established for certain types of lens.

621.387.42 **3085**

Low-Voltage Counter Tubes.—P. Kunze & G. Schulz. (*Ann. Phys., Lpz.*, 16th Jan 1953, Vol. 11, Nos. 4/7, pp. 225–238.) Discussion of design requirements and report of experimental investigations on tubes with Ne, He, Ar and Hg-vapour filling.

621.387.424 **3086**

The Functioning of Geiger-Müller Self-Quenching Counters.—D. Blanc. (*J. Phys. Radium*, April 1953, Vol. 14, No. 4, pp. 260–269.) Review of theory of operation, different types, and applications. 152 references.

621.387.424 **3087**

Geiger-Müller Counters with External Cathode and Pure Methyl-Alcohol Filling, for Intense Radiation.—D. Blanc. (*J. Phys. Radium*, April 1953, Vol. 14, No. 4, pp. 271–272.) A counter of this type, 8-cm long and 2-cm diameter, with vapour pressure of 2 cm Hg, has a threshold voltage of 1·3 kV, a plateau of 500 V with a slope of 2% per 100 V, and a counting rate of 20 000/min. It can deal with radiation much more intense than counters with usual fillings.

621.387.424 **3088**

The Photoelectric Sensitivity of Self-Quenching Geiger Counters with Glass Envelope and External Cathode.—H. Schwarz. (*Z. Phys.*, 17th April 1953, Vol. 134, No. 5, pp. 540–545.)

621.387.424 **3089**

The Maze Counter.—A. Aron. (*Z. Phys.*, 17th April 1953, Vol. 134, No. 5, pp. 622–641.) Investigation of the characteristics of G-M counters with glass envelope and external cathode.

PROPAGATION OF WAVES

621.396.11 **3090**

Radiocommunication on Frequencies exceeding Predicted Values.—E. V. Appleton & W. J. G. Beynon. (*Proc. Instn elect. Engrs*, Part III, July 1953, Vol. 100, No. 66, pp. 192–198.) Communication data for B.B.C. transmissions during 1943 at frequencies around 20 Mc/s showing cases of satisfactory reception of frequencies higher by 1–6 Mc/s than the predicted m.u.f. are analysed. A high proportion of the apparent anomalies occurring during the winter and equinoctial months can be explained in terms of reflection from the normal E region and of variability of F_2 normal-incidence critical frequency. The remainder can be attributed to reflections from the abnormal-E layer. Analysis of Japanese 30-Mc/s data for the years 1935–1939 leads to similar conclusions.

621.396.11 : 535.42 **3091**

Some V.H.F. Experiments upon the Diffraction Effect of Hills.—G. C. Rider. (*Marconi Rev.*, 2nd Quarter 1953, Vol. 16, No. 109, pp. 96–106.) Experiments at 160·2 Mc/s on propagation over steep hills show that the signal increase obtained when passive repeater aerials are used may be satisfactorily predicted by the method normally used for calculating the field in the shadow of a knife edge. Passive repeaters are expected to be most useful in the u.h.f. band. For prediction of field strength across a valley, when transmitter and/or receiver are close to the hill, the spherical diffraction treatment gives results in closer agreement with experiment than the knife-edge method. A preliminary survey of the polarization and direction of arrival of radio waves at receiver positions round the hill was also made.

621.396.11 : 551.510.535 **3092**

C.C.I.R.-U.R.S.I. Co-operation. Exchange of Information for Forecasts and Warnings.—(*U.R.S.I. Inform. Bull.*, March/April 1953, No. 78, pp. 42–54.) Report on the steps taken in the U.S.A. The current codes used by the C.R.P.L. for forecasts and cosmic data are given in an appendix.

621.396.11 : 551.510.535 **3093**

Some Calculations of Ray Paths in the Ionosphere.—S. K. H. Forsgren. (*Chalmers tek. Högsk. Handl.*, 1951, No. 104, 23 pp.) Reprint. See 782 of 1952.

621.396.11 : 551.510.535 (98) **3094**

Polar Blackouts Recorded at the Kiruna Observatory.—R. Lindquist. (*Chalmers tek. Högsk. Handl.*, 1951, No. 103, 25 pp.) Reprint. See 783 of 1952.

621.396.11 : 621.317.353.3† **3905**

The Theory of the Interaction of Radio Waves in the Ionosphere.—I. M. Vilenski. (*Zh. eksp. teor. Fiz.*, May 1952, Vol. 22, No. 5, pp. 544–561.) An elementary theory of the Luxemburg effect was given by Bailey (1934 Abstracts, p. 606). In a subsequent paper by Ginzburg, a kinetic theory of the phenomenon was presented in which collisions between electrons and ions were taken into account. These two papers, however, considered only the depth and phase of the cross-modulation of the waves which had passed through the 'disturbed' region, and the high-frequency terms of the interaction were neglected. A more detailed investigation, taking account of all terms, is here given. In addition to cross-modulation, 'side' waves of combination frequencies are found to appear. Formulae are derived for determining the amplitudes of the 'side' waves, and the effect of collisions between electrons and ions is considered for various cases.

621.396.11.029.53 **3096**

The Influence of the Ionosphere on Medium-Wave Broadcasting.—G. J. Phillips. (*B.B.C. Quart.*, Spring 1953, Vol. 8, No. 1, pp. 40–54.) A general summarized description of the ionosphere and its characteristic effects, with particular reference to those affecting m.f. broadcasting. Reduction of fading achieved by use of mast radiators designed to minimize the sky wave may be limited by the diffuseness of reflection from the ionosphere. Quantitative measurements of the angular scatter on reflection by the ionosphere are being made at medium frequencies.

621.396.11.029.64 **3097**

Microwave Propagation in the Optical Range.—O. F. Perers, B. K. E. Stjernberg & S. K. H. Forsgren. (*Chalmers tek. Högsk. Handl.*, 1951, No. 108, 20 pp.) Reprint. See 789 of 1952.

621.396.812.3 **3098**

On the Frequency of the Scintillation Fading of Microwaves.—O. Tukizi. (*J. phys. Soc. Japan*, Jan./Feb. 1953, Vol. 8, No. 1, pp. 130–131.) Experimental results are reported which appear to support the theoretical prediction of proportionality between the scintillation-fading frequency and the mean wind speed over the propagation path.

RECEPTION

621.396.621 **3099**

Counter-Circuit Multiplex Receiver.—A. R. Vallarino, H. A. Snow & C. Greenwald. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 178–181.) In a f.m. multiplex subcarrier system, a r.f. receiver demodulates the r.f. signal and passes the resulting f.m. 1-Mc/s subcarrier to a second receiver for demodulation. The subcarrier receiver described involves no alignment problems and introduces only 0·2% distortion over the modulation frequency band 250 c/s–150 kc/s. Details are given of the 6-stage limiter using cathode-coupled clippers, and of the counter-discriminator circuit. Output voltage is a linear function of input frequency.

