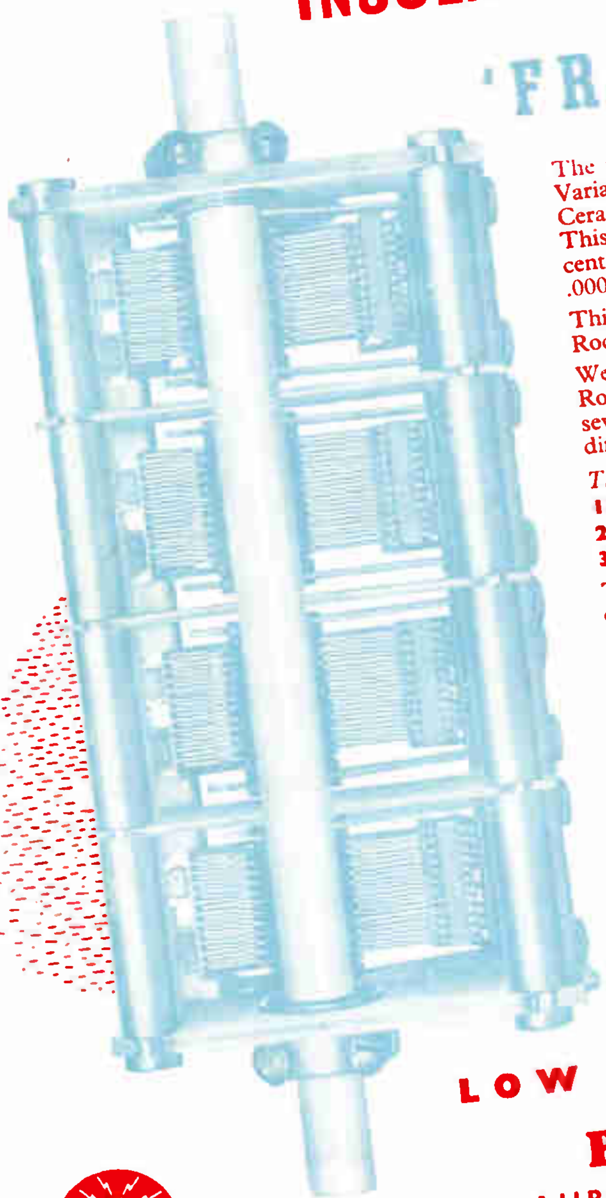
WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

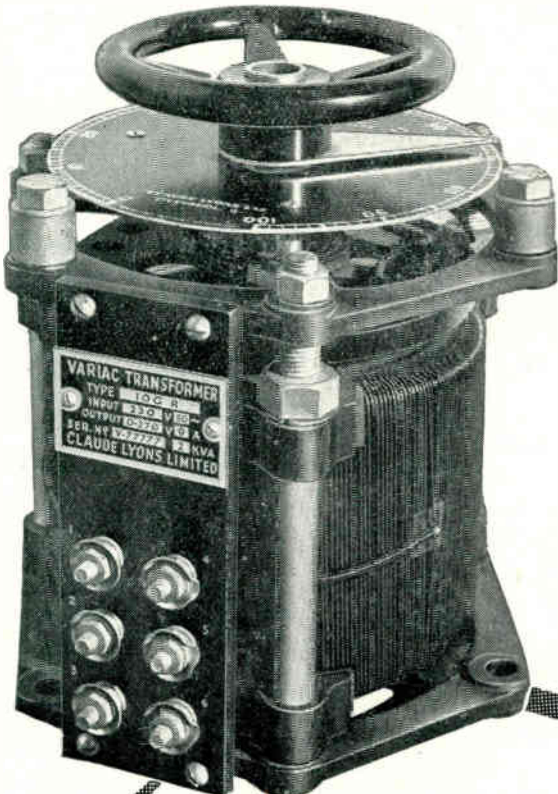
MARCH 1953

VOL. 30 No. 3 THREE SHILLINGS AND SIXPENCE



'VARIAC' voltage regulating transformers

Reg'd Trade Mark



'VARIAC' infinitely variable voltage-regulating transformers are designed and constructed for many years of trouble-free operation. Not only do they supply perfectly smooth control of voltage from zero, but many types (including 100-R) also furnish output voltages considerable above line voltage while the 100-RM/2B is a dual output type. Illustration left shows Type 100-R, Rating 2,000 va., Input Voltage 230/115, Output Voltage 0-270.

Left—
Type 200 C.U.H.

Right—
Type 50-B.

SERIES 100 'VARIAC' TRANSFORMERS.

Type	Load Rating	Specifications: Input Voltage	Current: Rated	Current: Maximum	Output Voltage	No-Load Loss	Net Price ‡ £ s. d.
100-K	2000 va.	115	15 a.	17.5 a.	0-115	20 watts	17 17 0
100-KM	2000 va.	115	15 a.	17.5 a.	0-115	20 watts	18 12 0
100-L	2000 va.	230/115	8 a.	9 a.	0-230	25 watts	17 17 0
100-LM	2000 va.	230/115	8 a.	9 a.	0-230	25 watts	18 12 0
100-Q	2000 va.	115	15 a.	17.5 a.	0-135	20 watts	18 9 0
100-QM	2000 va.	415	15 a.	17.5 a.	0-135	20 watts	19 4 0
100-R	2000 va.	230/115	8 a.	9 a.	0-270	30 watts	18 9 0
100-RM	2000 va.	230/115	8 a.	9 a.	0-270	30 watts	19 4 0
100-LH	1200 va.	480/240	2 a.	2.5 a.	0-480	25 watts	21 15 0
500-L	1450 va.	180	8 a.	9 a.	0-180	25 watts	17 17 0
2000-K	1000 va.	125	8 a.	9 a.	0-125	25 watts	17 17 0

* For 500 cycles † For 2,000 cycle service ‡ All 'VARIAC' prices plus 20% as from 23rd Feb. 1952

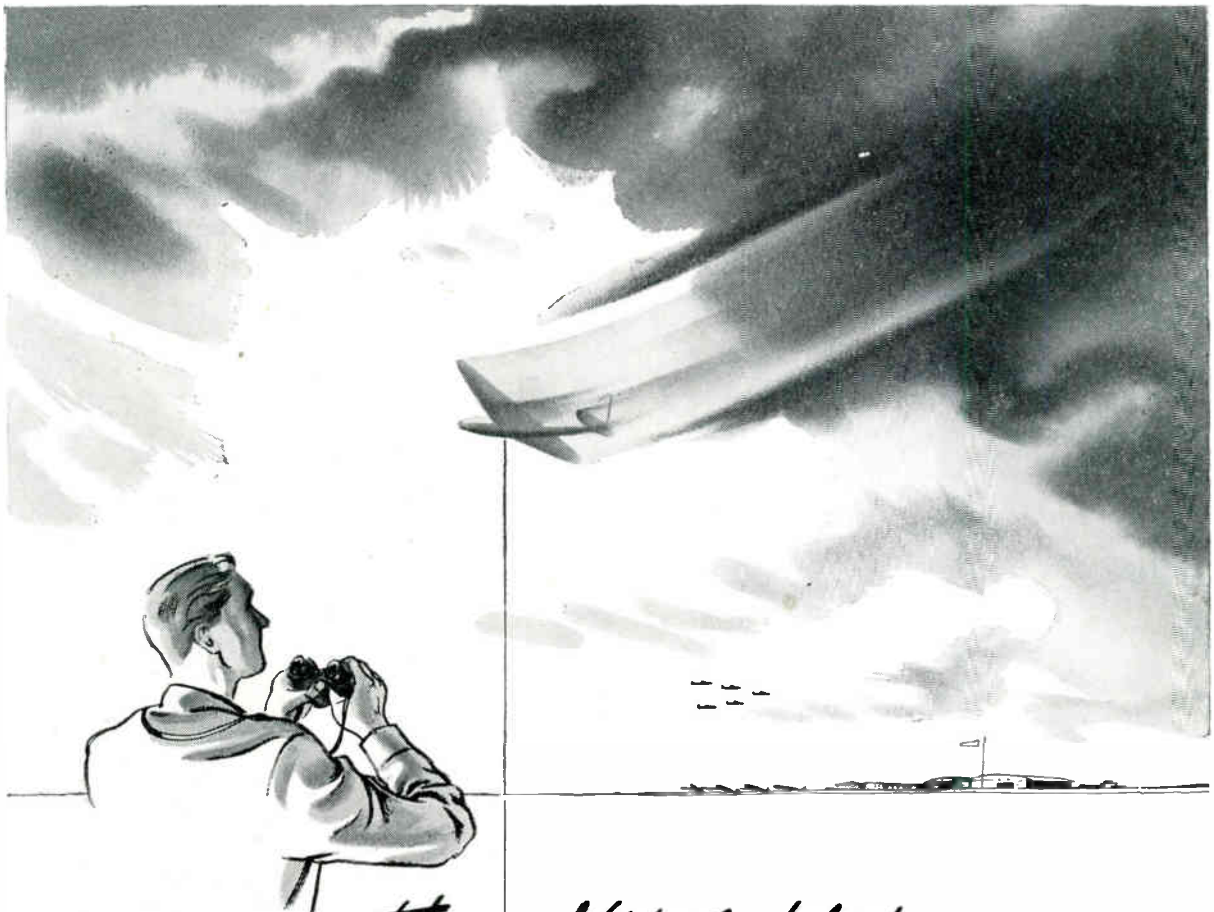
PROMPT DELIVERY

Write for catalogue V549 which gives full details of 'VARIAC' transformers and suggestions for use.

CLAUDE LYONS LIMITED

ELECTRICAL AND RADIO LABORATORY APPARATUS, ETC.

180 Tottenham Court Road, London, W.1; and 76 Oldhall Street, Liverpool 3, Lancs.



The Vital Link . . .

. . . upon which the pilot's safety depends is the consistent functioning of his radio and radar equipment. One dry or H.R. joint among the thousands of soldered joints in these precision instruments may mean the failure of a whole circuit endangering both pilot and machine.

FAULTLESS FLUXING PRESERVES THE VITAL LINK

Dry or H.R. joints are impossible with Superspeed for the flux is always released in exactly the correct proportion. This faultless fluxing action is achieved by the unique STELLATE core which gives six points of rapid solder collapse. At soldering temperature the activated rosin flux is released immediately for effective spreading and wetting. Superspeed is being used more and more in the production of service equipment where faultless joints are essential.

Superspeed

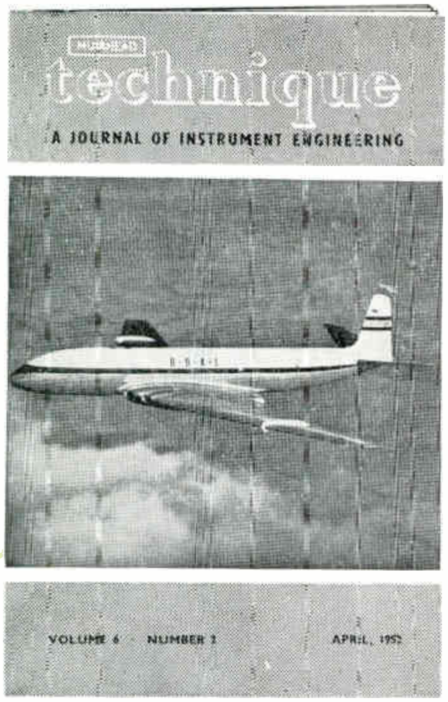
"WHITE FLASH" ACTIVATED ROSIN-CORED SOLDER

for general electrical, electronic and telecommunication work and all standard uses. A.I.D. and G.P.O. approved. Complies with M.O.S. Specification DTD 599. In all standard tin/lead alloys, 10-22 s.w.g. Sample of Superspeed and the comprehensive Superspeed booklet gladly sent on request. Technical advisers are available for free consultation.

MANUFACTURED BY THE **ENTHOVEN** *GROUP*

* Marketed by Enthoven Solders Ltd., Enthoven House, 89, Upper Thames Street, London, E.C.4. *

E3405



THIS INTERESTING QUARTERLY WILL BE MAILED *FREE* TO ANY PART OF THE WORLD

POST THIS COUPON TODAY

MUIRHEAD & CO. LTD · BECKENHAM · KENT · ENGLAND

Please mail me, free of charge, your quarterly journal "technique"

NAME ____________________

POSITION ____________________

COMPANY ____________________

ADDRESS ____________________

____

____________________ 76

MUIRHEAD & CO. LTD · BECKENHAM · KENT · ENGLAND

PRECISION ELECTRICAL INSTRUMENT MAKERS

A McGRAW-HILL BOOK PUBLISHED IN LONDON

PRINCIPLES OF RADAR

Third Edition

By J. F. REINTJES & G. T. COATE

A very thorough revision and expansion of the text originally prepared by the Massachusetts Institute of Technology Radar School Staff for use in their war training courses. Dealing with the fundamental concepts and techniques of pulse radar, it presents the engineering principles of the pulse circuits and high-frequency devices common to most radar systems. This exceptionally well illustrated book is now printed in a clearer type-face, and is outstanding in its clarity of expression.

985 pages 9 × 6 55s. 6d.

Obtainable from all booksellers

McGRAW-HILL

McGRAW-HILL PUBLISHING CO. LTD

McGraw-Hill House, 95 Farringdon Street, London, E.C.4

ALL-POWER

REGULATED POWER SUPPLIES

SERIES 500

An entirely new range of units, designed for the highest possible performance and overall efficiency.

Now entering into large-scale production and available for prompt delivery.

MODEL 501
(Fitted with end frames)

ABRIDGED DATA
(Further information on request)

	Data	Model 501	Model 502	Model 503	Model 504	Model 505
Main + VE. Stabilizer	Output	200-500 V 250 mA	200-500 V 250 mA	0-500 V 250 mA	0-500 V 250 mA	0-500 V 150 mA
	Number of Ranges	2	2	3	3	1
	Voltage Stabilization	±0·02%	±0·002%	±0·1%	±0·002%	±0·1%
	Effective output Resistance (max.)	0·2Ω	0·002Ω	0·5Ω	0·02Ω	0·5Ω
	Output Ripple (rms. max.)	2 mV	1 mV	3 mV	1 mV	3 mV
— VE. Supply Stabilizer	Outputs	—	—	250V 5mA 0-250V 1mA	250V 5mA 0-250V 1mA	250V 5mA 0-250V 1mA
	Voltage Stabilization	—	—	±0·05%	±0·002%	±0·05%
	Output Resistance (max.)	—	—	1Ω	0·01Ω	1Ω
	Output Ripple (rms. max.)	—	—	2mV	1mV	2mV
Unstabilized + VE. H.T. supply 250mA max.		470V 630V	470V 630V	320V 470V 630V	320V 470V 630V	630V
Unstabilized A.C. supply		6·3V 10A	6·3V 10A	6·3V 10A	6·3V 10A	6·3V 10A
Price		**£70**	**£91**	**£81**	**£100**	**£75**

STANDARD UNITS—All models are supplied as standard for mounting in 19" racks and are fitted with fully protective covers.

EXTRAS—To convert from rack mounting to bench use the following extras are available:—

POLISHED HARD-WOOD REINFORCED END FRAMES .. £1 15s. 0d. per pair

STEEL INSTRUMENT CASE OF NEW DESIGN .. £4 10s. 0d. each

PRICES—Prices are quoted net ex works and are subject to variation without notice

ALL-POWER TRANSFORMERS LTD. **CHERTSEY ROAD, BYFLEET, SURREY**
TEL.: BYFLEET 3224/5

MARCONI communication systems

serve mankind

Wherever men live together (in industrial areas or in tropical archipelagoes) prosperity is largely a matter of communications.

Island communities seem blessedly self-contained. But, for effective government, hygiene, safety, husbandry, commerce — welfare, in short, and prosperity — archipelagoes are increasingly interdependent. Marconi's invention is bringing to island peoples a fuller and richer way of life.

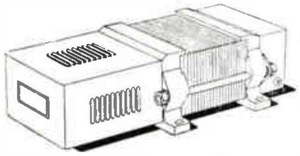
SURVEYED

PLANNED

INSTALLED

MARCONI'S WIRELESS TELEGRAPH COMPANY LTD · CHELMSFORD · ESSEX

constant light source voltage



Constantly reliable readings or indications of any photometric device require a *constant* light source, which in turn demands *constant voltage*. Advance Constant Voltage Transformers keep the mains supply voltage steady to within ± 1% with input variations of up to ± 15%.

Full details in Folder S.15/V available on request.

Can now be supplied with multiple secondary windings to provide stabilisation on a number of outputs. A specification of your requirements may enable us to provide you with CONSTANT VOLTAGE throughout your equipment.

Advance

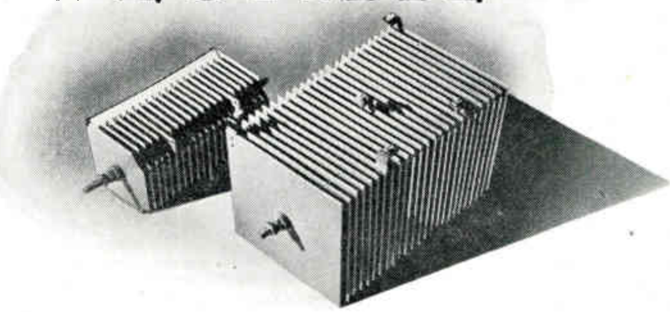
ADVANCE COMPONENTS LTD., BACK ROAD, SHERNHALL STREET, WALTHAMSTOW, LONDON, E.17

Telephone : LARkswood 4366/7/8 *Grams* : Attenuate, Walt, London

---

G.E.C. SELENIUM RECTIFIERS

Full particulars of the G. E. C. range of Selenium Rectifiers are available on application to your nearest G.E.C. Branch, E.S. & V. Department, Magnet House, Kingsway, London, W.C.2, or the address below—



LOW FIRST COST — *MAXIMUM STABILITY* — *LONG LIFE*

HIGH EFFICIENCY — *NO MAINTENANCE* — *COOL RUNNING*

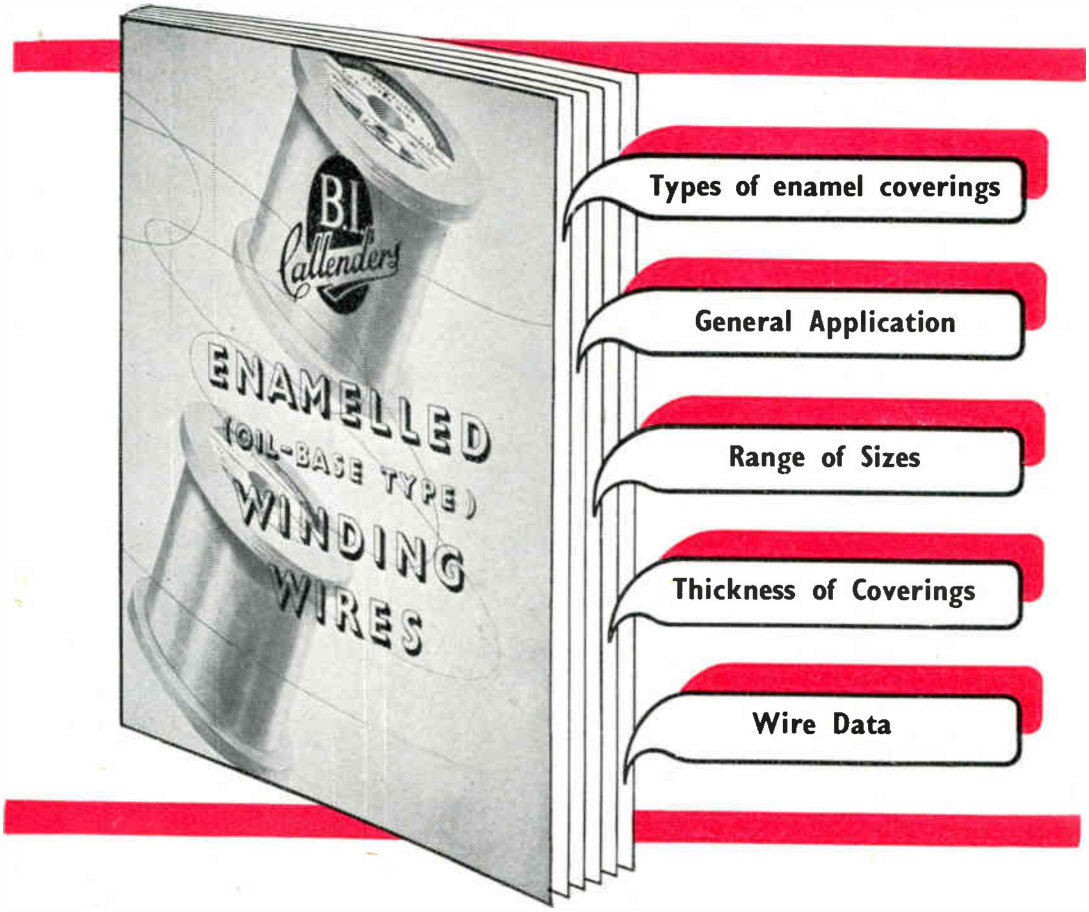
SALFORD ELECTRICAL INSTRUMENTS LTD · SALFORD 3 · LANCS

A SUBSIDIARY OF THE GENERAL ELECTRIC CO. LTD OF ENGLAND


this
will
fix
it
. . . fix it better,
more quickly and
neatly.
Unbrako Socket Screws
have extra strength
and toughness.
They are made of
high-tensile nickel-chrome
steel alloy.
They stay put
until you want to remove them.
Then—out they come,
smoothly, easily, without damage.
No wonder so many designers and engineers specify Unbrako!
Available for immediate delivery.
Send for free samples.
UNBRAKO SOCKET SCREW COMPANY LTD.
Manufacturers of Unbrako standard socket screws and special screws to A.I.D. requirements.
COVENTRY
ENGLAND
UNBRAKO

HERE'S USEFUL INFORMATION ON

BICC ENAMELLED WINDING WIRES



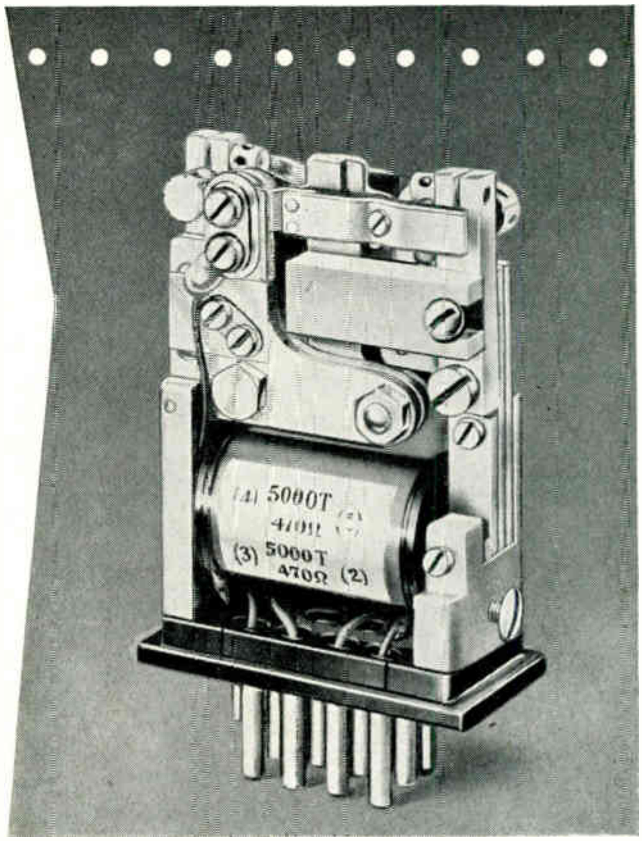
If you use enamelled winding wires for electrical purposes, send for a copy of this publication. It contains essential information on the range of BICC oil-base enamel covered winding wires, and includes tables and other useful data.

Send for Publication No. 303

BRITISH INSULATED CALLENDER'S CABLES LIMITED
NORFOLK HOUSE, NORFOLK STREET, LONDON, W.C.2

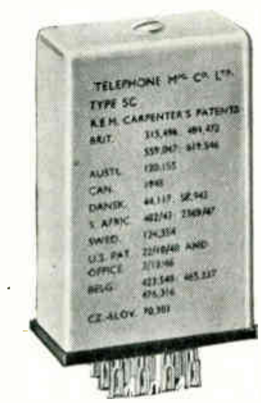
sturdy and strong-

The compact, rugged design of the Type 5 Carpenter Polarized Relay is just one of the many reasons why this small-size relay is so frequently specified in many fields of scientific research and electrical engineering.



Other outstanding features include :—

HIGH OPERATIONAL SPEED
FREEDOM FROM CONTACT REBOUND
IMMUNITY FROM POSITIONAL ERROR
GOOD CONTACT PRESSURES
HIGH SENSITIVITY
ACCURACY OF SIGNAL REPETITION
EXCEPTIONAL THERMAL STABILITY



Type 5 Relay with cover.

Plug or solder tag base optional.

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

TMC

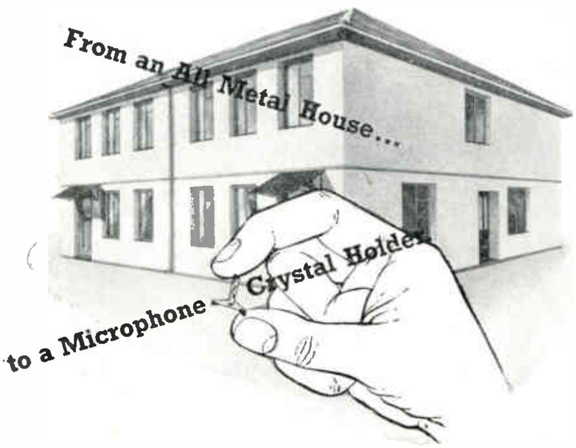
CARPENTER POLARIZED RELAY TYPE 5

Manufactured by the Sole Licensees:—

TELEPHONE MANUFACTURING CO. LTD

Contractors to Governments of the British Commonwealth and other Nations.

HOLLINGSWORTH WORKS · DULWICH · LONDON S.E.21 · *Telephone: Gipsy Hill 2211 (10 lines)*



Success in our business depends largely on versatility, and we think we have achieved it. Our range of products is a wide one—**from an All-Metal House to a Microphone Crystal Holder.**

Our Research and Development Department is available to the trade for assisting in the design of products to give the most economic method of production, and finishing, from our self-contained factory.

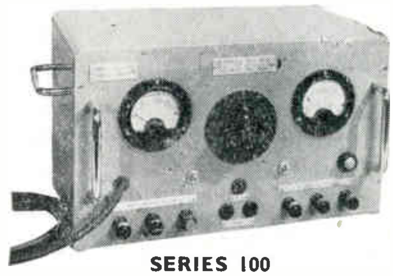
Give us the opportunity of quoting for your pressings and fabrications.

ADAMS BROS. & BURNLEY LTD.

Elmgrove Road, Harrow, Middlesex.

Tele: Harrow 6411 (5 lines)

POWER SUPPLY UNIT



SERIES 100

DC Output, Stabilised, 200-350V 120mA, continuously variable.

AC Output, Unstabilised 0-4-6·3V 3·0A.

Bench Model, as illustrated ... ... £58.
Rack Mounting Model ... ... £59.

Delivery ex-stock.

Available with negative bias supply at £63 and £64.
Power Supply Units and other special electronic equipment made to customers' specifications.

Pamphlet and full particulars from :—

HARVEY ELECTRONICS LTD.

Precision Mechanical and Electrical Engineers

273, FARNBOROUGH ROAD, FARNBOROUGH, HAMPSHIRE. *Telephone:* FARNBOROUGH 1120

Radio and Radar Technique

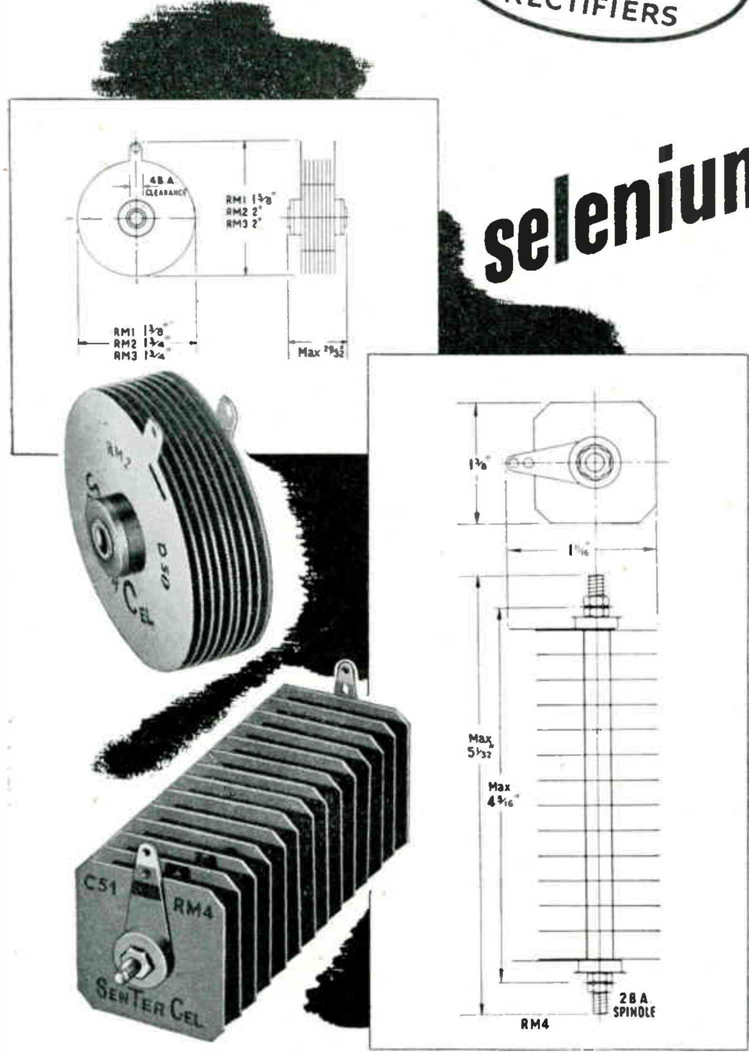
by A. T. Starr
M.A., Ph.D.
M.I.E.E.

The author is well-known as an outstanding worker in the field of radio and radar. In this important new book he surveys present-day knowledge of the essential methods and techniques, paying particular attention to noise, microwave techniques, waveforms, pulse circuit techniques, and electronic tubes. A first-rate book for designers and other advanced workers. Profusely illustrated. 830 pages. 75s. net.

Pitman *Parker Street, Kingsway, London, W.C.2*

10 advantages of SELENIUM SenTerCel RECTIFIERS miniature selenium rectifiers

REGISTERED TRADE MARK



Compare these outstanding features with those of the rectifiers which at present you are using :—

- **Less wiring**
- **Unlimited instantaneous overload such as the charging current of de-formed electrolytic capacitors.**
- **Far lower heat dissipation.**
- **No " warming-up " period.**
- **No valve-holder.**
- **Practically indestructible in normal service.**
- **No limit to size of electrolytic capacitor.**
- **Saves weight.**
- **Saves space.**
- **Low in cost.**

Study these **RATINGS**

TYPE	RM1		RM2		RM3		RM4		
Maximum ambient temperature	35°C	55°C	35°C	55°C	35°C	55°C	35°C	40°C	55°C
Maximum output current (mean)	60mA	30mA	100mA	60mA	120mA	90mA	275mA	250mA	125mA
Maximum input voltage (r.m.s.)	125V		125V		125V		250V		
Maximum peak inverse voltage	350V		350V		350V		700V		
Max. instantaneous peak current	Unlimited		Unlimited		Unlimited		Unlimited		
Weight	1 oz.		1.4 oz.		2 oz.		4.5 oz.		

REGISTERED TRADE MARK
SELENIUM SenTerCel RECTIFIERS

Standard Telephones and Cables Limited

(Registered Office : Connaught House, Aldwych, W.C.2)

RECTIFIER DIVISION: Warwick Road, Boreham Wood, Hertfordshire.
Telephone : Elstree 2401 Telegrams: Sentercel, Borehamwood

WIRELESS ENGINEER

The Journal of Radio Research and Progress

Managing Editor: HUGH S. POCOCK. M.I.E.E.
Editor: W. T. COCKING, M.I.E.E. Technical Editor: Professor G. W. O. HOWE, D.Sc., LL.D., M.I.E.E.