621.396.722 : 551.594.6 **3100**

Existing and Planned Spherics Stations.—(*U.R.S.I. Inform. Bull.*, March/April 1953, No. 78, pp. 38–41.) The geographical position, the frequency of operation and type of equipment of 52 stations are tabulated.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **3101**

Information Theory.—P. M. Woodward. (*Brit. J. appl. Phys.*, May 1953, Vol. 4, No. 5, pp. 129–133.) An introduction to the early work of Hartley and the newer statistical theory developed by Shannon.

621.396.1 : 621.396.931 **3102**

Channel Spacings at 152 to 174 Mc/s.—H. E. Strauss. (*Commun. Engng*, March/April 1953, Vol. 13, No. 2, pp. 17–21 . . 48.) Mobile-radio tests were made on frequencies around 160 Mc/s, using channel spacings of 60, 30 and 20 kc/s. Tests covered range, adjacent-channel interference and intermodulation. A system having 30-kc/s channel spacing and ±7·5-kc/s modulation deviation was 4 or 5 db more sensitive to noise than the system with 60-kc/s channel spacing. A system with 20-kc/s channel spacing and ±5-kc/s modulation deviation was 8 db more sensitive to noise than that with 60-kc/s spacing and had 15% less range.

621.396.41.018.424 **3103**

Bandwidth of Multiplex Channels.—J. S. Smith. (*Commun. Engng*, May/June 1953, Vol. 13, No. 3, pp. 22–23 . . 43.) Discussion of factors involved in the relation of channel response speed to the subcarrier bandwidth.

621.396.619.13 : 621.3.018.78 **3104**

The Distortion Introduced by Frequency Modulation of an Oscillator by a Perfect Reactance Valve.—P. S. Brandon. (*Marconi Rev.*, 2nd Quarter 1953, Vol. 16, No. 109, pp. 88–95.) Assuming no distortion present when the instantaneous angular frequency is given by $1/(LC)^{\frac{1}{2}}$, analysis shows that a change in the inductance or capacitance of the oscillatory circuit introduces a form of amplitude modulation which in turn causes an unwanted change in frequency. Approximate expressions for the fundamental and second harmonic of this form of distortion are developed in terms of modulation depth, angular frequency of modulation, mean oscillator frequency and deviation.

621.396.619.16 **3105**

Consequences of the Amplitude Spectrum [transformation] **in Pulse Technology.**—H. Kleinwächter & H. Weiss. (*Funk u. Ton*, May 1953, Vol. 7, No. 5, pp. 221–229.) A general expression relating pulse duration and bandwidth required for transmission of square-wave pulse trains is derived by Fourier analysis. The bandwidth for which the signal/noise ratio is a maximum and the mean power required for p.a.m. transmission are calculated.

621.396.619.16 **3106**

Pulse-Code Modulation Systems.—A. J. Oxford. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 859–865.) Principles of p.c.m. systems are reviewed and two modulator systems are considered in detail. In one system, the incoming signal sample charges a capacitor, which is then 'examined' successively by 5 pulses, each of 30-μs duration, a charge of logarithmically decreasing amplitude being added to or subtracted from the capacitor before the arrival of the next pulse. This provides a five-digit code for transmission. In a simplified system, the residual charge on the capacitor is amplified after each pulse, so that a fixed amount of charge can be added to or subtracted from the capacitor. Demodulation and synchronization arrangements are also dealt with.

621.396.65 **3107**

A New Multiplex Microwave System.—N. B. Tharp. (*Commun. Engng*, March/April 1953, Vol. 13, No. 2, pp. 31–34 . . 46.) Types FR and FJ equipments operating together at about 1·9 kMc/s can provide 30 two-way voice-frequency channels. The Type-FR equipment comprises transmitter-receiver, oscillator-modulator, a.f.c., power-supply, service-channel and test-meter units. The Type-FJ equipment is the multiplexing unit.

621.396.65 **3108**

Microwave Developments Overseas.—V. J. Nexon. (*Commun. Engng*, May/June 1953, Vol. 13, No. 3, pp. 19–21 . . 46.) A survey of the v.h.f. and u.h.f. radio link systems used in 16 countries in Europe, North Africa and the Near East, and in Australia.

621.396.65 : 621.396.41 **3109**

Multichannel Radio Link between Mestre and Trieste.—L. Bernardi. (*Poste e Telecomunicazioni*, April 1953, Vol. 6, No. 4, pp. 159–167.) A description is given of the equipment used for this link, which operates in the frequency band 1·3–1·6 kMc/s and has a relay station at Clauzetto, the lengths of the two sections being respectively 98 and 92 km. Parabolic aerials are used at all three stations, the aerial power being 2·1 W. Performance curves for the whole system are given.

621.396.65 : 621.396.41.029.64 **3110**

A Multichannel Microwave Relay System.—R. D. Boadle. (*Proc. Instn Radio Engrs, Aust.*, April 1953, Vol. 14, No. 4, pp. 80–89.) Description of the development in Australia of equipment based on an American design [3283 of 1947 (Gerlach) and back references].

SUBSIDIARY APPARATUS

621-526 **3111**

The Application of Nonlinear Techniques to Servomechanisms.—K. C. Mathews & R. C. Boe. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 10–21.)

621-526 **3112**

The Transient Response of a Single-Point Nonlinear Servomechanism.—K. N. Burns. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 22–30.)

621-526 **3113**

Servo-System Comparators.—M. Cooperstein. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 31–36.)

621-526 : 621.314.3† **3114**

Compensation of a Magnetic-Amplifier Servo System.—H. H. Woodson & C. V. Thrower. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 158–165.)

621-526 : 621.392.5.076.12 **3115**

Synthesis of Compensation Networks for Carrier-Frequency Servomechanisms.—R. S. Carlson & J. G. Truxal. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 37–49.)

621.3.013.783 **3116**

Screening of Test Rooms and Instruments from Electromagnetic Fields.—J. Deutsch & O. Zinke. (*Frequenz*, April 1953, Vol. 7, No. 4, pp. 94–101.) Screens made of wire mesh, metal foil and sheet, and of materials with a high permittivity or permeability are considered and expressions are given for the attenuation of incident e.m. waves. Experimental curves for sheet and mesh screens are given. The construction of a large screened room is described and the attenuation of magnetic fields in this room (>70 db for frequencies from 200 kc/s to 500 Mc/s) is compared with that of other rooms. A novel type of honeycomb wire-mesh-strip window screen is described.

621.352.32 **3117**

Reactivating the Dry Cell.—R. W. Hallows. (*Wireless World*, Aug. 1953, Vol. 59, No. 8, pp. 344–347.) Discussion of the effect on cell life of repeated passage of current in the reverse direction.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24/.26 **3118**

The Television Transmission of the English Coronation Ceremonies.—A. W. M. Paling. (*Tijdschr. ned. Radiogenoot.*, May 1953, Vol. 18, No. 2, pp. 167–174.) An illustrated description of the radio-link system and line-standard converter devised to transmit pictures of the Coronation ceremonies to Western Europe.

621.397.335 : 621-526 .001.11 **3119**

Servomechanism Theory applied to A.F.C. Circuit Design.—G. D. Doland. (*Tele-Tech*, May 1953, Vol. 12, No. 5, pp. 95–97 . . 155.)

621.397.5 **3120**

Television Study Centre.—M. Boella. (*Ricerca sci.*, May 1953, Vol. 23, No. 5, pp. 799–802.) A survey of activities up to the end of 1952. Of the investigations undertaken during these first few years, the most important are the choice of the Italian television standard and the construction of a scanning unit.

621.397.5 : 519.272 **3121**

An Application of Autocorrelation Theory to the Video Signal of Television.—M. B. Ritterman. (*Sylvania Technologist*, July 1952, Vol. 5, No. 3, pp. 70–75.) Autocorrelation theory is used to analyse the power-density spectrum of a video signal. Statistical parameters representing motion and complexity in the picture are introduced. The problem is simplified by considering a signal representing only blacks and whites, but extension of the conclusions to cover intermediate greys is justified by reference to the work of Kretzmer (2904 of 1952).

621.397.5 : 519.272 **3122**

Application of Autocorrelation Theory to the Video Signal of Television.—M. B. Ritterman. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 201–207.) The power-density spectrum of the television video signal is derived, certain assumptions being made about the autocorrelation function.

621.397.5 : 535.623 **3123**

Colorimetry in Color Television.—F. J. Bingley. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 838–851.) Relations between colorimetric quantities and the electrical signals into which they are encoded for transmission are considered. The specific type of video signal envisaged comprises a monochrome component together with a colour carrier, though the colorimetric equations derived apply to any type of system.

621.397.5 : 535.623 **3124**

The National Television Systems Committee Color Television Transmission: Part 2.—R. M. Bowie & D. C. Livingston. (*Sylvania Technologist*, April 1952, Vol. 5, No. 2, pp. 36–40.) A mathematical treatment is given of the method of vestigial-sideband modulation of the colour sub-carriers described in part 1 [2343 of 1952 (Bowie & Tyson)], and the question of colour phase alternation is considered quantitatively.

621.397.5 : 778.5 **3125**

Reversible Process for Recording on and Electronic Reproduction from Cinematograph Films.—P. Mandel. (*Onde élect.*, April 1953, Vol. 33, No. 313, pp. 206–213.) Description, with discussion of the performance and limitations, of the flying-spot method, which can be used either to derive television pictures from cinematograph films or, conversely, to record such pictures on films.

621.397.61 **3126**

Television Transmitter at Wenvoe, Glamorganshire.—(*Engineering, Lond.*, 22nd August 1952, Vol. 174, No. 4517, pp. 249–250.) Description of medium-power equipment in operation pending installation of high-power transmitters.

621.397.61 : 656.135 **3127**

Mobile Television Transmitters.—E. W. Hayes. (*Wireless World*, Aug. 1953, Vol. 59, No. 8, pp. 372–375.) A description is given of the B.B.C. mobile transmitters which provided a television service for N.E. England, Northern Ireland, and the Brighton area much more quickly than would have been possible if permanent stations had to be constructed. The equipment, of a rugged type, is installed in a trailer 17 ft by 7 ft by 7 ft high. Duplicate vision transmitters, with peak-white output power $\sim$ 500 W are provided; the duplicate sound transmitters have a maximum carrier power of 200 W. The frequency of the vision signals is adjustable over the whole range of the present broadcasting band (41–68 Mc/s). The aerials used near Newcastle-on-Tyne and near Belfast were of the batwing type mounted on 250-ft lattice masts. The station on the downs near Brighton receives its vision signals direct from Alexandra Palace, using a Yagi aerial. The transmitters feed a vertically polarized dipole fitted with a reflector and mounted on a 70-ft mast. The Belfast station obtains its vision programme by direct reception of the Kirk o' Shotts transmissions until a permanent programme link with Britain can be provided.

621.397.611.2 **3128**

A Camera Tube with Photoconductive Layer: the Conductron.—B. Bartels & M. Munsch. (*Le Vide*, March 1953, Vol. 8, No. 44, pp. 1320–1325.) Description of the characteristics of the conductron in its present stage of development, with an indication of possible improvements. At present the internal photoelectric inertia is about 0·1 sec, which should be reduced below 0·04 sec if the tube is to be used for ordinary television. See also 1489 of 1951 (Veith).

621.397.62 : 535.623 **3129**

Theory of Synchronous Demodulator as used in N.T.S.C. Color Television Receiver.—D. C. Livingston. (*Sylvania Technologist*, July 1952, Vol. 5, No. 3, pp. 60–63.) The operation of the synchronous demodulator [see also 253 of January (McGregor)] is analysed, and possible distortion effects are studied. These may result from rectification of the colour subcarrier or from incomplete exclusion of the luminance signal. The analysis indicates that each of these may produce perceptible distortion. See also 3124 above.

621.397.62 : 535.623 **3130**

A 42-Tube Compatible Color Television Receiver.—K. E. Farr. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 208–217.) See 1501 of May.

621.397.62 : 535.623 **3131**

The PDF [post-deflection-focusing] **Chromatron — a Single- or Multi-gun Tricolor Cathode-Ray Tube.**—R. Dressler. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 851–858.) Red, green and blue phosphor strips, 15-mil wide, are deposited on the image plate and are backed with Al. Directly behind the image plate is a grid of parallel wires with one set of wires behind the red and another set behind the blue strips. The c.r. tube cone is made in two sections to permit insertion of the colour structure after the front plate has been sealed on. Varying the potential difference between the two sets of wires directs the beam on to one of the sets of coloured strips. A focusing and accelerating potential is applied between the wire grid and the Al coating. Colour distortion due to nonlinear sweeps or misalignment of gun position is absent. Analysis of operation is given, and a three-gun version is also described. The tube will operate with existing colour-television transmission systems.