Volume 30 · Number 3

CONTENTS

MARCH 1953

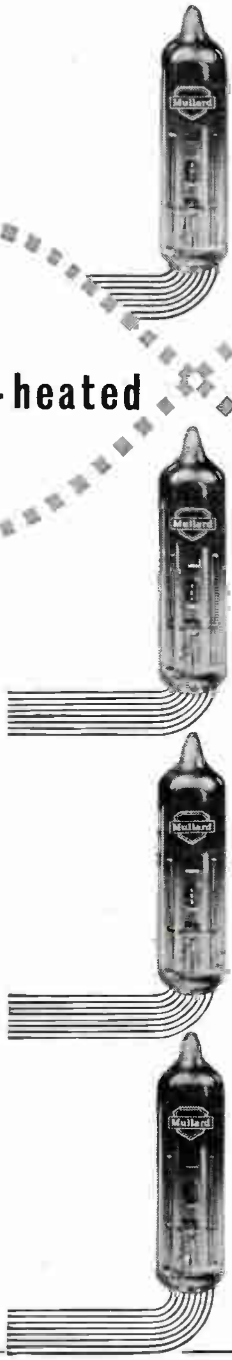
Published on the sixth of each month

Annual Subscription: Home and overseas, 1 year £2 4s. 6d.; 6 months £1 2s. 3d.; Canada and U.S.A. $7.00

Editorial, Advertising and Publishing Offices: Dorset House, Stamford Street, London, S.E.1
Telephone: Waterloo 3333 (60 *lines*) • *Telegrams: Wirenger, Sedist, London*

BRANCH OFFICES AT BIRMINGHAM · MANCHESTER AND GLASGOW

AVAILABLE SOON FOR COMMUNICATIONS AND INDUSTRIAL EQUIPMENT



indirectly-heated *subminiatures*

These new Mullard indirectly-heated subminiature valves, characterised by their extremely robust construction, excellent electrical performance, low heater consumption and small physical dimensions, will shortly be made available for communications and industrial equipment.

Developed originally for Service applications such as guided missiles and fire control systems, they will provide designers with types specially suited to all electronic applications where space is limited and where shock of impact or high *g* vibration is encountered.

The electrical performance of these subminiatures is equal to, and in certain cases even better than, that expected from valves of a much greater size. The EF72 R.F. amplifier, for example, which is suitable for use in the first stage of telecommunications receivers, combines many of the qualities of larger low-noise receiver input valves, with the ability to work at higher frequencies.

To enable experimental and development work to be carried out with these indirectly-heated subminiatures, a limited number of samples can be made available now. Designers who require further information are advised to send their enquiries to the address below as soon as possible.

Type No.	Description	Filament or Heater (V)	(mA)	$V_a = V_{g2}$ (V)	$-V_{g1}$ (V)	I_a (mA)	I_{g2} (mA)	g_m (mA/V)
EA76	Single diode (5 mm. bulb) ...	6.3	150	150 (max.)	—	9.0 (max.)	—	—
EC70	U.H.F. triode oscillator ...	6.3	150	100	2.0	13	—	5.5
EF70	High slope R.F. pentode with short suppressor grid base ...	6.3	200	100	2.0	3.0	2.5	2.5
EF71	Variable-mu R.F. pentode ...	6.3	150	100	1.2	7.2	2.2	4.5
EF72	High slope R.F. pentode ...	6.3	150	100	1.4	7.0	2.2	5.0
EF73	High slope pentode for industrial applications ...	6.3	200	100	2.0	7.5	2.5	5.25
EY70	Half-wave rectifier ...	6.3	450	250 (max.)	—	45 (max.)	—	—
DY70	High voltage rectifier (directly heated) ...	1.25	140	10KV (P.I.V.)	—	2.0 (max.)	—	—

Mullard

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

MVT 130

WIRELESS ENGINEER

Vol. 30 MARCH 1953 No. 3

Folded Dipoles

A DISADVANTAGE of the simple dipole is its low input impedance, which is only a fraction of the characteristic impedance of the ordinary line connecting the dipole to the transmitter or receiver. This necessitates the use of some matching device. This can be avoided by using a folded dipole, the principal advantage of which is its high input impedance. Another advantage is its broader bandwidth, the tuning of the folded dipole not being so sharp as that of the simple dipole. In 1940 J. D. Kraus[1] discussed various types and gave the results of measurements as shown in Fig. 1. The wire used was No. 12 B and S gauge and the overall width d of each dipole was 0·015 λ; the frequency was adjusted in each case to give resonance and the input resistance was determined by means of the standing waves on the transmission line of 570 ohms impedance. The length l of the dipole in terms of the wavelength, and the input resistance are given in each case in Fig. 1. The frequency was about 14 Mc/s. It is important to note that these

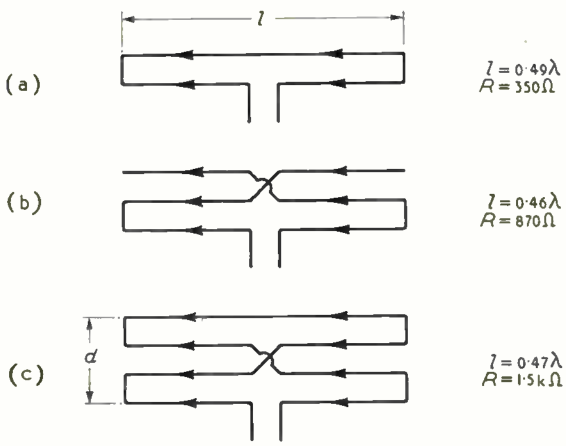


Fig. 1

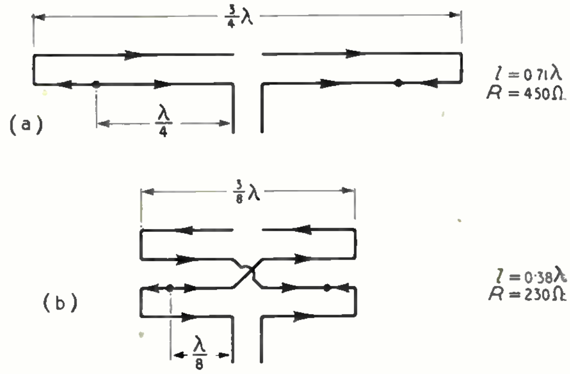


Fig. 2

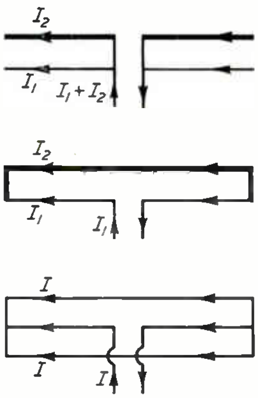
experiments were made with the dipoles horizontal and only about 25 ft above the ground; as $\lambda/2$ was about 35 ft, it is probable that the results were appreciably affected by the proximity of the ground. The input resistances are from 20 to 30% above the calculated values; the high-frequency resistance of the wires constituting the dipoles is only about 1% or less of the measured values and can be ignored.

Two interesting examples of a different type are shown in Fig. 2. The theoretical lengths are indicated on the diagrams; the actual measured lengths for resonance are given beside them, together with the input resistances. These do not lend themselves to such simple calculation as do those shown in Fig. 1.

It is well known that a simple dipole at its

[1] *Electronics*, Jan. 1940, Vol. 13, p. 26; see also P. S. Carter, *R.C.A. Rev.*, Oct. 1939, Vol. 4, p. 168.

resonant frequency has a radiation resistance of 73 ohms, which is not appreciably affected if, instead of a single conductor, two or three conductors are used in close proximity, as shown in Fig. 3(a). With a single-wire dipole taking a current I the power input is 73 I^2, and if the single wire is replaced by two wires of different diameters so that the current in one is I_1, and that in the other I_2, the radiated power will be 73 $(I_1 + I_2)^2$. If now the two elements of the dipole are arranged as in Fig. 3(b), so that the current is only supplied directly to one and induced in the other, for the same currents the radiated power will be unchanged, but the power supplied is equal to $I_1{}^2R$ where R is the input resistance. Hence



(a) (b) (c)

Fig. 3

$$R = 73\left(\frac{I_1 + I_2}{I_1}\right)^2 = 73\ (1 + n)^2$$

where n is the ratio of the current in the secondary element to that in the primary element. If the two elements are similar and $I_1 = I_2$, then $R = 73 \times 4 = 292\ \Omega$. With three elements, as shown in Fig. 1(b) or in Fig. 3(c), the result will be the same as in Fig. 3(b) with $I_2 = 2I_1$, that is $n = 2$ and $R = 73 \times 9 = 657\ \Omega$, but only if the currents in the three elements are equal, and to obtain this result in Fig. 3(c) it is necessary for the central primary element to have twice the diameter of the secondary elements.

We turn now to a consideration of this question of multiple dipoles with elements of the same or different diameters.[2] Fig. 4 represents a simple dipole of length $\lambda/2$ and exaggerated radius a. At the moment of zero current the charge is distributed according to a cosine curve as shown. We wish to calculate the potential at O, the centre point of one end, due to the charge, allowing for retardation, that is to say, the charge at A is not to be taken as it is at this moment, but as it was about half a cycle earlier. Were it not for this retardation the potential at O would be given by the formula

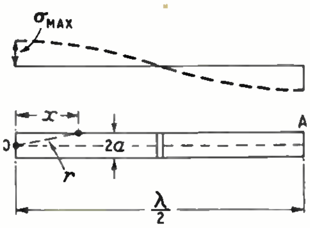


Fig. 4

$$\psi = \int_0^{\lambda/2} \sigma_{max} \frac{1}{r} \cos 2\pi \frac{x}{\lambda}\,.\ dx$$

where σ_{max} is the charge per unit length at the end, x is the axial distance from O, and $r = \sqrt{x^2 + a^2}$. To allow for retardation this has to be multiplied by $\cos 2\pi \frac{r}{\lambda}$ giving

$$\psi = \int_0^{\lambda/2} \sigma_{max} \frac{1}{r} \cos 2\pi \frac{r}{\lambda}\,.\ \cos 2\pi \frac{x}{\lambda}\,.\ dx$$

which may be written

$$\psi = \tfrac{1}{2}\,\sigma_{max} \left[\int_0^{\lambda/2} \frac{1}{r} \cos 2\pi \frac{(r + x)}{\lambda} dx + \int_0^{\lambda/2} \frac{1}{r} \cos 2\pi \frac{(r - x)}{\lambda} dx \right]$$

If λ, r and x are expressed as angles ($\lambda = 2\pi$) then $\frac{1}{r} \cos (r + x)\, dx$ may be written

$$\frac{r + x}{r} \left[\frac{\cos (r + x)}{r + x} d\,(r + x) \right] \frac{dx}{d(r + x)}$$

in which, since

$$\frac{d(r+x)}{dx} = \frac{d[x + (x^2 + a^2)^{\frac{1}{2}}]}{dx}$$

$$= 1 + \frac{x}{\sqrt{x^2 + a^2}} = 1 + \frac{x}{r}$$

$$\frac{r + x}{r} \times \frac{dx}{d(r + x)} = 1.$$

Similarly $\dfrac{r - x}{r} \times \dfrac{dx}{d(r - x)} = -1$

and the two upper limits are approximately $\sqrt{\pi^2 + a^2} + \pi = 2\pi$ and $\sqrt{\pi^2 + a^2} - \pi = a^2/2\pi$.

Hence[3] putting $\text{Ci}\ x = \int_\infty^x \frac{\cos t}{t} dt$ and

$$\text{Cin}\ x = \int_0^x \frac{1 - \cos t}{t} dt$$

$$\psi = \tfrac{1}{2}\,\sigma_{max} \left[\int_a^{2\pi} \frac{\cos (r + x)}{r+x} d(r + x) - \int_a^{a^2/2\pi} \frac{\cos (r - x)}{r - x} d(r - x) \right]$$

$$= \tfrac{1}{2}\,\sigma_{max} \left[-\text{Ci}(a) + \text{Ci}(2\pi) + \text{Ci}(a) - \text{Ci}(a^2/2\pi) \right]$$

$$= \tfrac{1}{2}\,\sigma_{max} \left[\text{Ci}(2\pi) - \text{Ci}(a^2/2\pi) \right]$$

[2] W. B. Roberts, *R.C.A. Rev.*, 1947, Vol. 8, p. 289. Rudolf Guertler, *Proc. Instn Radio Engrs, Australia*, April 1949, Vol. 10, p. 95 and Nov. 1952, Vol. 13, p. 389. It is on these papers by Dr. Guertler that this editorial article is largely based.

[3] See Jahnke and Emde. "Tables of Higher Functions," p. 3.

Putting $\mathrm{Ci}(x) = 0{\cdot}577 + \log_e x - \mathrm{Cin}(x)$, in which $0{\cdot}577$ is Euler's constant, we have

$$\psi = \tfrac{1}{2}\,\sigma_{max}\left[\log_e 2\pi - \mathrm{Cin}(2\pi) - \log_e a^2/2\pi + \mathrm{Cin}(a^2/2\pi)\right]$$

which, since $\mathrm{Cin}\,(a^2/2\pi)$ is negligibly small,

$$= \sigma_{max}\,\tfrac{1}{2}\left[\log_e \frac{4\pi^2}{a^2} - \mathrm{Cin}(2\pi)\right]$$

$$= \sigma_{max}\left[\log_e \frac{2\pi}{a} - \tfrac{1}{2}\,\mathrm{Cin}(2\pi)\right].$$

This is the potential at O due to the charge on the dipole at the moment of maximum charge.

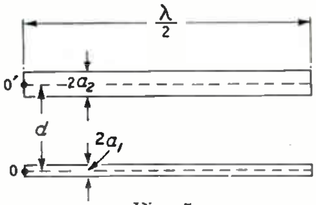


Fig. 5

If now there is a parallel conductor of radius a_2 at a distance d (Fig. 5), the charge on which per unit length is n times as great, the potential at O due to this second conductor will be

$$\psi' = n\sigma_{max}\left[\log_e \frac{2\pi}{d} - \tfrac{1}{2}\,\mathrm{Cin}(2\pi)\right]$$

and therefore the resultant total potential at O will be

$$\phi = \psi + \psi'$$

$$= \sigma_{max}\left[\log_e \frac{\lambda^{n+1}}{ad^n} - \frac{n+1}{2}\,\mathrm{Cin}(2\pi)\right]$$

The angular measure 2π has been replaced by the linear measure λ, and a and d are also now in linear measure.

Similarly, the potential ϕ' at the point O′ on the second element will be the sum of two components,

$$n\sigma_{max}\left[\log_e \frac{\lambda}{a_2} - \tfrac{1}{2}\,\mathrm{Cin}(2\pi)\right]$$ due to its own charge, and

$$\sigma_{max}\left[\log_e \frac{\lambda}{d} - \tfrac{1}{2}\,\mathrm{Cin}(2\pi)\right]$$ due to the charge on the other element. Hence

$$\phi' = \sigma_{max}\left[\log_e \frac{\lambda^{n+1}}{d\;a_2{}^n} - \frac{n+1}{2}\,\mathrm{Cin}(2\pi)\right]$$

When connected together as in a folded dipole the two potentials ϕ and ϕ' are equal, and therefore $a_1 d^n = d a_2{}^n$ or $(d/a_2)^n = d/a_1$, where a_1 and a_2 are the radii of the two elements. Hence the total charge must so distribute itself between the two elements that the ratio n of the one charge to the other is given by the formula

$$n = \log\,(d/a_1)\;/\;\log\,(d/a_2)$$

In calculating the potential at O or O′ due to the charge on the other element, it has been assumed that it is uniformly distributed around the wire, and can therefore be assumed to be on its axis without any great error. If the wires are so close together that the distribution is greatly distorted, this assumption is not justified. We will now apply this result to the symmetrical three-element dipole shown in Fig. 3(c). If the radius of the central element is a_1 and that of the two outer elements a_2, and if the distance between the central axis and either of the outer axes is d, and the charge per unit length on each of the outer elements is m times that on the central element, then, omitting unnecessary factors and $\mathrm{Cin}\,(2\pi)$,

potential at O due to its own charge $= \log \lambda/a_1$
potential at O due to outer element $= 2m \log\lambda/d$
potential at O′ due to its own charge $= m \log\lambda/a_2$
potential at O′ due to central element $= \log\lambda/d$
potential at O′ due to other element $= m \log\lambda/2d$

Equating the two resultant potentials

$$\log \lambda/a_1 + \log\left(\frac{\lambda}{d}\right)^{2m} = \log\left(\frac{\lambda}{a_2}\right)^{m} + \log\left(\frac{\lambda}{2d}\right)^{m} + \log \lambda/d$$

from which

$$m = \frac{\log d/a_1}{\log d/2a_2}$$

From this the current ratio m can be calculated for any values of a_1, a_2 and d. If $a_1 = 2a_2$, $m = 1$ and, since there are two secondary elements, the transformation ratio of the input resistances $(1 + n)^2$ becomes $(1 + 2m)^2$ which is equal to 9.