621.397.621.2 : 621.385.832 **3132**

A Steel Picture-Tube for Television Reception.—J. de Gier, T. Hagenberg, H. J. Meerkamp van Embden, J. A. M. Smelt & O. L. van Steenis. (*Philips tech. Rev.*, April 1953, Vol. 14, No. 10, pp. 281–291.) The development of steel c.r. tubes and of the lead-free 'iron-glass' screen is described.

621.397.645 **3133**

Video Amplifiers for Optimum Transient Response.—W. K. Squires & H. L. Newman. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 218–230.) From Laplace-transform analysis of typical video coupling networks, relations among the various parameters resulting in optimum transient response are derived. A design procedure, based on these relations and on figures of merit for the valves used, is presented which is applicable to both triode and pentode amplifiers. Examination of response curves and tests of actual circuits show that the triode amplifier is definitely superior to the pentode amplifier when attainment of maximum voltage is the major criterion. In addition, the triode has certain advantages as regards the signal/noise ratio of the synchronizing signal.

621.397.743 **3134**

Television Coverage of the National Political Conventions.—R. W. Ralston & B. D. Wickline. (*Elect. Engng, N.Y.*, May 1953, Vol. 72, No. 5, pp. 383–389.) Detailed description of the wire and microwave links used for the 1952 Chicago Conventions.

TRANSMISSION

621.396.61.029.53/.55 **3135**

400-Watt H.F. Channellized Transmitter Type HC.100.—P. W. Jinkings. (*Marconi Rev.*, 2nd Quarter 1953, Vol. 16, No. 109, pp. 69–87.) Six editions of this self-contained equipment, which is suitable for use in the tropics, cover the range 1·6–27·5 Mc/s.

VALVES AND THERMIONICS

621.314.632 : 546.482.21 **3136**

Cadmium Sulfide as a Crystal Rectifier.—G. Strull. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 510–521.) Tests on CdS crystals produced for X-ray detection showed they were n-type semiconductors with both rectification and photoelectric properties. The optimum operating voltage for rectification is 6–8 V, but the range of voltage extends from 0·1 V to 40 V. The current at any given voltage can be varied by changing the intensity of illumination. The best performance is obtained with single crystals mounted so that the field is down the length of the crystal. The resistance of a typical crystal is $\sim$ 100 kΩ with 6 V applied in the forward direction and an illumination intensity of 30 lumens/ft^2.

621.383.2 **3137**

On the Fatigue of Ag-Cs Photoelectric Tubes.—G. Kuwabara. (*J. phys. Soc. Japan*, March/April 1953, Vol. 8, No. 2, pp. 229–233.)

621.383.2 **3138**

The Appearance of an Alternating Component in the Photocurrent of Photocells under Strong Illumination.—L. A. Vainshtein & L. P. Malyavkin. (*Zh. tekh. Fiz.*, Aug. 1952, Vol. 22, No. 8, pp. 1315–1317.) It has been observed that when an Sb-Cs photocell is illuminated by a powerful source of light an a.f. component appears in the photocurrent, sometimes together with a h.f. component of the order of several megacycles per second. Experiments are described and oscillograms are shown. It is suggested that these oscillations appear in the space charge, and that the presence of gas is necessary for this. In the case of vacuum cells, gas is apparently liberated as a result of the partial destruction of the photolayer by intense illumination.

621.383.27 **3139**

On the Development of Secondary-Electron Multipliers.—F. Eckart. (*Ann. Phys., Lpz.*, 16th Jan. 1953, Vol. 11, Nos. 4/7, pp. 181–202.) The secondary-emission properties of various materials and metal-based emissive layers are tabulated and the sensitivities of different photocathodes are noted. Different designs of multiplier are described and their noise and dark-current characteristics are discussed. The marked decrease in dark current of Ag-Cs_2O multipliers of mesh type is achieved by increasing the insulation path between the individual electrodes. 83 references.

621.383.4 **3140**

Action of a Stream of Electrons on the Current Sensitivity of Photocells.—B. P. Angelov, E. M. Lobanov & S. V. Starodubtsev. (*C. R. Acad. Sci. U.R.S.S.*, 1st Aug. 1952, Vol. 85, No. 4, pp. 733–735. In Russian.) A report of an investigation of the dependence of the current in the external circuit of Se and Ag_2S photocells on the energy (800–20 000 eV) and intensity of the primary electron stream.

621.383.4 **3141**

Tentative Interpretation of the Laws governing the Characteristics of Photocells.—G. Blet. (*J. Phys. Radium*, April 1953, Vol. 14, No. 4, pp. 241–250.) Analysis of the action of the blocking layer in a photocell enables a scheme to be developed which permits qualitative and quantitative verification of observations relating to the laws connecting illumination, voltage and current. Formulae derived give results in good agreement with experiment and permit calculation of the quantum output of a photocell.

621.383.4 : 535.33-1 **3142**

The Application of Lead Selenide Photoconductive Cells to Infra-red Spectroscopy.—V. Roberts & A. S. Young. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 199–200.)

621.385.012.6 **3143**

Investigation of the Fine Structure of Valve Characteristics.—H. Schneider. (*NachrTech.*, June 1953, Vol. 3, No. 6, pp. 274–279.) An electric current can be expressed in terms of a Taylor series with electric potential as the independent variable. If the differential coefficients are known, the series can be evaluated. Two methods of evaluating the differential coefficients, illustrated by considering the I_a/V_g characteristic of a triode, are given in detail and the experimental determination of the coefficients is described.

621.385.029.6 : 621.396.615.142.2 **3144**

Equations for the Oscillations in Uniform Electron Beams.—Yu. A. Katsman. (*Zh. tekh. Fiz.*, Sept. 1952, Vol. 22, No. 9, pp. 1467–1476.) The electrodynamic processes taking place in klystrons are discussed.

621.385.029.63/.64 : 621.396.822 **3145**

Noise Measurements on a Traveling-Wave Tube.—B. N. Agdur & C. G. L. Åsdal. (*Chalmers tek. Högsk. Handl.*, 1951, No. 106, 9 pp.) Reprint. See 1152 of 1952.

621.385.029.64/.65 **3146**

Experimental Observation of Double-Stream Amplification.—B. N. Agdur. (*Chalmers tek. Högsk. Handl.*, 1951, No. 105, 13 pp.) Reprint. See 1154 of 1952.

621.385.029.64 **3147**

The Resistive-Wall Amplifier.—C. K. Birdsall, G. R. Brewer & A. V. Haeff. (*Proc. Inst. Radio Engrs*, July 1953, Vol. 41, No. 7, pp. 865–875.) In the resistive-wall amplifier, the gain is obtained through interaction between the electron-stream charge and the wall charge induced by the stream. Increasing bunching of the electrons takes place and hence wave amplification results. In contrast with the travelling-wave valve, there is no interaction between the highly-damped circuit wave and the stream. Although some energy is dissipated in wall loss, there is a finite rate of growth for all stream currents. Theory is confirmed by experiments performed at 1–4 kMc/s on streams passed through glass tubes coated with tin oxide as the resistive material. Helices or resonant cavities were used as input and output coupling circuits. Gain and bandwidth are comparable to those of low-power travelling-wave valves. The resistive-wall amplifier is relatively insensitive to changes in circuit parameters or operating voltages, and there is no tendency to self-oscillation, in view of the high isolation between input and output. Amplifiers of this type with honeycomb or sandwich structure are expected to give better performance.

621.385.032.213.2 **3148**

Metal Capillary Cathodes.—H. Katz. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 597–603.) The operating conditions of cathodes basically similar to the L cathode [773 of 1951 (Lemmens et al.)] are discussed, and descriptions are given of five variants with working temperatures ranging from 1 500° to 850°C and pulsed emission from 3 to 5 A/cm². A low working temperature is advantageous, as surface evaporation is reduced. The best cathode is that with lowest evaporation and highest migration speed of the substance used for reducing the work function of the tungsten emitting surface. Substances other than tungsten, which may be better as regards low evaporation and high migration speed, may not maintain their porosity at the operating temperatures, but porous tungsten may serve as a base for surface layers with better properties. A considerable increase of emission was obtained from areas around marks made on a cathode surface with an ordinary pencil.

621.385.032.216 : 621.317.336 **3149**

Cathode-Interface Impedance and Its Measurement.—H. M. Wagner. (*Proc. nat. Electronics Conf., Chicago*, 1952, Vol. 8, pp. 553–561.) In valves with oxide-coated cathodes, failure is sometimes due to the formation of an interface impedance layer between the coating and its metal base. This type of failure occurs more frequently when valves are operating in circuits involving periods of anode-current cut-off [see 1526 of 1951 (Waymouth)]. The effect of interface impedance is approximately equivalent to the insertion of a resistor shunted by a capacitor in the cathode lead. The effect is studied for various pulse waveforms and a method for determining the value of the impedance in any particular valve is described which makes use of rectangular voltage pulses. The measurement of low values of impedance is made possible by use of a difference amplifier; this produces an output signal that represents the difference between the voltage across the load of the valve under test and a variable fraction of the input-pulse voltage.

621.385.032.216 : 621.396.822 **3150**

Temperature Dependence of Low-Frequency Fluctuations in Thermionic Emitters.—T. B. Tomlinson. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 611–615.) A method is described for measuring the temperature variation of the flicker effect in diodes with oxide-coated cathodes. The results show that in most valves the fluctuations of emission are minimum for a cathode temperature near 1 000°K. Two separate effects appear to be involved, one with a rising and the other with a falling temperature characteristic. Various theories of the effect are discussed; none is found completely satisfactory. See also 281 of January.

621.385.032.216.1 : 546.841.4-31 **3151**

Measurements relating to the Filament of Directly Heated Thoria Cathodes.—G. Mesnard. (*Le Vide*, March 1953, Vol. 8, No. 44, pp. 1326–1331.) The variations of heating current in the tungsten wire carrying the thoria coating, and of the wire resistance and applied power, are shown graphically as a function of temperature to indicate the effect of various heat treatments of the cathodes. The effect of the thickness of the thoria coating, and its relation to the spectral emissivity in the red region and to the thermionic emission, are also discussed. See also 2506, 2507 and 2508 of August.

621.385.032.216.2 **3152**

Double Activators in Oxide-Cathode Base Alloys.—A. Eisenstein, H. John & J. H. Affleck. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 631–632.) The existence of Si as a minor impurity in a W-Ni cathode base metal with 4·7% W was found to result in the formation, towards the end of a 6 000-hour life test, of a high-resistance Ba_2SiO_4 interface compound, the Si replacing W in the Ba_3WO_6 interface layer formed at first. Similar reactions between Ti, Al and other interface compounds were examined.

621.385.23 **3153**

Squirrel-Cage Filament Structures.—A. M. Hardie. (*Wireless Engr*, Aug. 1953, Vol. 30, No. 8, pp. 196–204.) "The solution of problems of space-charge-limited current flow is discussed with particular reference to the squirrel-cage filament structure commonly employed in high-power valves. Experiments are described which were made to determine the solid-cathode diameter equivalent to squirrel-cage structures containing an even number of wires up to 16. An example is given of the large error caused in the estimation of valve characteristics by an optimistic assumption of cathode diameter."

621.387 **3154**

Arc Drop and Deionization Time in Inert-Gas Thyratrons.—S. Pakswer & R. Meyer. (*J. appl. Phys.*, May 1953, Vol. 24, No. 5, pp. 501–509.)

621.387 : 621.316.722 **3155**

Variations in the Characteristics of Some Corona-Stabilizer Tubes.—F. A. Benson & J. P. Smith. (*J. sci. Instrum.*, June 1953, Vol. 30, No. 6, pp. 192–194.) Measurements made on six 500-V tubes are reported. The effects of ambient-temperature changes and of excess current are noted. An indication is given of the drift of running voltage to be expected during the first 1 200 hours of life. The properties of the tubes when operated with reversed polarity are discussed.

621.387 : 621.395 **3156**

New Thyratrons for Telephone-Circuit Engineering: Part 1 — Introduction.—K. L. Rau. (*Frequenz*, May 1953, Vol. 7, No. 5, pp. 134–139.) Static characteristics of the simple, coincidence and anticoincidence types of thyratron are discussed and the effects of grid resistance, grid-cathode spacing and gas pressure are shown graphically. The operation of these thyratrons is compared with that of relays.

621.396.615.141.1 **3157**

Motion of Electrons in Retarding-Field Tubes under Optimum Receiving Conditions.—M. Hirashima. (*J. phys. Soc. Japan*, March/April 1953, Vol. 8, No. 2, pp. 182–193.) Measurements are described of the anode-current distribution in a valve with tungsten-filament cathode, spiral grid, mesh anode, and a collecting electrode, arranged coaxially. The energy distribution of electrons flowing into the mesh anode was also determined. A signal of frequency near 550 Mc/s, modulated at 1 kc/s, was applied to grid and anode via a Lecher-line system.