Another interesting example given by Dr. Guertler is that in which $2a_2 = \sqrt{da_1}$; i.e. the diameter of the outer elements is the geometric mean of the distance d and the radius of the inner element. In this case $m = 2$ and therefore $n = 4$, and the resistance ratio becomes 25.

In his recent article Dr. Guertler gives nomograms from which one can read off the input resistance ratio for various values of a_1, a_2 and d for folded dipoles of two and three elements. He also gives the results of some measurements showing very close agreement with the calculated values.

G. W. O. H.

RC OR DIRECT-COUPLED POWER STAGE

Conditions for Maximum Efficiency

By E. F. Good, M. A.

SUMMARY.—It is shown that in a resistance-capacitance-coupled or direct-coupled output stage consisting of an ideal triode of anode resistance, r_a, an anode feed resistance, R_D, and a given load resistance, R_L, maximum efficiency is obtained when R_D has a definite and finite relationship with r_a and R_L. The relationship is evaluated for a number of particular cases.

Introduction

IT is shown in many textbooks that an ideal triode of anode resistance, r_a, working in class A with a fixed h.t. voltage, and choke- or transformer-coupled to a resistance load, R_L, can deliver into the load maximum a.c. power when $R_L = 2r_a$. It has been pointed out, however, by Prof. Howe,[1] that this is only one solution to the problem of optimum matching for a power output stage and that, if quiescent anode dissipation instead of h.t. voltage is taken as the only limiting factor, the maximum power output rises continuously as the ratio R_L/r_a is increased, and the efficiency of the stage rises from 25% towards a limit of 50%.

In a recent paper by M. G. Scroggie[2] a solution is given to the problem of determining the conditions for getting maximum power from a triode output stage when RC coupling is used, the h.t. supply voltage is fixed, and R_L and R_D (the anode feed resistance) may be varied. This again is only one solution to the problem of optimum matching; for example, the designer may be presented with a fixed value of R_L, perhaps a fixed value of r_a, and the only other fixed condition may be the maximum voltage that is to be developed across the load; the choice of h.t. voltage may be quite open. In these circumstances, rather than work out a design for minimum h.t. voltage, the designer may wish to achieve maximum power efficiency, so that the d.c. input power and the heat dissipated by the circuit may be as low as possible—especially if, as is often the case nowadays, a large amount of circuitry must be compressed into a small amount of space.

For a transformer-coupled stage there is, as already indicated, no finite solution to the problem of maximum efficiency; the efficiency continuously increases as the ratio R_L/r_a is increased, although in practice there is a limit set by the maximum h.t. voltage that may safely be used. Now consider an RC-coupled stage, Fig. 1. We know from Scroggie's paper that for a given value of maximum voltage to be developed across R_L, and a given value of r_a, the h.t. voltage required will be at a minimum when R_D has a certain finite value. (For example if $r_a = R_L$, then R_D should be equal to $\sqrt{2}R_L$.) If R_D is reduced below this value, the h.t. voltage required is increased and, since the shunting effect of R_D across R_L is increased, the maximum alternating current through the valve, and consequently the direct current, is also increased. Consequently the d.c. power input to the stage is increased and the efficiency reduced. If, on the other hand, the value of R_D is increased above the value given by Scroggie's analysis, the h.t. voltage must again be increased but, since the shunting effect of R_D across R_L is reduced, the alternating current which has to flow through the valve, and consequently the direct current, is reduced.

It seems, therefore, that although a reduction of R_D below the value calculated from Scroggie's analysis results in an increase of the d.c. power input to the stage, an increase in the value of R_D, by reducing the h.t. current required, may result in a net decrease in the power consumed.

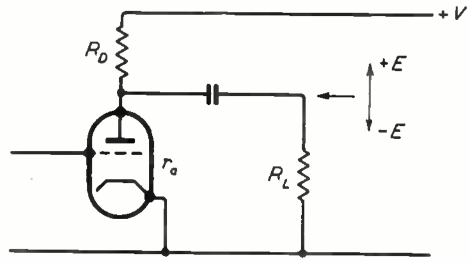


Fig. 1. *RC-coupled stage.*

Of course, if R_D is increased indefinitely the h.t. voltage required also increases without limit, while the h.t. current does not fall below a minimum value equal to the peak value of the alternating current required in the load. It is clear, therefore, that if r_a and R_L are finite, maximum efficiency (and consequently minimum waste of power as heat) will be obtained for a finite value of R_D. At the moment we do not know if this is the same as the value for minimum h.t. voltage; it may be a higher value.

MS accepted by the Editor, April 1952

Before proceeding to find an answer to the problem, however, it is useful to look again at Fig. 1 and notice that if the lower end of the load, R_L, is taken to a fixed voltage equal to the steady voltage on the anode of the valve, the coupling capacitor can be dispensed with, and the working conditions of the circuit are not altered in any way. This means that a solution for the RC-coupled circuit of Fig. 1 is the same as a solution for the direct-coupled circuit of Fig. 2; and it is in terms of Fig. 2 that the problem will now be stated and some solutions obtained.

Problem

Given that the triode valve in Fig. 2 has idealized characteristics as shown in Fig. 3, and that the value of its anode resistance is r_a; find working conditions to give a maximum voltage swing of $\pm E$ across the given resistance load, R_L, with minimum quiescent power consumption from the h.t. supplies. [Note: If the signal is purely a.c. (i.e., contains no d.c. component) and the stage is distortionless (i.e., works in strict class A) the direct component of the anode current does not change when the signal is applied, and the power drawn from the h.t. supplies remains constant at the quiescent value.]

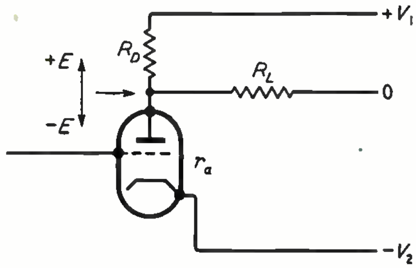


Fig. 2. *Direct-coupled stage equivalent to Fig. 1.*

Solution

If we assume that maximum efficiency can be obtained only if the valve is driven from cutoff to $V_g = 0$, the dynamic load line, PQR, must lie across the valve curves as shown in Fig. 3. (Note: PQR represents the net load, R_L and R_D in parallel. SQT represents R_D alone, and becomes the load line if R_L is removed.)

From Fig. 3, or by applying Ohm's law directly, we can now write down three equations:—

At the point R the valve is cut off, and the output voltage (i.e., the voltage across R_L) is $+E$.

$$\therefore E = \frac{R_L}{R_D + R_L} V_1$$

$$\text{i.e., } V_1 = \frac{R_D + R_L}{R_L} E \quad \dots \quad \dots \quad (1)$$

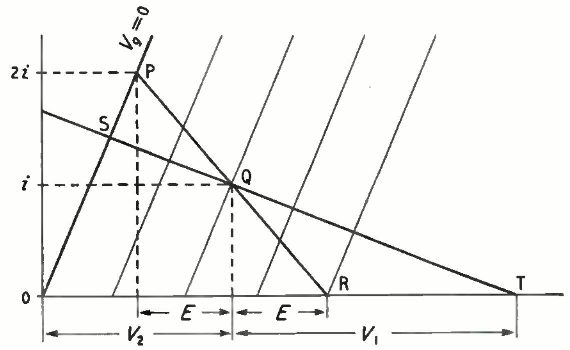


Fig. 3. *Characteristics of ideal triode.*

At the quiescent point, Q, the output voltage $= 0$, and the anode current is given by

$$i = \frac{V_1}{R_D}$$

or, substituting from (1)

$$i = \frac{R_D - R_L}{R_D R_L} E \quad \dots \quad \dots \quad \dots \quad (2)$$

At the full-on point, P, the output voltage is $-E$, and the anode current is $2i$.

$$\therefore \frac{V_2 - E}{r_a} = 2i = \frac{2(R_D + R_L)}{R_D R_L} E$$

$$\text{i.e., } V_2 = \left\{1 + \frac{2r_a\,(R_D + R_L)}{R_D R_L}\right\} E$$

$$= \frac{R_D R_L + 2r_a(R_D + R_L)}{R_D R_L} E \quad \dots \quad (3)$$

Hence, the power taken from the h.t. supplies is given by

$$W = (V_1 + V_2)i = \left\{\frac{R_D(R_D + R_L)}{R_D R_L} + \frac{R_D R_L + 2r_a(R_D + R_L)}{R_D R_L}\right\} \frac{R_D + R_L}{R_D R_L} E^2$$

$$= \frac{\{R_D^2 + 2(R_L + r_a)R_D + 2R_L r_a\}(R_D + R_L)}{R_D^2 R_L^2} E^2$$

$$= \frac{R_D^3 + (3R_L + 2r_a)R_D^2 + 2R_L(R_L + 2r_a)R_D + 2R_L^2 r_a}{R_D^2 R_L^2} E^2 \quad \dots \quad \dots \quad (4)$$

To obtain the required minimum value of W, we differentiate with respect to R_D, and equate to zero. Thus

$$\frac{dW}{dR_D} = \left\{1 - \frac{2R_L(R_L + 2r_a)}{R_D^2} - \frac{4R_L^2 r_a}{R_D^3}\right\} \frac{E^2}{R_L^2} = 0$$

$$\text{i.e., } R_D^3 - 2R_L(R_L + 2r_a)R_D - 4R_L^2 r_a = 0 \quad (5)$$

This is a cubic equation which is not apparently factorizable; so it seems best to start with a solution for the limiting case, $r_a = 0$, and then obtain approximate solutions for a series of numerical values of r_a and R_L.

For $r_a = 0$, Equ. (5) becomes

$$R_D^2 - 2R_L^2 = 0$$

i.e., $R_D = \sqrt{2}R_L$ .. .. .. (6)

to 1·625 (approx.); and so on until we are able to construct a table of R_D/R_L against r_a/R_L (Table 1) and a curve (Fig. 4).

Now Equ. (4) gives the power consumption of the stage, and the maximum output power (sine-wave) developed in R_L is $E^2/2R_L$. Hence the efficiency of the stage at maximum output is given by

$$\eta = \frac{E^2}{2R} \cdot \frac{1}{W} = \frac{R_D^2 R_L}{2\{R_D^3 + (3R_L + 2r_a)R_D^2 + 2R_L(R_L + 2r_a)R_D - 2R_L^2 r_a\}} \quad (7)$$

Thus for $r_a/R_L = 0$, the optimum value of R_D/R_L is 1·4 (approx.). Proceeding then to $r_a/R_L = 0{\cdot}1$, we find the optimum value of R_D/R_L increases

Substituting corresponding values of r_a/R_L and R_D/R_L found already, we get the values given under η in Table 1 and by the second curve in Fig. 4.

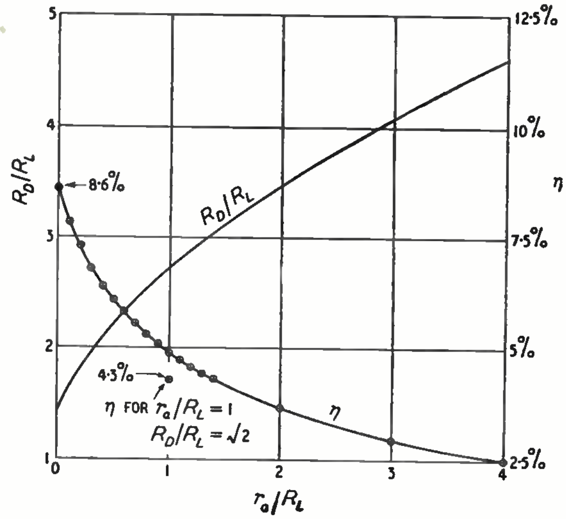


Fig. 4. *Optimum R_D/R_L versus r_a/R_L and corresponding efficiency η.*

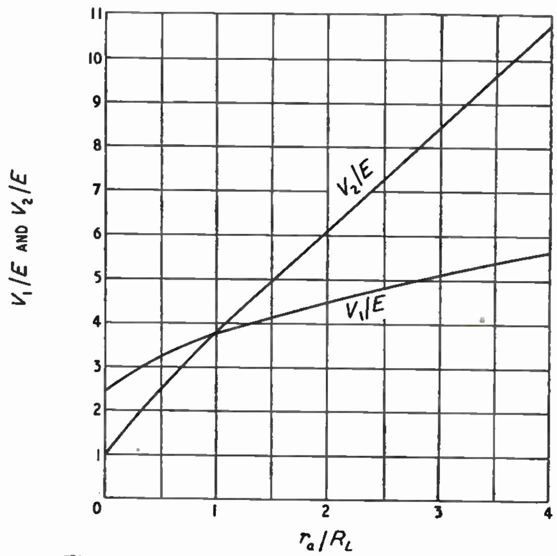


Fig. 5. *Optimum V_1 and V_2 versus r_a/R_L.*

TABLE 1

Optimum Values of Anode-Feed Resistance, R_D, and Corresponding Efficiency, η.

r_a/R_L	R_D/R_L	η %
0	1·41	8·6
0·1	1·63	7·85
0·2	1·80	7·3
0·3	1·95	6·8
0·4	2·09	6·4
0·5	2·21	6·1
0·6	2·33	5·8
0·7	2·44	5·55
0·8	2·54	5·3
0·9	2·64	5·1
1·0	2·73	4·9
1·4	3·05	4·3
2·0	3·50	3·65
3·0	4·11	2·95
4·0	4·63	2·5

TABLE 2

Required H.T. Voltages, V_1 and V_2.

r_a/R_L	V_1/E	V_2/E
0	2·41	1·00
0·1	2·63	1·32
0·2	2·80	1·62
0·3	2·95	1·91
0·4	3·09	2·18
0·5	3·21	2·45
0·6	3·33	2·71
0·7	3·44	2·97
0·8	3·54	3·23
0·9	3·64	3·48
1·0	3·73	3·73
1·4	4·05	4·72
2·0	4·50	6·14
3·0	5·11	8·46
4·0	5·63	10·74

We see then that the greatest efficiency that can be obtained is 8·6%, when $r_a = 0$; and that as r_a/R_L is increased the efficiency falls to still lower values. In addition it is interesting to calculate the efficiency for Scroggie's arrangement for maximum output power when r_a and $V_1 + V_2$ are fixed; i.e., $r_a/R_L = 1$, $R_D/R_L = \sqrt{2}$. The result is $\frac{1}{4\{2\sqrt{2}+3\}}$ which equals 4·3% (approx.), or just one half the efficiency that can be achieved when $r_a/R_L = 0$.

To complete the design of the stage when E has been given and r_a, R_L, and R_D have been fixed, we return to equations (1) and (2) and substitute values from Table 1 to obtain corresponding values of V_1 and V_2. Results are given in Table 2 and in the curves of Fig. 5.

Such then are the results for an ideal triode. For any real triode, of course, allowance must be made for curvature of the valve characteristics and, in general, the net effect is that higher h.t. voltages and a higher mean anode current will be needed for a given 'undistorted' output. In a practical design it will also be necessary to make allowance for possible variations between different specimens of the type of valve chosen. Nevertheless the value of R_D calculated for the nearest ideal triode will be an entirely suitable value to use with a real valve, and it will remain true that for a given output with minimum h.t. power input a triode with the lowest possible r_a should be used.

One other point is worth mentioning, and that is the relevance of the above analysis when a pentode or beam-tetrode type valve is used. When a pentode-type valve is used efficiently the load line lies across the valve curves so that the point, P, lies just below the knee in the characteristic

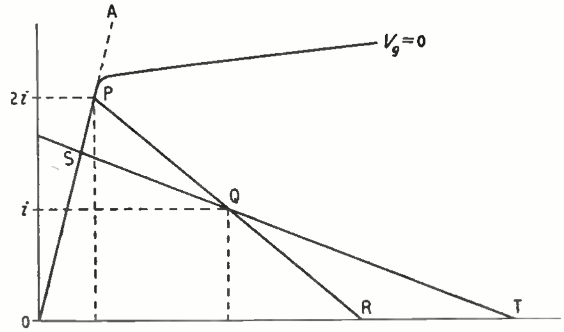


Fig. 6. *Basic pentode characteristic.*

(Fig. 6). Consequently, the effective value of r_a for present purposes is not the usual high value but the low value representing the line OA. Indeed in most cases this value will be so low compared with R_L that it may be taken as almost zero, and so we arrive at the useful rule that with a pentode R_D should be 1·5 × R_L (approx.).