621.396.615.141.2 **3158**

Theory of the Multisegment Magnetron.—L. E. Bakhrakh. (*Zh. tekh. Fiz.*, June 1952, Vol. 22, No. 6, pp. 1008–1015.) The phase focusing of an electron beam passing consecutively through a number of phase lenses is investigated and the results obtained are used for explaining the maintenance of oscillations in a multi-segment magnetron.

621.396.615.142.2 **3159**

Experimental Investigation of the Oscillation Mode and Bunching Efficiency of a Reflex Klystron.—M. de Bennetot. (*Ann. Radioélect.*, April 1953, Vol. 8, No. 32, pp. 103–110.) The method of determining the oscillation mode, previously described by Hamilton (3570 of 1948), is illustrated. From a law of similitude in electron optics which takes account of space charge, two relations are deduced which show how the phase of the modulated current varies with slight change of (*a*) the accelerating voltage, (*b*) the reflector voltage. These relations enable the bunching efficiency to be determined. Results of measurements on a Type-KR331 klystron are shown graphically. Both the oscillation mode and the bunching efficiency must be known if a satisfactory explanation of the operation of the reflex klystron is to be given, and, since the combination of curved reflector, space charge and drift space is not very amenable to mathematical treatment, experimental determination of the bunching efficiency is particularly useful for obtaining design data.

621.396.615.142.2 **3160**

A Coupled-Resonator Reflex Klystron.—E. D. Reed. (*Bell Syst. tech. J.*, May 1953, Vol. 32, No. 3, pp. 715–766.) Output-power/repeller-voltage and frequency/repeller-voltage characteristics suitable for various applications of variable-frequency oscillators are obtained by means of a reflex klystron having two coupled resonators instead of the usual single resonator. The theory of the system is developed and an experimental valve and circuit are described. The characteristics can be adjusted to suit a particular use by varying the coupling between the resonators and the ratio of their Q values. This advantage is gained at the expense of using greater power to supply the losses in the second resonator.

621.396.615.142.2 : 621.384.622.1 **3161**

High-Power Klystron Amplifier.—J. Jasberg. (*Electronics*, July 1953, Vol. 26, No. 7, pp. 244–248.) Description of a gridless three-cavity klystron with oxide-coated indirectly-heated cathode, designed for the 1 000-MeV linear electron accelerator at Stanford University. Beam spreading is prevented by focusing coils surrounding the cavities. Pulses of 2-μs duration at a repetition frequency of 60/sec are used, and peak output power is 17 MW at 10·5-cm wavelength.

MISCELLANEOUS

53.08 : 061.4 **3162**

Industrial Instrumentation.—(*Elect. Rev., Lond.*, 3rd July 1953, Vol. 153, No. 1, pp. 3–13.) A survey of instruments exhibited at the British Instrument Industries Exhibition at Olympia, London, 30th June–11th July 1953.

621.39 : 061.4 **3163**

20th National Radio Exhibition.—(*Wireless World*, Sept. 1953, Vol. 59, No. 9, pp. 397–406.) Classified guide to the principal exhibits.

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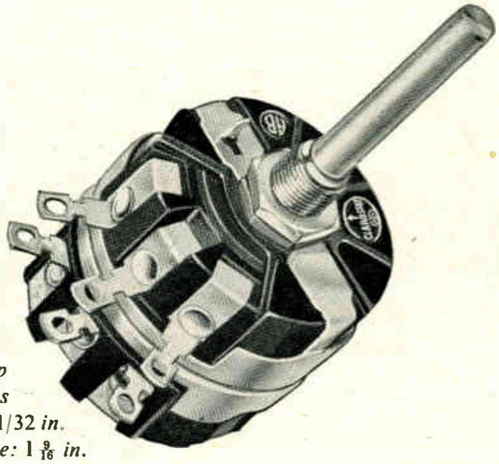

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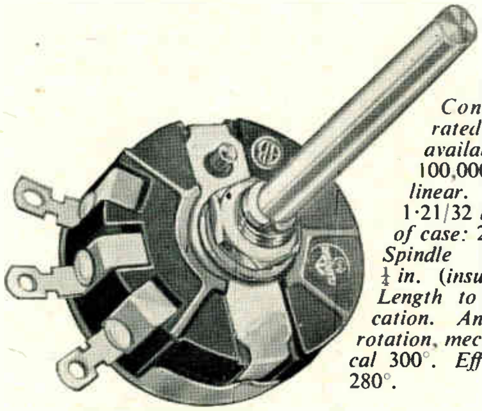

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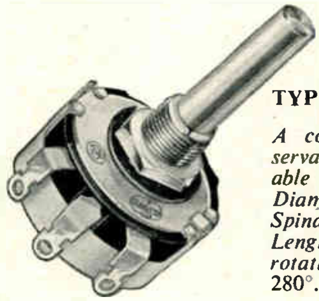

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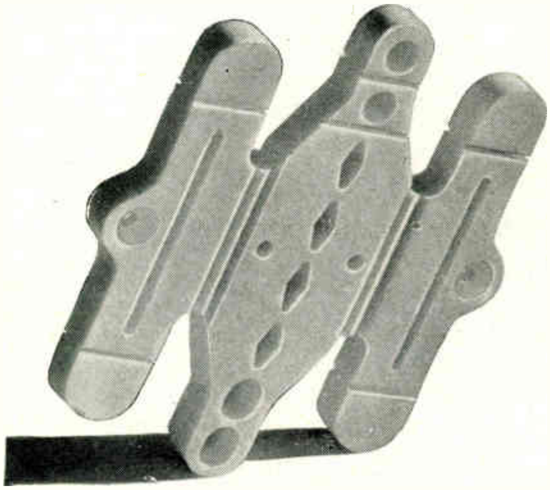

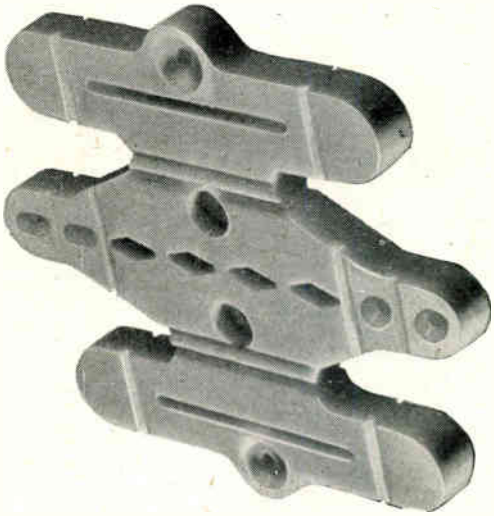

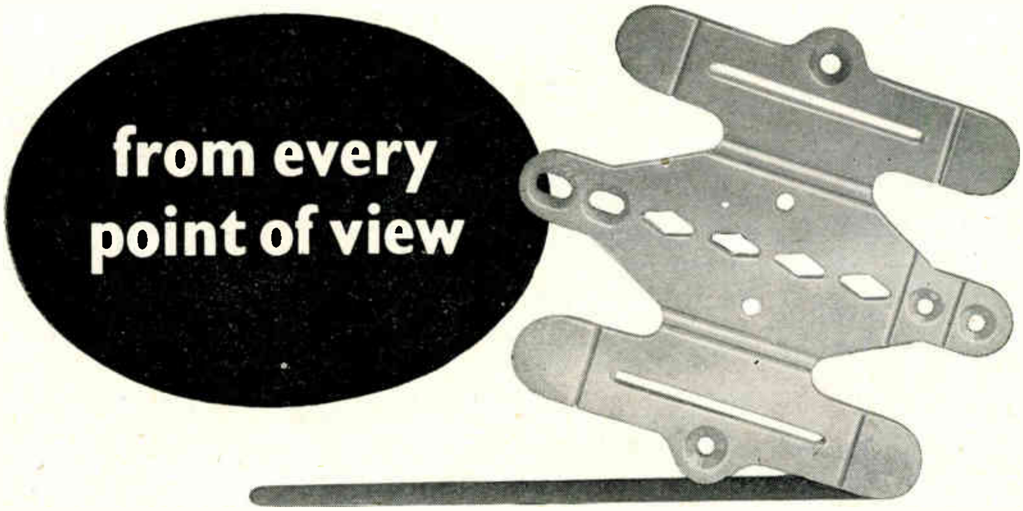

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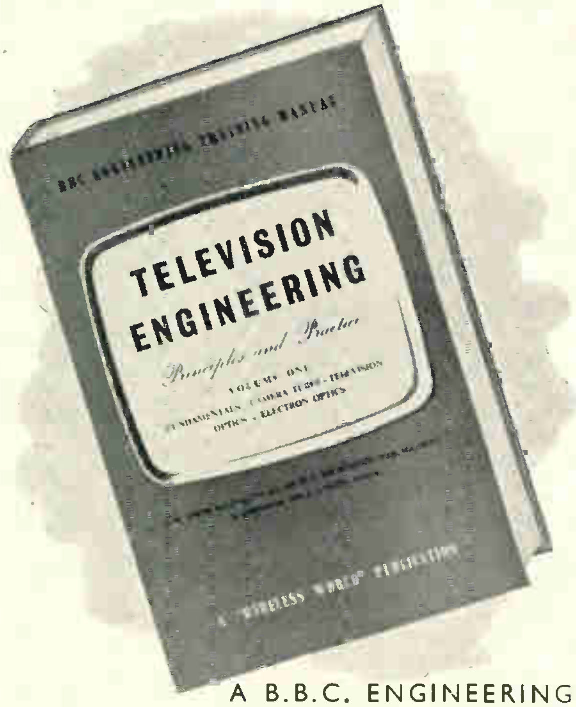
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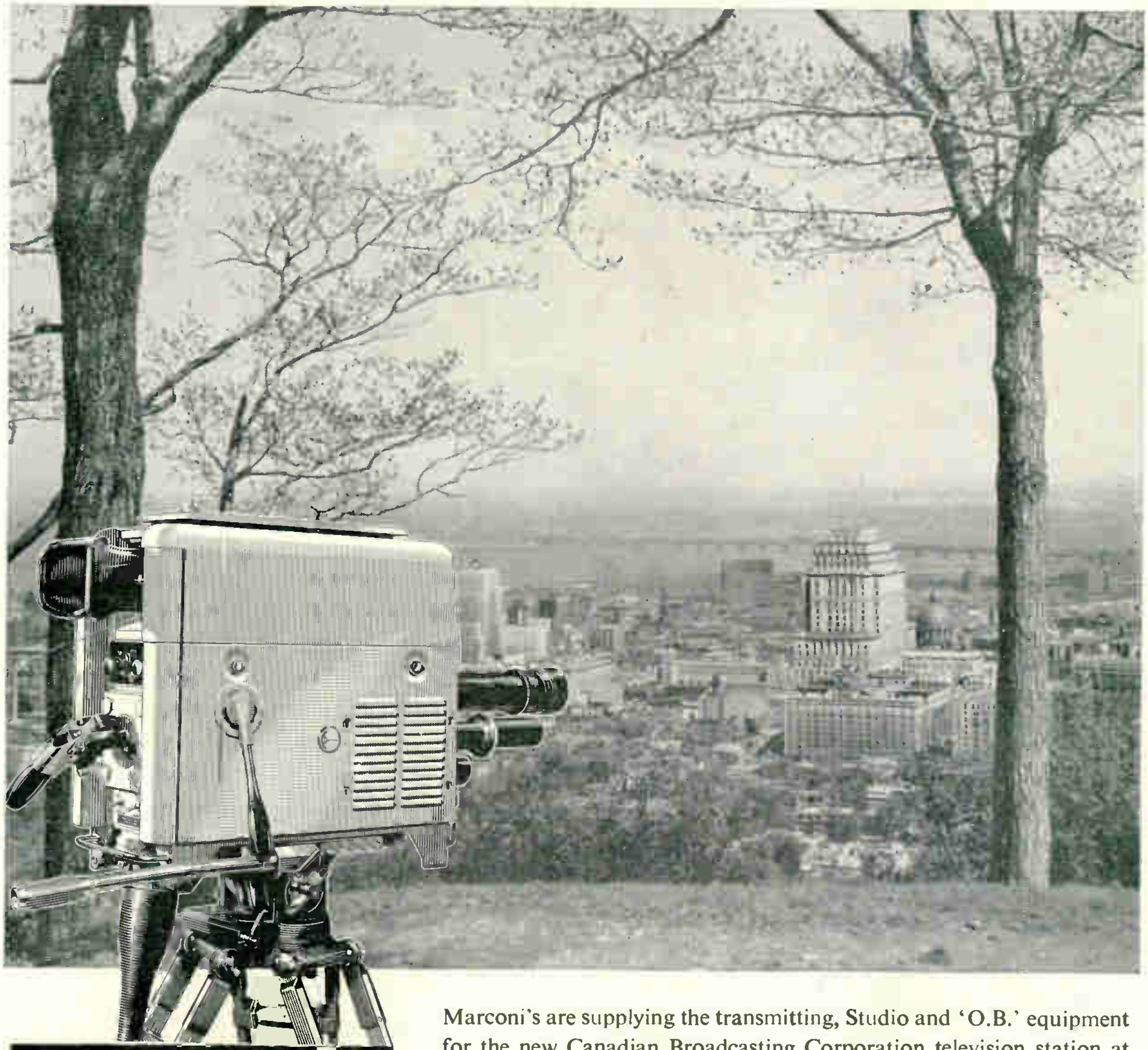

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University Graduates in Electrical Engineering, Physics or Chemistry are offered employment in the Capacitor Laboratory of a large firm of communications engineers in East London. Previous experience is not necessary and the field covered includes fundamental work on dielectric and capacitor design. Consideration will be given to London University Graduates who are eligible and Graduates of the Institution of Electrical Engineers. Applications should be addressed to Personnel Dept., Standard Telephones & Cables Ltd., Woolwich, London, E.16.