REFERENCES

[1] "Optimum Conditions in Class A Amplifiers", Editorial, *Wireless Engineer*, February 1943, Vol. 20, p. 53.
[2] "RC-Coupled Power Stage", by M. G. Scroggie, B.Sc., M.I.E.E., *Wireless Engineer*, March 1950, Vol. 27, p. 81.

VIBRATION TEST FOR VALVES

Use of Repeated Impacts

By C. C. Eaglesfield, M.A., A.M.I.E.E.

(Standard Telephones and Cables, Ltd.)

SUMMARY.—A machine is described which gives a repeated-impact test to valves. The advantages are that a wide band of resonances is excited and that the test lasts only a short time. The impact train is under sufficient control to be closely specified.

A short description is also given of the kind of routine adopted for tests of this kind, including a check on the microphony, which is excited by this machine in a rather basic way.

Introduction

FOR a very long time there has been a demand for valves which work under conditions of heavy vibration without giving trouble. In 1938 Standard Telephones & Cables introduced a series of valves (4033-AF, 4307-AF, etc.) specially tested for use in the standard aircraft radio equipment, every valve being subject to a run of half an hour under working conditions on a bumping table. Today the demand is growing rapidly and it is therefore necessary to consider carefully the mechanical tests which can be used to ensure the quality of the production.

Since the internal structure of any ordinary valve has many mechanical resonances, it is clearly not sufficient to vibrate at one or more arbitrary frequencies. Such a test might strike a resonance in one type and miss it in another, and the trend of design might well be to construct valves whose resonances avoided the specified frequencies. One way around this difficulty is to

MS accepted by the Editor, July 1952

use a 'gliding tone' covering a band of frequencies, which would ensure that no resonance was missed.

A simpler way of achieving the same result is to use impact testing, which also is effectively aperiodic. The advantages of impact testing are indeed generally agreed, but at first sight it seemed to the writer that the difficulties of specifying and controlling the mechanical conditions might prove very great. However, it was found in practice that the problem was easier than had been expected and the following description of a very simple apparatus may therefore be of general interest.

Instead of having a few large impacts, there are advantages in using a train of impacts, each so small that the valve is not overstressed, but with a large cumulative effect. All the impacts could lie in the same direction or they could, for instance, alternate in opposite directions. Again they might lie in different planes so that the valve was shocked from every direction. The method that has been used, chosen for its practical simplicity, is to make the impacts all lie in the same direction and to arrange the valve so that its three main axes are equally inclined to the line of impacts.

A description will be given in later sections of the simple vibration gear which has been constructed, the methods used in checking the nature of the impacts it gives, and the routine that has been followed when testing valves in this way. For the moment we can consider the specification of the impact train.

Specification of the Impact Train

For the purpose in hand we may consider a valve to be a rigid envelope containing internal parts whose supports have a number of mechanical resonances. The effect on the valve of any applied vibration is determined by the amplitude to which the internal parts are excited and the applied vibration is specified by the movement of the valve envelope. In what follows, velocities and accelerations are to be understood as referring to the envelope.

For the sake of completeness it is desirable to include some elementary discussion of impulses and pulses but the reader to whom the subject is familiar may omit much of this section.

A single impact will throw the resonances into oscillation having an amplitude proportional to the size of the impact. This size is the area under the acceleration-time curve, which is usually in the shape of a pulse. The area is, of course, the velocity-change.

If the impact is a mathematically ideal impulse, the acceleration is infinite and the duration infinitesimal, but the area is finite. An impact of finite duration can be as effective as an impulse if the resonance considered has a periodic time large compared to the duration of the pulse. Thus the duration of the pulse decides whether a resonance is excited at all. The duration is more difficult to define than the size.

This question of relating the duration of the pulse to the frequency of the resonances excited can be dealt with in the following way. A pulse of finite duration may be regarded as the output of a low-pass filter excited by an impulse. The relation between the duration of the pulse and the pass-band of the filter depends on the definitions used and is in general rather complicated. The writer has dealt with this question previously.[1]

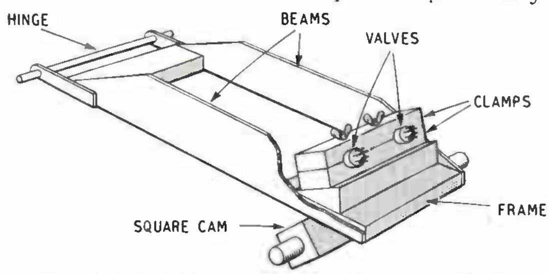


Fig. 1. *A general view of the impact machine.*

However, there is a simple empirical relation which is adequate for the present purpose. This is that

$$f_c = 1/2t_c.$$

In this f_c is the cut-off frequency at which the transmission of the filter is reduced to 70% of the low-frequency transmission, and t_c is defined as though the acceleration were constant and equal to the actual peak acceleration. This time is easy to estimate from a given velocity-time curve, if the transition is substantially smooth, without undue oscillation.

It is a matter of judgment to decide, in practice, whether the transition can be regarded as sufficiently smooth.

It is now possible to replace an input of finite duration by an impulsive impact acting through a low-pass filter. Then for resonances well below f_c in frequency, the size of the impact is determined only by the velocity change; a resonance at f_c is excited to 70% of what would be calculated from the velocity change; and resonances well above f_c are barely excited at all.

We have now reached the point where we can see how to define each impact in terms of a velocity change and a duration time. In addition it is necessary to specify a repetition frequency. The numerical values that have been used for these three parameters of the impact train will be dealt with later, but it is useful to mention here some considerations in choosing them.

[1] "Transition Time and Pass Band," by C. C. Eaglesfield, *Proc. Inst. Radio Engrs*, Feb. 1947, Vol. 35, No. 2.

The repetition frequency should be fairly low, to allow time for the resonances to decay away between impacts. On the other hand the combination of the repetition frequency and the size of impact should be chosen so that a fairly severe dose of treatment is administered in a reasonable time.

The duration of the impact should not be made very short. It is true that a short impact will excite a wide range of resonances, but it is liable to vary from one machine to another or even in the same machine according to the details of the set-up.

The compromises arrived at will be discussed in a later section.

The Impact Machine

The important parts of the machine are shown in Fig. 1, in which will be seen a frame attached by two deep beams to a hinge. The frame bears on a square cam which is made to rotate. Now if the angular speed of the cam were very slow, the frame would follow the cam exactly and would always be in contact with it. At higher cam speeds "bouncing" occurs and it is possible to choose a size of cam, in relation to its speed, such that each of its edges strikes the frame in turn. This is the adjustment that has been used in this machine and the free riding of the frame, which rises and falls between blows under gravitational influence alone, has been checked both by stroboscopic inspection and by verifying electrically that contact between the frame and the cam lasts only a very short time at each blow.

On the frame there is an arrangement for supporting a pair of valves, clamps being provided to hold them down. Not shown in the figure, but used in practice, are valve sockets to make contact to the pins of the valves.

The materials used in the construction were aluminium for the beams, and wood for the frame and clamps. The frame has a face plate of hard tool steel and the cam is also of tool steel. A sheet of aluminium was put between the face plate and the frame. The clamps are lined with baize, to reduce the possibility of crushing the glass envelope of the valve.

The clatter and vibration from this machine was objectionable and therefore it was enclosed in a sound-proof box. To prevent it from shaking itself to pieces, all wood-screws were cemented in and lock-washers were used throughout.

Measurement of the Shock Train

After some experiment it was decided to use a repetition frequency of thirteen per second.

Now this frequency effectively determines the size of the shock which itself depends on the drop. We can visualize the frame bouncing on the cam and therefore having a time $1/2f$ (f being the repetition frequency) in which to drop. It is very easy to calculate the velocity change at each impact, if we imagine the whole mass concentrated in the frame and neglect the constraining effect of the beams. The velocity change is $g/f = 32/13 = 2{\cdot}46$ ft/sec.

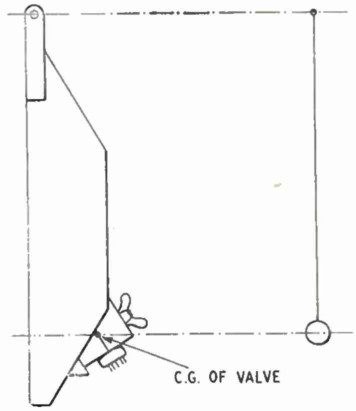


Fig. 2. *This illustrates a simple way of checking that the valve is put at the centre of percussion.*

The effect of the inertia of the beams can be eliminated by placing the valves at the centre of percussion of the system. When this is done, the motion of the valve is again that of a small free mass and the size of shock again is g/f. The easiest way to check the adjustment is to hang the frame from its hinges, as shown in Fig. 2, and verify that the periodic time is the same as that of the bob pendulum shown at the side.

The duration of the shocks has been measured by mounting a coil (moulded in plastic) on the frame and using a magnet to produce an e.m.f. in the coil (see Fig. 3). The e.m.f. is proportional to the velocity and by connecting the coil to an oscillograph and photographing the trace, a record of the velocity-time curve was made. This is shown in Fig. 4(a) which shows that the velocity changes smoothly from one value to another at the shock without undue oscillation. Fig. 4(b) shows the transition on an expanded time scale, and from it the transition time was estimated at $0{\cdot}4$ millisecond.

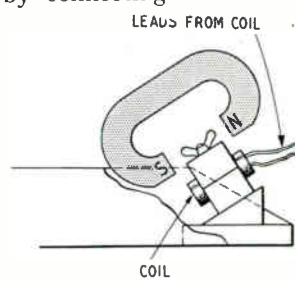


Fig. 3. *A diagrammatic illustration of the way the coil and magnet are arranged.*

This measurement was made using coils moulded to the approximate size and weight of valves of each common type. The coils were clamped in the place where the valve was put. Also the measurement was made both with and without baize in the clamps. The result was always the same, from which it seems fair to deduce that the impact duration of 0·4 millisecond is characteristic of the machine.

The transition time is evidently determined by the resonant frequency of the beams, and it should therefore be possible to repeat this time on similar machines. It is worth remarking that aluminium was used for the beams partly because of its lightness and partly for its high mechanical damping. The smooth transition that has been achieved is due to this damping, both in the beams and in the aluminium plate sandwiched between the hard face plate and the frame.

Experience with this machine suggests that it should not be difficult to make machines to give a specified impact train. A specification which this machine would fit would be

Repetition frequency: 13 per sec.
Size of each impact: 2·5 ft/sec.
Duration of each impact: 0·4 millisecond.

In writing a specification, it would be as well to provide limits, at any rate for the duration of the impact, which could hardly be reproduced exactly.

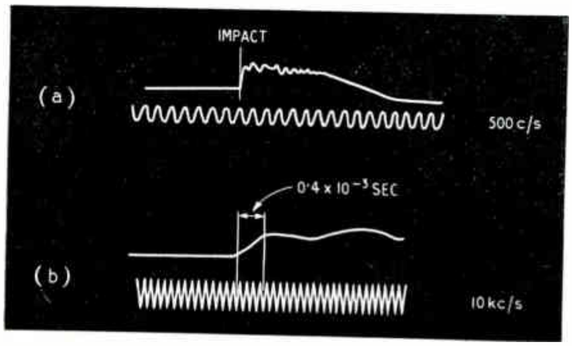


Fig. 4. (*a*) *A reproduction of the photographic trace, showing the velocity-time curve;* (*b*) *an enlargement of the part of the trace near the impact.*

While the figures given above seem very suitable for the purpose in hand, other specifications could doubtless be met by suitable simple machines.

As a matter of interest, it may be remarked that the above figures correspond to a cut-off frequency of 1,250 c/s and a peak acceleration of 190 g; that is, 190 × 32 ft/sec^2.

The Routine of Valve Testing

There is one further thing which has not so far been mentioned that has been found valuable in this machine. Since the mechanical excitation is of a rather general kind, it serves as a good basis for a test of microphony. Accordingly, the microphonic output of the valve is checked during the test. This output, at the start of the test, is quite a general measure of the valve's microphony, and any increase in the output is an indication that wear has taken place, probably in the insulators.

It is usually sufficient to strap the valve as a diode and pass a moderate direct current through it and a high resistance in series. A suitable millivoltmeter is connected between anode and cathode to indicate the microphony. This meter should preferably respond to the mean arithmetic voltage; this is a good practical compromise, although there are theoretical arguments for using an r.m.s. meter. But a peak voltmeter would be objectionable.

Since the whole process of repeated impact testing must be regarded as semi-destructive, it has been found best to proceed on a sampling basis. The routine is very simple. From each batch of tested valves, a random sample is taken and vibrated; the fate of the batch depends on the number of rejections in the sample.

To pass, a valve must still pass all the clauses of the normal test specification after vibration, and its microphony at the end of the period must be below a limit set for the type. (The normal limits are applied for the ordinary tests.)

The sample size is laid down according to the batch size and the permissible number of rejects is also specified. If this is not exceeded the batch is passed; otherwise the whole batch is rejected.

The samples themselves are destroyed. A suitable period of vibration for this test has been found to be half an hour.

Acknowledgment

The writer is grateful to a number of his colleagues whose help made possible the development of this machine, in particular to Mr. C. H. Foulkes, who first suggested the desirability of a test of this type, and to Mr. J. Hookings, who made the machine and did much of the work. Finally, acknowledgment is due to the Directors of Standard Telephones & Cables for permission to publish this article.

PHYSICAL SOCIETY'S EXHIBITION

This year's exhibition is from Monday, 13th April to Friday, 17th April, at the Imperial College of Science & Technology, Imperial Institute Road, Exhibition Road, London, S.W.7. Apart from one heavy exhibit in the Mathematics Department of the Huxley Building in Exhibition Road (for which entrance tickets will not be needed), the exhibition is to be all in the main building of Imperial College. Applications for tickets should be made to the Exhibition Secretary, 1 Lowther Gardens, Prince Consort Road, London, S.W.7. The tickets are of four kinds; whole day for 13th or 17th April; whole day for 14th, 15th or 16th April; morning, 10 a.m.-1 p.m., 14th-17th; and evening, 5.30 p.m.-8 p.m., 13th-16th.

The exhibition is open to Fellows and the press only from 10.30 a.m. to 2 p.m. on 13th April and for all ticket holders from 2 p.m. to 8 p.m. On other days the hours are 10 a.m. to 8 p.m. except on the 17th when the exhibition closes at 5 p.m.

FOUR-TERMINAL NETWORKS

Fundamental Theory

By O. P. D. Cutteridge, M.Sc.(Eng.)

(Queen Mary College, University of London)

SUMMARY.—The object of this paper is to present the steady-state theory of passive 4-terminal networks (4-poles) in a manner that, taken as a whole, is believed to have the elements of novelty and further, to be more rigorous than the classical approaches to the subject. The general 4-pole is regarded as being composed of n meshes and the external equations of the 4-pole are obtained by solving the n-mesh equations using determinants. Iterative and image impedances are obtained as impedance properties of a single 4-pole and a suitable propagation function is associated with each pair of impedances. The relationship of these functions to the open- and short-circuit impedance functions of the 4-pole is derived. Conditions in 4-poles in two types of cascade are shown to be immediately deducible from the external equations of a single 4-pole. The paper concludes with tables giving expressions for the image functions in terms of the iterative functions (and vice versa) and giving the various external equations of a 4-pole written in terms of the iterative and image functions.

Introduction

CLASSICAL 4-terminal network theory was developed more or less simultaneously in a number of countries at the beginning of the twentieth century for dealing with telephone lines and their associated networks. The theory has long ceased to be the interest of telephone engineers solely, and today is used throughout the whole field of communication engineering and electronics. Applications of the theory to subjects quite outside the communications field have been given; for example, in pure physics, to wave propagation in crystals[1] and to thermodynamics.[2] Classical 4-terminal network theory is really the algebra of linear (passive) one-dimensional transmission systems and a large number of diverse subjects could, perhaps, be usefully linked together under this heading.