Research Assistant required, to work on Psycho-Acoustics and related studies of Speech Analysis. £400 p.a. Research experience essential, though not necessarily in this field. Apply to E. C. Cherry, Electrical Engineering Department, Imperia College, S.W.7.

An Engineer is required for the layout and development, to the prototype production state, of electronic equipment of a new type. Applicants, who must be British born, should preferably have had experience of work of this kind in the radio, radar or communications field and should have a degree or suitable alternative training and experience. The post will carry a starting salary in accordance with age and experience. Write giving full details to the Personnel Officer, Ericsson Telephones Limited, Beeston, Nottingham.

The Parsons and Marine Engineering Turbine Research and Development Association requires a Research Engineer for the development of specialised measuring apparatus used in connection with research on steam and gas turbine machinery. The applicants should have an Honours degree in Electrical Engineering or Physics, and should have industrial and laboratory experience in electronic techniques as applied to mechanical engineering. A good salary will be paid, depending upon age, qualifications and experience. Candidates should apply in writing to the Research Director, Pametrada Research Station, Wallsend-on-Tyne.

Belling & Lee, Ltd., Cambridge Arterial Road, Enfield, Middlesex, require research assistants in connection with work on electronic components, fuses, interference suppressors and television aerials. Applicants must be graduates of the I.E.E. or possess equivalent qualifications together with similar laboratory experience. Salary will be commensurate with previous experience: five-day week, contributory pension scheme. Applications must be detailed and concise, and will be treated as confidential.

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A number of vacancies exist in the Electronics Division of our Research Laboratories for work on Electronic Switching and Computing. Applicants for some posts should have a good degree and for others a minimum of Higher National and preferably several year's electronic experience. Write giving full details to the Personnel Officer, Ericsson Telephones Limited, Beeston, Nottingham.

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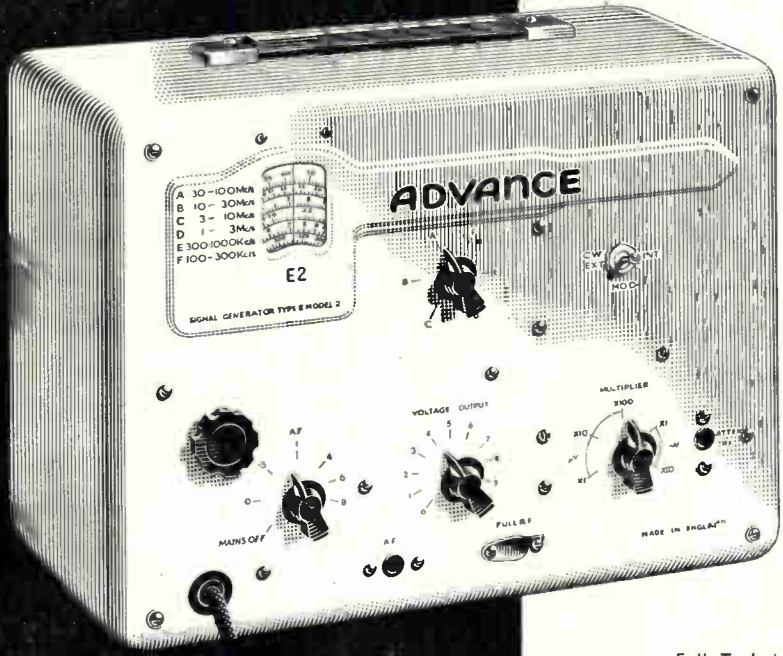

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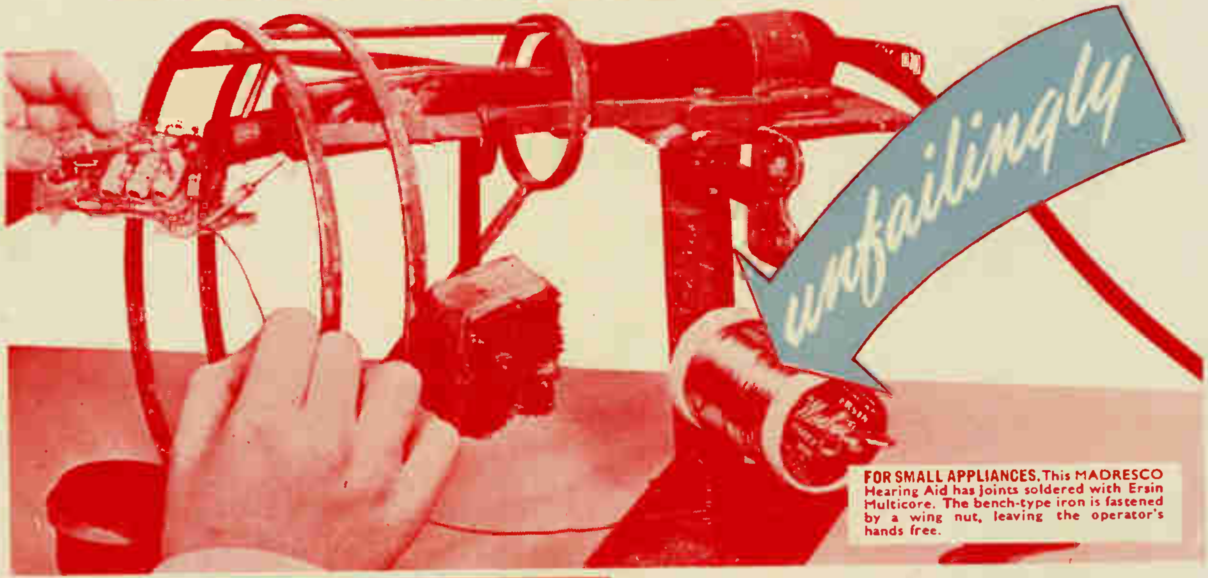
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