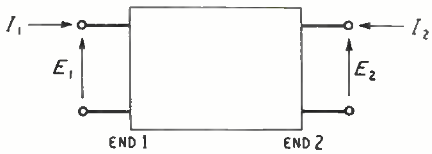


Fig. 1. *General form of 4-pole network defining the terminal voltages and currents.*

Passive 4-terminal network theory is normally developed in one of two ways; either around the conception of the equivalent T and π networks[3] or from the difference-equations treatment of a cascade of 4-poles.[4] This latter is an exact parallel with the well-known differential-equations treatment of the uniform transmission line, except that an additional parameter appears due to the asymmetry of the general 4-pole. In the present treatment the external equations of a single 4-pole are obtained from the more fundamental mesh equations. The usual iterative and image functions are discussed with reference to a single 4-pole and their existence as one-valued functions investigated. Conditions in 4-poles in two types of cascade are then shown to be immediately deducible from the external equations of a single 4-pole.

MS accepted by the Editor, April 1952

1. Derivation of Fundamental Steady-State Equations

The general passive 4-pole with linear, time-invariant, elements can be regarded as being composed of n meshes, mesh 1 being completed by voltage E_1 at end 1 and mesh 2 being completed by voltage E_2 at end 2. Since the 4-pole is being taken to be passive, no alternators are present in the interior meshes of the networks and so the following equations are obtained by applying Kirchhoff's voltage law to each of the n meshes in turn:—

$$\left.\begin{aligned}
E_1 &= Z_{11}I_1 + Z_{12}I_2 + Z_{13}I_3 + \ldots + Z_{1n}I_n \\
E_2 &= Z_{21}I_1 + Z_{22}I_2 + Z_{23}I_3 + \ldots + Z_{2n}I_n \\
0 &= Z_{31}I_1 + Z_{32}I_2 + Z_{33}I_3 + \ldots + Z_{3n}I_n \\
0 &= Z_{41}I_1 + Z_{42}I_2 + Z_{43}I_3 + \ldots + Z_{4n}I_n \\
&\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \\
&\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \\
0 &= Z_{n1}I_1 + Z_{n2}I_2 + Z_{n3}I_3 + \ldots + Z_{nn}I_n
\end{aligned}\right\} \quad (1)$$

$I_1, I_2, I_3 \ldots I_n$ being Maxwell's mesh currents, I_1 and I_2 defined as in Fig. 1 and the sense of circulation of the others being immaterial. Z_{ii} ($i = 1, 2, 3, \ldots n$) is the self impedance of the i^{th} mesh and Z_{ij} ($i \neq j$: $i, j = 1, 2, 3, \ldots n$) is the mutual impedance between meshes i and j; i.e., the total impedance common to meshes i and j. If the 4-pole did not contain unilateral components, such as radio valves to couple one mesh to another, then Z_{ij} would equal Z_{ji}. However, it is generally taken that restricting the 4-pole to be passive excludes radio valves as coupling components by virtue of the equivalent alternator circuit. It should be noted that equations (1) hold whether the 4-pole is being driven from end 1 into a load at end 2 or vice versa, or being driven from both ends 1 and 2 simultaneously.

Solving equations (1) by the method of determinants (Cramer's rule) we have

$$I_j = \frac{1}{\Delta}\sum_{i=1}^{n} E_i D_{ij} \qquad j = 1, 2 \quad \ldots \quad (2)$$

where Δ is the determinant of the Z terms, D_{ij} are cofactors and $D_{ij} = D_{ji}$ if $Z_{ij} = Z_{ji}$.

Since D_{ij}/Δ has the dimensions of admittance we write $D_{ij}/\Delta = y_{ij}$ and (2) becomes:—

$$I_j = \sum_{i=1}^{n} E_i\, y_{ij} \qquad j = 1, 2 \quad \ldots \quad (3)$$

Since impedance properties of the 4-pole will be used later, it is better to have the fundamental external relationships in the form:—

$$E_j = \sum_{i=1}^{n} I_i\, z_{ij} \qquad j = 1, 2 \quad \ldots \quad (4)$$

where

$$\left.\begin{aligned} z_{12} = z_{21} &= -\frac{y_{12}}{y_{11}\,y_{22} - y_{12}^2} \\ z_{11} &= \frac{y_{22}}{y_{11}\,y_{22} - y_{12}^2} \\ z_{22} &= \frac{y_{11}}{y_{11}\,y_{22} - y_{12}^2} \end{aligned}\right\} \quad \ldots \quad \ldots \quad (5)$$

By reference to equations (3) and (4) the y and z terms can be interpreted as terms of the network's open- and short-circuit, driving-point and transfer impedances and admittances.

The three functions (of frequency) y_{11}, y_{12}, y_{22} (or z_{11}, z_{12}, z_{22}) completely determine the external properties of a passive 4-pole. That is to say, given the three functions y_{11}, y_{12}, y_{22} (or z_{11}, z_{12}, z_{22}) and any two of the quantities E_1, E_2, I_1, I_2 then the remaining two of the latter quantities can be determined. For 4-poles containing radio valves as coupling elements, four functions would have to be used completely to determine the external properties. On the other hand a symmetrical passive 4-pole, that is a passive 4-pole whose external behaviour is unchanged on interchanging ends 1 and 2, only requires two functions to determine its external behaviour.

2. Open- and Short-Circuit Impedances

Let us write Z_{sc1} (Z_{oc1}) for the impedance seen looking into end 1 with end 2 short-circuited (open-circuited) and Z_{sc2} (Z_{oc2}) for the impedance looking into end 2 of the 4-pole with end 1 short-circuited (open-circuited).

If we open-circuit end 2, $I_2 = 0$ and using equations (4) we have:—

$$Z_{oc1} = (E_1/I_1)_{I_2=0} = z_{11} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

We may note here that by virtue of the symmetry of our notation and reference directions we may interchange suffixes 1 and 2 in all our equations. This will often be done in future working. From (6), for example, we have immediately:

$$Z_{oc2} = z_{22} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

If we now short-circuit end 2, $E_2 = 0$,

$$\therefore\ Z_{sc1} = (E_1/I_1)_{E_2=0} = z_{11} - \frac{z_{12}^2}{z_{22}} = \frac{|z|}{z_{22}} \quad \ldots \quad \ldots \quad (8)$$

$$\text{where}\quad |z| = \begin{vmatrix} z_{11} & z_{12} \\ z_{12} & z_{22} \end{vmatrix} = z_{11}z_{22} - z_{12}^2 \quad \ldots \quad (9)$$

Interchanging suffixes 1 and 2 in (8) gives:—

$$Z_{sc2} = \frac{|z|}{z_{11}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

So, for any 4-pole:—

$$Z_{oc1}Z_{sc2} = Z_{oc2}Z_{sc1} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

$$\text{Since } z_{12}^2 = Z_{oc2}\,(Z_{oc1} - Z_{sc1}) \quad \ldots \quad \ldots \quad (12)$$

a knowledge of any three of the open- and short-circuit impedances of the 4-pole may be said to 'almost determine' its external behaviour. Two externally different 4-poles could actually be constructed having identical open- and short-circuit impedances, but with the transfer impedance of one being the negative of the transfer impedance of the other. Thus the two 4-poles could be made externally identical merely by interchanging the output (or input) leads of one of them.

Since, for a symmetrical 4-pole, $z_{11} = z_{22}$,

$$Z_{oc1} = Z_{oc2} = Z_{oc} \text{ say} \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

$$Z_{sc1} = Z_{sc2} = Z_{sc} \text{ say} \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

3. Iterative Impedances and Iterative Propagation Function

Suppose an idealized constant-voltage generator having zero internal impedance be connected to end 1 of a 4-pole and end 2 be closed by a passive impedance Z_2 (i.e., the real part of Z_2 is $\geqslant 0$). Let the impedance that the generator looks into at end 1 be Z_1. Now, in general, the input impedance, Z_1, of the 4-pole and the terminating impedance, Z_2, will have different values. The question arises whether it is possible to satisfy the equation:—

$$Z_1 = Z_2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

for any values of Z_2 and, if so, how many? The answer to this question will now be investigated.

The general external equations for a 4-pole, (4), written in full, were:—

$$\begin{aligned} E_1 &= z_{11}I_1 + z_{12}I_2 \\ E_2 &= z_{12}I_1 + z_{22}I_2 \end{aligned} \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

and the following boundary conditions have to be satisfied in this case:—

$$E_1/I_1 = Z_1; \quad -E_2/I_2 = Z_2 \quad \ldots \quad \ldots \quad (17)$$

Substituting (17) in (16) yields finally:—

$$Z_1 = z_{11} - \frac{z_{12}^2}{Z_2 + z_{22}} \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

Thus, Z_1 will equal Z_2 for values of Z_2 satisfying the equation

$$Z_2^2 + Z_2 (z_{22} - z_{11}) + (z_{12}^2 - z_{11} z_{22}) = 0 \quad (19)$$

Since we have made the proviso that the real part of Z_2 must be greater than or equal to zero, the number of solutions to (15) depends on the distribution of the roots of (19) on the complex Z_2 plane.

Let us now make the following definition:—

The iterative impedance looking into end 1 of a 4-pole is Z_{01} if, when an impedance Z_{01} closes end 2, the impedance seen looking into end 1 is Z_{01}. Similarly the iterative impedance looking into end 2 of a 4-pole is Z_{02} if, when an impedance Z_{02} closes end 1, the impedance seen looking into end 2 is Z_{02}.

That these impedance functions always exist as passive impedances and can both be made one-valued will now be shown.

It is evident that, if one identifies a certain root of (19) with Z_{01}, then Z_{02} is minus the other root of (19). In the Appendix it is demonstrated that with physically-realizable passive 4-poles the real parts of the roots of (19) must be of opposite sign with the real part of one root zero as the limiting case. Thus it is always possible to choose the impedance functions Z_{01} and Z_{02} from the roots of (19) in such a manner that both Z_{01} and Z_{02} exist as passive impedances and are both one-valued.

From (19) we have:—

$$Z_{01} = \frac{z_{11} - z_{22} + \sqrt{(z_{11} + z_{22})^2 - 4z_{12}^2}}{2} \quad (20)$$

and interchanging suffixes 1 and 2 gives:—

$$Z_{02} = \frac{z_{22} - z_{11} + \sqrt{(z_{11} + z_{22})^2 - 4z_{12}^2}}{2} \quad (21)$$

The following results are easily derived using the results of Section 2,

$$Z_{01} \cdot Z_{02} = Z_{oc1} \cdot Z_{sc2} = Z_{oc2} \cdot Z_{sc1} = |z| \quad \ldots \quad (22)$$

$$Z_{01} - Z_{02} = Z_{oc1} - Z_{oc2} \quad \ldots \quad (23)$$

$$Z_{01} = \frac{Z_{oc1} - Z_{oc2} + \sqrt{(Z_{oc1} - Z_{oc2})^2 + 4Z_{sc1} \cdot Z_{oc2}}}{2} \quad \ldots \quad \ldots \quad \ldots \quad (24)$$

$$Z_{02} = \frac{Z_{oc2} - Z_{oc1} + \sqrt{(Z_{oc1} - Z_{oc2})^2 + 4Z_{sc1} \cdot Z_{oc2}}}{2} \quad \ldots \quad \ldots \quad \ldots \quad (25)$$

For a symmetrical 4-pole,

$$Z_{01} = Z_{02} = Z_0, \text{ say } = \sqrt{z_{11}^2 - z_{12}^2} = \sqrt{Z_{oc} Z_{sc}} \quad \ldots \quad \ldots \quad \ldots \quad (26)$$

Thus the two normally-different iterative impedances of a 4-pole become equal when the 4-pole is symmetrical.

Iterative Propagation Function

One other function, in addition to the two iterative impedances, is necessary in order to determine the external properties of a 4-terminal network and it can be arbitrary except that it must be independent of Z_{01} and Z_{02}.

However, a function expressing the voltage gain (or attenuation) of a network will probably be the most generally useful and, moreover, following the usual communication-engineering arguments, the logarithm of the voltage (or current) ratio would be preferable to the plain ratio of voltages. Thus we have the definition:—

The iterative propagation function, γ, of a 4-terminal network is the natural logarithm of the ratio input current to output current when the network is working into the appropriate iterative impedance. We shall assume for the moment that there are two possible values of γ for a given 4-pole; γ_1, say, when driving from end 1 into a load Z_{01} at end 2 and γ_2 when driving from end 2 into a load Z_{02} at end 1, as shown in Fig. 2(a) and (b). The equations for these two cases are then:—

$$\gamma_1 = \log\left(\frac{I_1}{-I_2}\right) \quad \ldots \quad \ldots \quad \ldots \quad (27)$$

and

$$\gamma_2 = \log\left(\frac{I_2'}{-I_1'}\right) \quad \ldots \quad \ldots \quad \ldots \quad (28)$$

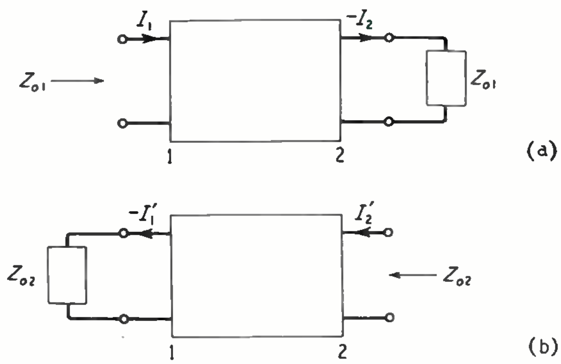


Fig. 2. *Iterative propagation function of network driven at end* 1(a), *and driven at end* 2(b).

We shall now show that for any passive 4-terminal network:—

$$\gamma_1 = \gamma_2 = \gamma, \text{ say.}$$

Taking the general external equations (16) and inserting the appropriate boundary conditions for driving from end 1 into a load impedance equal to Z_{01} at end 2 we have:—

$$I_1 Z_{01} = z_{11} I_1 + z_{12} I_2$$
$$-I_2 Z_{01} = z_{12} I_1 + z_{22} I_2 \quad \ldots \quad \ldots \quad (29)$$

Therefore,

$$z_{11} + z_{12} (I_2/I_1) = -z_{12} (I_1/I_2) - z_{22}$$

and writing $(I_1/I_2) = \exp.(\gamma_1)$, we have,

$$\exp.(\gamma_1) + \exp.(-\gamma_1) = \frac{z_{11} + z_{22}}{z_{12}} \quad \ldots \quad (30)$$

$$\text{Therefore, } \cosh \gamma_1 = \frac{z_{11} + z_{22}}{2z_{12}} \quad \ldots \quad \ldots \quad (31)$$

Since interchanging suffixes 1 and 2 leaves the right-hand side of (31) unchanged, we see that:—

$$\gamma_1 = \gamma_2 = \gamma, \text{ say,}$$

and γ, the iterative propagation function is given by:—

$$\cosh \gamma = \frac{z_{11} + z_{22}}{2z_{12}} \quad .. \quad .. \quad .. \quad (32)$$

Thus the three independent functions Z_{01}, Z_{02} and γ can be used to determine the external properties of a 4-terminal network.

4. Image Impedances and the Image Propagation Function

Let end 2 of a 4-terminal network be closed by an impedance Z_{I2} and let the impedance then seen looking into end 1 be Z_{I1}. Now let end 1 be closed by an impedance equal to Z_{I1} and let the impedance now seen looking into end 2 be Z'_{I2}. We shall now investigate whether it is possible to satisfy the equation:—

$$Z'_{I2} = Z_{I2}$$

for any values of Z_{I2} and if so for how many values?

The equations corresponding to the above boundary conditions are as follows:—

$$E_1/I_1 = Z_{I1}; \quad -E_2/I_2 = Z_{I2} \quad .. \quad (33)$$

$$-E_1/I_1 = Z_{I1}; \quad E_2/I_2 = Z'_{I2} \quad .. \quad (34)$$

Substituting equations (33) and (34) in the general external equations (16), we see, after some manipulation, that Z'_{I2} will equal Z_{I2} for values of Z_{I2} satisfying the equation,

$$Z_{I2}^2 + \frac{z_{22}z_{12}^2}{z_{11}} - z_{22}^2 = 0 \quad .. \quad .. \quad (35)$$

The roots of (35) lie either one in each vertical half-plane or both along the imaginary axis. If, in the latter case, we conventionally agree to take the value in the upper half-plane and bearing in mind we restricted Z_{I2} to be a passive impedance, it is seen that Z_{I2} always exists as a one-valued passive impedance. It now follows from (16) that Z_{I1} also always exists as a one-valued impedance and is further a passive impedance, since both Z_{I2} and the 4-pole are passive. We thus define two image impedances as follows:—

The two image impedances of a 4-terminal network, Z_{I1} and Z_{I2} (appropriate to end 1 and end 2 of the network respectively), are of such a value that

(*a*) an impedance Z_{I1} is seen looking into end 1 when end 2 is closed by Z_{I2} and

(*b*) an impedance Z_{I2} is seen looking into end 2 when end 1 is closed by Z_{I1}.

Solving (35) we have:—

$$Z_{I2} = \sqrt{z_{22}^2 - \frac{z_{22}z_{12}^2}{z_{11}}} \quad .. \quad .. \quad .. \quad (36)$$

and interchanging suffixes 1 and 2 gives:—

$$Z_{I1} = \sqrt{z_{11}^2 - \frac{z_{11}z_{12}^2}{z_{22}}} \quad .. \quad .. \quad .. \quad (37)$$

These results are obtained in a neater form bv writing $|z|$ for $z_{11}z_{22} - z_{12}^2$ as before,

$$Z_{I1} = \sqrt{\frac{z_{11}\,|z|}{z_{22}}} \quad .. \quad .. \quad .. \quad .. \quad (38)$$

$$Z_{I2} = \sqrt{\frac{z_{22}\,|z|}{z_{11}}} \quad .. \quad .. \quad .. \quad .. \quad (39)$$

Dividing and multiplying (38) by (39) gives:—

$$Z_{I1}/Z_{I2} = z_{11}/z_{22} \quad .. \quad .. \quad .. \quad (40)$$

$$\text{and } Z_{I1} \, . \, Z_{I2} = |z| \quad .. \quad .. \quad .. \quad .. \quad (41)$$

Substituting (6) and (8) in (38) yields the simple expressions for the image impedances in terms of the open- and short-circuit impedances of the 4-pole:—

$$Z_{I1} = \sqrt{Z_{sc1} \, . \, Z_{oc1}} \quad .. \quad .. \quad .. \quad (42)$$

and interchanging suffixes 1 and 2 gives:—

$$Z_{I2} = \sqrt{Z_{sc2} \, . \, Z_{oc2}} \quad .. \quad .. \quad .. \quad (43)$$

Since the open- and short-circuit impedances of a 4-pole can frequently be obtained by inspection, these equations are often used in practice to obtain the values of the image impedances of the network. From (22) and (41) we have the interesting results:—

$$Z_{I1} \, . \, Z_{I2} = Z_{01} \, . \, Z_{02} = Z_{oc1} \, . \, Z_{sc2} = Z_{oc2} \, . \, Z_{sc1} \quad (44)$$

For a symmetrical 4-pole:—

$$Z_{I1} = Z_{I2} = Z_{01} = Z_{02} = \sqrt{Z_{sc} \, . \, Z_{oc}} = Z_o \quad (45)$$

Thus the four normally different impedance functions of a 4-pole become equal when the 4-pole is symmetrical.

Image Propagation Function

It will be found more convenient to have the propagation function associated with the two image impedances expressed as a ratio of input to output volt-amperes rather than a ratio of input to output currents (or voltages). When working into iterative impedances, voltage, current and volt-ampere ratios all give the same result but when working into image impedances the results normally differ. Thus we have the definition:—

The image propagation function, θ, of a 4-terminal network is one-half of the natural logarithm of the ratio input volt-amperes to output volt-amperes when the network is working into the appropriate image impedance.

As with the iterative propagation function, we shall first assume that there are two possible values of θ, θ_1, say, when driving from end 1 into a load Z_{I2} at end 2 and θ_2 when driving from end 2 into a load Z_{I1} at end 1, and later we shall show that the values of the two are equal. They give the equations

$$\theta_1 = \tfrac{1}{2} \log \left(E_1 I_1 / - E_2 I_2\right) \quad .. \quad .. \quad (46)$$

$$\text{and} \quad \theta_2 = \tfrac{1}{2} \log \left(E_2' I_2' / - E_1' I_1'\right) \quad .. \quad .. \quad (47)$$

We shall now show that for any passive 4-terminal network:—

$$\theta_1 = \theta_2 = \theta, \text{ say}$$

Taking the general external equations (16) and inserting the appropriate boundary conditions for driving from end 1 into a load Z_{I2} at end 2 we have:—

$$\left.\begin{aligned} 1 &= z_{11}/Z_{I1} + z_{12}(I_2/E_1) \\ 1 &= z_{12}(I_1/E_2) - z_{22}/Z_{I2} \end{aligned}\right\} \quad .. \quad .. \quad (48)$$

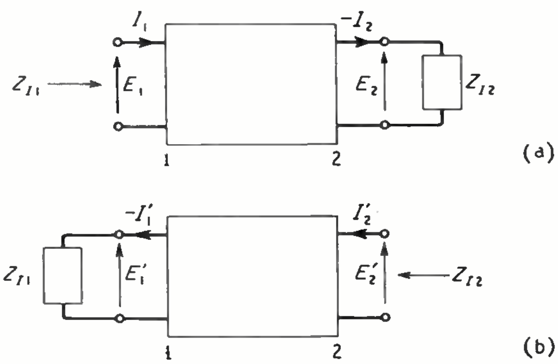


Fig. 3. *Image propagation function of network driven at end* 1(a), *and driven at end* 2(b).

From (48) we obtain:—

$$\exp.(2\theta_1) = E_1 I_1/(-E_2 I_2)$$
$$= \frac{z_{22}Z_{I1} + Z_{I1}Z_{I2}}{z_{11}Z_{I2} - Z_{I1}Z_{I2}} \quad .. \quad .. \quad (49)$$

Using (40) we see that interchanging suffixes 1 and 2 leaves the right-hand side of (49) unchanged and hence:—

$$\theta_1 = \theta_2 = \theta$$

and θ, the image propagation function, is given by:—

$$\tanh \theta = \frac{\exp.(2\theta) - 1}{\exp.(2\theta) + 1}$$
$$= \frac{Z_{I2}}{z_{22}} \quad .. \quad .. \quad .. \quad (50)$$

Using the value previously obtained for Z_{I2} in (39) we have finally:—

$$\tanh \theta = \sqrt{\frac{|z|}{z_{11} \cdot z_{22}}} \quad .. \quad .. \quad (51)$$

Thus the independent functions Z_{I1}, Z_{I2}, and θ can be used to determine the external properties of a 4-terminal network.

Substituting (6)-(10) in (51) yields the simple expressions for the image propagation function in terms of the open- and short-circuit impedances of the 4-pole:—

$$\tanh \theta = \sqrt{\frac{Z_{sc1}}{Z_{oc1}}} = \sqrt{\frac{Z_{sc2}}{Z_{oc2}}} \quad .. \quad .. \quad (52)$$

Like (42) and (43), this equation is also often used in practice.

For a symmetrical 4-pole, using (13), (14) we have:—

$$\tanh \theta = \frac{\sqrt{z_{11}^2 - z_{12}^2}}{z_{11}} = \sqrt{\frac{Z_{sc}}{Z_{oc}}} \quad .. \quad (53)$$

Comparing with (32) we see that the normally different image and iterative propagation functions are identical when the network is symmetrical.

Thus when symmetrical 4-poles are considered, the image and iterative functions reduce to the same pair of functions determining the 4-pole's external behaviour.

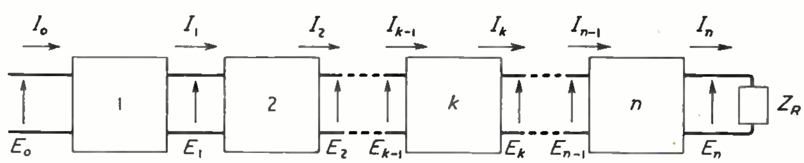
5. 4-Poles Connected in Cascade

By virtue of our definitions of the iterative and image functions the resultant functions for two types of cascade follow immediately; viz, (a) a cascade such that iterative impedances corresponding to all left-hand ends of all 4-poles are the same, iterative impedances corresponding to all right-hand ends of all 4-poles are the same, while values of the iterative propagation functions are immaterial and (b) a cascade such that image impedances at any two ends to be joined are equal (i.e., matched image impedances at all junctions). The external equations for these cascades are hence obtained straight away from the external equations of a single 4-pole. We shall first consider the case of n identical 4-poles cascaded on an iterative basis and then n identical 4-poles cascaded on an image, or back-to-back, basis.

5.1. *Iterative Cascade of n Identical 4-Poles*

We shall here find it convenient to adopt the notation of Fig. 4 for the voltages and currents.

Fig. 4. *This diagram defines the notation for voltages and currents in a cascade of 4-poles.*



Equations (54) and (55) follow immediately from the appropriate pair of equations in Table 2 by replacing γ by $n\gamma$ and $k\gamma$ for (54) and (55) respectively and remembering the different notation now being employed:—

$$E_0 = \frac{Z_{01}\,\text{exp.}\,(n\gamma) + Z_{02}\,\text{exp.}\,(-n\gamma)}{Z_{01} + Z_{02}} E_n + \frac{Z_{01}Z_{02}[\text{exp.}\,(n\gamma) - \text{exp.}\,(-n\gamma)]}{Z_{01} + Z_{02}} I_n \quad \ldots \quad (54)$$

$$I_0 = \frac{\text{exp.}\,(n\gamma) - \text{exp.}\,(-n\gamma)}{Z_{01} + Z_{02}} E_n + \frac{Z_{02}\,\text{exp.}\,(n\gamma) + Z_{01}\,\text{exp.}\,(-n\gamma)}{Z_{01} + Z_{02}} I_n$$

$$E_k = \frac{Z_{02}\,\text{exp.}\,(k\gamma) + Z_{01}\,\text{exp.}\,(-k\gamma)}{Z_{01} + Z_{02}} E_0 - \frac{Z_{01}Z_{02}[\text{exp.}\,(k\gamma) - \text{exp.}\,(-k\gamma)]}{Z_{01} + Z_{02}} I_0$$

$$-I_k = \frac{\text{exp.}\,(k\gamma) - \text{exp.}\,(-k\gamma)}{Z_{01} + Z_{02}} E_0 - \frac{Z_{01}\,\text{exp.}\,(k\gamma) + Z_{02}\,\text{exp.}\,(-k\gamma)}{Z_{01} + Z_{02}} I_0 \quad \ldots \quad \ldots \quad (55)$$

and for the boundary condition at the far end we have:—

$$E_n/I_n = Z_R \quad \ldots \quad \ldots \quad \ldots \quad (56)$$

From these five equations the voltages and currents at any point along the cascade may be determined.

Solving for E_k and I_k in terms of E_0, for example, we have finally:—

$$E_k = E_0 \frac{Z_{01}\,\text{exp.}\,(\overline{n-k}\gamma) + Z_{02}r_R\,\text{exp.}\,(-\overline{n-k}\gamma)}{Z_{01}\,\text{exp.}\,(n\gamma) + Z_{02}r_R\,\text{exp.}\,(-n\gamma)} \quad \ldots \quad \ldots \quad \ldots \quad (57)$$

$$I_k = E_0 \frac{\text{exp.}\,(\overline{n-k}\gamma) - r_R\,\text{exp.}\,(-\overline{n-k}\gamma)}{Z_{01}\,\text{exp.}\,(n\gamma) + Z_{02}r_R\,\text{exp.}\,(-n\gamma)} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (58)$$

$$\text{where } r_R = \frac{Z_R - Z_{01}}{Z_R + Z_{02}} \quad \ldots \quad \ldots \quad \ldots \quad (59)$$

These equations are similar to the equations giving the voltage and current along a uniform transmission line, but are rather more complicated due to the asymmetry of the general 4-pole.

One can regard incident and reflected waves as being present in a cascade of 4-poles just as in uniform transmission lines. It can easily be seen from the above equations that the condition for absence of the reflected wave is that the chain should be terminated by the appropriate iterative impedance. A further interpretation of the iterative impedance, easily derived from the above equations, is that it is the impedance seen looking into the appropriate end of an infinite iterative chain of 4-poles, however terminated, but provided that the real part of γ is greater than zero. The insertion ratio is given in Table 4.

5.2. *Image or Back-to-Back Cascade of n Identical 4-Poles*

Adopting the same notation for the voltages and currents as with the iterative cascade, we have the following equations immediately from the appropriate pair of equations in Table 3 by replacing θ by $n\theta$ and $k\theta$ for (60) and (61), respectively, and noting the different notation now being employed:—

$$E_0 = \sqrt{\frac{Z_{I1}}{Z_{I2}}}(\cosh n\theta)E_n + \sqrt{Z_{I1}\,.\,Z_{I2}}(\sinh n\theta)I_n$$

$$I_0 = \frac{\sinh n\theta}{\sqrt{Z_{I1}\,.\,Z_{I2}}}E_n + \sqrt{\frac{Z_{I2}}{Z_{I1}}}(\cosh n\theta)I_n \quad \ldots \quad \ldots \quad \ldots \quad (60)$$

$$E_k = \sqrt{\frac{Z_{I2}}{Z_{I1}}}(\cosh k\theta)E_0 - \sqrt{Z_{I1}Z_{I2}}\,(\sinh k\theta)I_0$$

$$-I_k = \frac{\sinh k\theta}{\sqrt{Z_{I1}Z_{I2}}}E_0 - \sqrt{\frac{Z_{I1}}{Z_{I2}}}(\cosh k\theta)I_0 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (61)$$

and for the boundary condition at the far end we have again:—

$$E_n/I_n = Z_R \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (62)$$

The above equations (60) and (61) apply for both n and k odd and for end 1 of the first 4-pole to the left. Equations (61) and (62) actually give complete information from which all quantities in an image cascade can be calculated when it is borne in mind that:—

(a) If n is even, one must change suffix 2 to suffix 1 on the image impedances in (60).

(b) If k is even, one must change suffix 2 to suffix 1 on the image impedances in (61).

(c) If end 2 of the first 4-pole is to the left, one must interchange suffixes 1 and 2 on the image impedances in both (60) and (61) after making the changes in (a) and (b) above.

Solving for E_k and I_k in terms of E_0, for example, we have finally,

$$E_k = E_0\sqrt{\frac{Z_{I2}}{Z_{I1}}}\left[\frac{\text{exp.}\,(\overline{n-k}\theta) + r_{R2}\,\text{exp.}\,(-\overline{n-k}\theta)}{\text{exp.}\,(n\theta) + r_{R2}\,(-n\theta)}\right] \quad (63)$$

$$I_k = \frac{E_0}{\sqrt{Z_{I1}Z_{I2}}}\left[\frac{\text{exp.}\,(\overline{n-k}\theta) - r_{R2}\,\text{exp.}\,(-\overline{n-k}\theta)}{\text{exp.}\,(n\theta) + r_{R2}\,\text{exp.}\,(-n\theta)}\right] \quad (64)$$

$$\text{where } r_{R1} = \frac{Z_R - Z_{I1}}{Z_R + Z_{I1}};\ r_{R2} = \frac{Z_R - Z_{I2}}{Z_R + Z_{I2}} \quad \ldots \quad (65)$$

(63) and (64) are correct for both n and k odd:—

For n even: change suffix 2 on the reflection coefficients to suffix 1.

For k even: change suffix 2 on the image impedances to suffix 1.

A wave interpretation of the above equations is again possible, but a further complication arises here, since in addition to E_k and I_k being functions of a discrete variable, k, the form of the functions changes according to whether k is odd or even. However, as might be expected from the definition of the image propagation function, the volt-ampere function, $E_k I_k$, is the same function regardless of whether k is odd or even. From (65) we see that the reflected wave disappears when the load impedance is equal to the appropriate image impedance. From the above equations it can also be seen that the appropriate image impedance is seen on looking into an infinite image chain of 4-poles, however terminated, but provided that the real part of θ is greater than zero.

The insertion ratio in terms of the image functions is given in Table 4.

TABLE 1

Expressions for the Image Functions in Terms of the Iterative Functions and vice versa.

$$Z_{I1} = \sqrt{\frac{Z_{01}Z_{02}\,[Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)]}{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)}}$$

$$Z_{I2} = \sqrt{\frac{Z_{01}Z_{02}\,[Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)]}{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)}}$$

$$\sinh\theta = \frac{\sqrt{Z_{01}Z_{02}}}{Z_{01}+Z_{02}}\,[\exp.(\gamma) - \exp.(-\gamma)]$$

$$= \frac{\sqrt{Z_{01}Z_{02}}}{\frac{1}{2}(Z_{01}+Z_{02})}\sinh\gamma$$

i.e., $\sinh\theta =$

$$\frac{\text{geometric mean of iterative impedances}}{\text{arithmetic mean of iterative impedances}}\sinh\gamma$$

$$Z_{01} = \frac{(Z_{I1}-Z_{I2})\cosh\theta + \sqrt{(Z_{I1}+Z_{I2})^2\cosh^2\theta - 4Z_{I1}Z_{I2}}}{2\sinh\theta}$$

$$Z_{02} = \frac{(Z_{I2}-Z_{I1})\cosh\theta + \sqrt{(Z_{I1}+Z_{I2})^2\cosh^2\theta - 4Z_{I1}Z_{I2}}}{2\sinh\theta}$$

$$\cosh\gamma = \frac{\frac{1}{2}(Z_{I1}+Z_{I2})}{\sqrt{Z_{I1}Z_{I2}}}\cosh\theta$$

i.e., $\cosh\gamma = \dfrac{\text{arithmetic mean of image impedances}}{\text{geometric mean of image impedances}}\cosh\theta$

TABLE 2

Evidently the external passive 4-pole equations (3), (4) can be written in six different ways altogether* viz:—

$$\left.\begin{aligned} I_1 &= y_{11}E_1 + y_{12}E_2 \\ I_2 &= y_{12}E_1 + y_{22}E_2 \end{aligned}\right\} \qquad \left.\begin{aligned} E_1 &= z_{11}I_1 + z_{12}I_2 \\ E_2 &= z_{12}I_1 + z_{22}I_2 \end{aligned}\right\}$$

$$\left.\begin{aligned} I_1 &= g_{11}E_1 + g_{12}I_2 \\ E_2 &= g_{21}E_1 + g_{22}I_2 \end{aligned}\right\} \qquad \left.\begin{aligned} E_1 &= h_{11}I_1 + h_{12}E_2 \\ I_2 &= h_{21}I_1 + h_{22}E_2 \end{aligned}\right\}$$

$$\left.\begin{aligned} E_1 &= AE_2 - BI_2 \\ I_1 &= CE_2 - DI_2 \end{aligned}\right\} \qquad \left.\begin{aligned} E_2 &= DE_1 - BI_1 \\ I_2 &= CE_1 - AI_1. \end{aligned}\right\}$$

The following table expresses the various 4-pole functions in terms of the iterative functions:—

$$y_{11} = \frac{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)}{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]} = \frac{(Z_{02}-Z_{01}) + (Z_{01}+Z_{02})\coth\gamma}{2\,Z_{01}Z_{02}}$$

$$y_{12} = -\frac{Z_{01}+Z_{02}}{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]} = -\frac{Z_{01}+Z_{02}}{2\,Z_{01}Z_{02}\sinh\gamma}$$

$$y_{22} = \frac{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)}{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]} = \frac{(Z_{01}-Z_{02}) + (Z_{01}+Z_{02})\coth\gamma}{2\,Z_{01}Z_{02}}$$

$$z_{11} = \frac{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)}{\exp.(\gamma) - \exp.(-\gamma)} = \frac{(Z_{01}-Z_{02}) + (Z_{01}+Z_{02})\coth\gamma}{2}$$

$$z_{12} = \frac{Z_{01}+Z_{02}}{\exp.(\gamma) - \exp.(-\gamma)} = \frac{Z_{01}+Z_{02}}{2\sinh\gamma}$$

$$z_{22} = \frac{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)}{\exp.(\gamma) - \exp.(-\gamma)} = \frac{(Z_{02}-Z_{01}) + (Z_{01}+Z_{02})\coth\gamma}{2}$$

$$g_{11} = \frac{\exp.(\gamma) - \exp.(-\gamma)}{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)} = \frac{2}{(Z_{01}-Z_{02}) + (Z_{01}+Z_{02})\coth\gamma}$$

$$g_{12} = -g_{21} = -\frac{Z_{01}+Z_{02}}{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)} = -\frac{Z_{01}+Z_{02}}{(Z_{01}-Z_{02})\sinh\gamma + (Z_{01}+Z_{02})\cosh\gamma}$$

$$g_{22} = \frac{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]}{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)} = \frac{2\,Z_{01}Z_{02}}{(Z_{01}-Z_{02}) + (Z_{01}+Z_{02})\coth\gamma}$$

$$h_{11} = \frac{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]}{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)} = \frac{2\,Z_{01}Z_{02}}{(Z_{02}-Z_{01}) + (Z_{01}+Z_{02})\coth\gamma}$$

$$h_{12} = -h_{21} = \frac{Z_{01}+Z_{02}}{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)} = \frac{Z_{01}+Z_{02}}{(Z_{02}-Z_{01})\sinh\gamma + (Z_{01}+Z_{02})\cosh\gamma}$$

$$h_{22} = \frac{\exp.(\gamma) - \exp.(-\gamma)}{Z_{02}\exp.(\gamma) + Z_{01}\exp.(-\gamma)} = \frac{2}{(Z_{02}-Z_{01}) + (Z_{01}+Z_{02})\coth\gamma}$$

$$A = \frac{Z_{01}\exp.(\gamma) + Z_{02}\exp.(-\gamma)}{Z_{01}+Z_{02}} = \frac{(Z_{01}+Z_{02})\cosh\gamma + (Z_{01}-Z_{02})\sinh\gamma}{Z_{01}+Z_{02}}$$

$$B = \frac{Z_{01}Z_{02}\,[\exp.(\gamma) - \exp.(-\gamma)]}{Z_{01}+Z_{02}} = \frac{2Z_{01}Z_{02}\sinh\gamma}{Z_{01}+Z_{02}}$$

* The notation used here follows E. A. Guillemin, "Communication Networks," Vol. II, pp. 135-138. This Table and Table 3 can be regarded as an extension of the table given on pp. 137-138 of Guillemin's book. It should be noted that while the y and z terms are all respectively admittances and impedances, they are not the reciprocals of each other. The g and h terms are not all of the same dimensions; thus, g_{11} is an admittance but g_{12} is a numeric.

$$C = \frac{\exp.(\gamma) - \exp.(-\gamma)}{Z_{01} + Z_{02}}$$
$$= \frac{2 \sinh \gamma}{Z_{01} + Z_{02}}$$
$$D = \frac{Z_{02} \exp.(\gamma) + Z_{01} \exp.(-\gamma)}{Z_{01} + Z_{02}}$$
$$= \frac{(Z_{01} + Z_{02}) \cosh \gamma + (Z_{02} - Z_{01}) \sinh \gamma}{Z_{01} + Z_{02}}$$

The determinants of the various coefficients are as follows:—

$$|z| = z_{11}z_{22} - z_{12}^2 = Z_{01}Z_{02}$$
$$|y| = y_{11}y_{22} - y_{12}^2 = 1/Z_{01}Z_{02}$$
$$|g| = g_{11}g_{22} - g_{21}g_{12} = \frac{Z_{02} \exp.(\gamma) + Z_{01} \exp.(-\gamma)}{Z_{01} \exp.(\gamma) + Z_{02} \exp.(-\gamma)}$$
$$= \frac{(Z_{02} - Z_{01}) + (Z_{01} + Z_{02}) \coth \gamma}{(Z_{01} - Z_{02}) + (Z_{01} + Z_{02}) \coth \gamma}$$
$$|h| = h_{11}h_{22} - h_{21}h_{12} = \frac{Z_{01} \exp.(\gamma) + Z_{02} \exp.(-\gamma)}{Z_{02} \exp.(\gamma) + Z_{01} \exp.(-\gamma)}$$
$$= \frac{(Z_{01} - Z_{02}) + (Z_{01} + Z_{02}) \coth \gamma}{(Z_{02} - Z_{01}) + (Z_{01} + Z_{02}) \coth \gamma}$$
$$\begin{vmatrix} A & B \\ C & D \end{vmatrix} = AD - BC = 1$$

TABLE 3

In this table the various 4-pole functions are expressed in terms of the image functions. The notation for the various functions is the same as in Table 2.

$$y_{11} = \frac{\coth \theta}{Z_{I1}}; \quad y_{12} = -\frac{1}{\sqrt{Z_{I1}Z_{I2}}} \cdot \frac{1}{\sinh \theta}; \quad y_{22} = \frac{\coth \theta}{Z_{I2}}$$
$$z_{11} = Z_{I1} \coth \theta; \quad z_{12} = \frac{\sqrt{Z_{I1}Z_{I2}}}{\sinh \theta}; \quad z_{22} = Z_{I2} \coth \theta$$
$$g_{11} = \frac{\tanh \theta}{Z_{I1}}; \quad g_{12} = -g_{21} = -\sqrt{\frac{Z_{I2}}{Z_{I1}}} \cdot \frac{1}{\cosh \theta};$$
$$g_{22} = Z_{I2} \tanh \theta$$
$$h_{11} = Z_{I1} \tanh \theta; \quad h_{12} = -h_{21} = \sqrt{\frac{Z_{I1}}{Z_{I2}}} \cdot \frac{1}{\cosh \theta};$$
$$h_{22} = \frac{\tanh \theta}{Z_{I2}}$$
$$A = \sqrt{\frac{Z_{I1}}{Z_{I2}}} \cosh \theta; \quad B = \sqrt{Z_{I1}Z_{I2}} \sinh \theta;$$
$$C = \frac{\sinh \theta}{\sqrt{Z_{I1}Z_{I2}}}; \quad D = \sqrt{\frac{Z_{I2}}{Z_{I1}}} \cosh \theta$$

The determinants of the various coefficients are as follows:—

$$|z| = Z_{I1}Z_{I2}; \quad |y| = 1/Z_{I1}Z_{I2}; \quad |g| = Z_{I2}/Z_{I1};$$
$$|h| = Z_{I1}/Z_{I2}; \quad \begin{vmatrix} A & B \\ C & D \end{vmatrix} = 1$$

TABLE 4

Expressions for the Insertion Ratio of a Single 4-pole in Terms of the Iterative and Image Functions.

From (57) of Section 5 we have, after a little algebra, and for the conditions of Fig. 5.:—

$$E_2 = E_g \cdot \frac{Z_R}{Z_R + Z_S} \cdot \exp.(-\gamma) \cdot (1 - r_R r_S) \cdot \{1 - r_R r_S \exp.(-2\gamma)\}^{-1}$$

where $r_R = \dfrac{Z_R - Z_{01}}{Z_R + Z_{02}}$ and $r_S = \dfrac{Z_S - Z_{02}}{Z_S + Z_{01}}$

If E_2' is the voltage across load Z_R without 4-pole (i.e., with Z_R directly connected to alternator),

$$E_2' = Eg \cdot \frac{Z_R}{Z_R + Z_S}$$

i.e., Insertion ratio $= \dfrac{E_2'}{E_2} = \exp.(\gamma) \cdot (1 - r_R r_S)^{-1} \cdot \{1 - r_R r_S \exp.(-2\gamma)\}$

The corresponding expressions in terms of the image functions are:—

$$E_2 = Eg \cdot \frac{Z_R \sqrt{\dfrac{Z_{I1}}{Z_{I2}}}}{Z_S + Z_R \dfrac{Z_{I1}}{Z_{I2}}} \cdot \exp.(-\theta) \cdot (1 - r_{S1} r_{R2}) \cdot \{1 - r_{S1} r_{R2} \exp.(-2\theta)\}^{-1}$$

where $r_{S1} = \dfrac{Z_S - Z_{I1}}{Z_S + Z_{I1}}$ and $r_{R2} = \dfrac{Z_R - Z_{I2}}{Z_R + Z_{I2}}$

$$\text{Insertion ratio} = \frac{\left(1 + \dfrac{Z_R}{Z_S}\right)\sqrt{\dfrac{Z_{I1}}{Z_{I2}}}}{1 + \dfrac{Z_R}{Z_S} \cdot \dfrac{Z_{I1}}{Z_{I2}}} \cdot \exp(\theta) \cdot (1 - r_{S1} r_{R2})^{-1} \{1 - r_{S1} r_{R2} \exp.(-2\theta)\}$$

Taking the real and imaginary parts of the natural logarithm of the expressions for the insertion ratio yields the 'insertion loss' in nepers and the 'insertion phase' in radians.

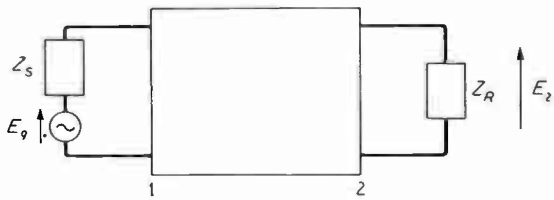


Fig. 5. *This diagram defines the notation used in obtaining the insertion ratio.*

APPENDIX

Let the general quadratic equation be

$$Z^2 + PZ + Q = 0$$

where P and Q are, in general, complex.

It is clear that if P equals zero then one root will lie in the left-hand half of the Z plane and the other will lie in the right-hand half of the Z plane, the only possible exception being both roots lying on the imaginary axis. If we write

$$Z = \zeta - P/2$$

the general quadratic equation becomes

$$\zeta^2 + (Q - P^2/4) = 0$$

and the roots of this equation lie one in the left-half ζ plane and one in the right-half ζ plane (or both along the imaginary axis) as before. Hence, if

Modulus of real part of $P/2 \leqslant$

Modulus of real part of $\sqrt{\dfrac{P^2}{4} - Q}$

the roots of the general quadratic will lie one in the left-half Z plane and one in the right-half Z plane (with the equality sign, one root, or possibly two roots, will lie on the imaginary axis).

Substituting this condition in (19) we have

Modulus of real part of $(z_{22} - z_{11}) \leqslant$

Modulus of real part of $\sqrt{(z_{11} + z_{22})^2 - 4z_{12}^2}$

Now Possenti[5] has recently shown that this relation is necessary for all physically-realizable passive 4-poles. Hence the real parts of the roots of (19) must be of opposite sign with the real part of either or both roots zero as a limiting case.

REFERENCES

[1] L. Brillouin, "Wave Propagation in Periodic Structures." (McGraw-Hill.)

[2] H. B. G. Casimer, "On Onsager's Principle of Microscopic Reversibility." *Rev. Mod. Phys*, 1945, Vol. 17, pp. 343-350.

[3] e.g., K. S. Johnson, "Transmission Circuits for Telephonic Communication." (D. Van Nostrand.)

[4] e.g., E. A. Guillemin, "Communication Networks," Vol. II. (John Wiley.)

[5] R. Possenti, "Quadripoli dissipativi con doppia impedenza iterativa." *Poste e Telecomunicazioni*, April 1949, Vol. 17, No. 4, pp. 225-230.

IMPEDANCE-MEASURING APPARATUS

For use with V.H.F. Aerials

By M. P. Beddoes, B.Sc., Graduate I.E.E.

1. Introduction

MANY types of bridges are available for use in the very-high-frequency (v.h.f.) region, but when used for aerial measurements all suffer from the disadvantage that the amplifier-detector necessary will pick up radiation from the aerial unless very careful precautions are taken. The null detector is usually a radio receiver tuned to the frequency of the source driving the bridge. Raising the output level of the source will help matters, as the sensitivity of the detector can be lowered and its distributed pick-up reduced. Unfortunately, this same raising of the driving power has the effect of heating the bridge elements, possibly to destruction.

The apparatus described is eminently adaptable for high-level operation, but, although a 'balance' is indicated by a minimum signal on the detector, it is not a bridge.

2. Principle of Operation

The apparatus which is described in detail in Section 3 is driven by a transmitter as shown in Fig. 1.

The impedometer is used solely to detect the wave being reflected from the standards unit, whose controls are adjusted for minimum reflected wave both with, and without, the aerial. From these two sets of readings the aerial admittance is immediately available by subtraction.

If $G_1 + jB_1$ is the admittance read off the standards unit controls with the aerial disconnected, and $G_2 + jB_2$ is the admittance read off when the aerial is connected, and $G_A + jB_A$ is the aerial admittance,

then,

$$G_A + jB_A = (G_1 - G_2) + j(B_1 - B_2) \quad \ldots . \quad (1)$$

The full output of 15 W is required from the transmitter only when the standards unit controls are near a 'balance'. There is, therefore, little risk of overloading the detector, and the elements in the standards unit.

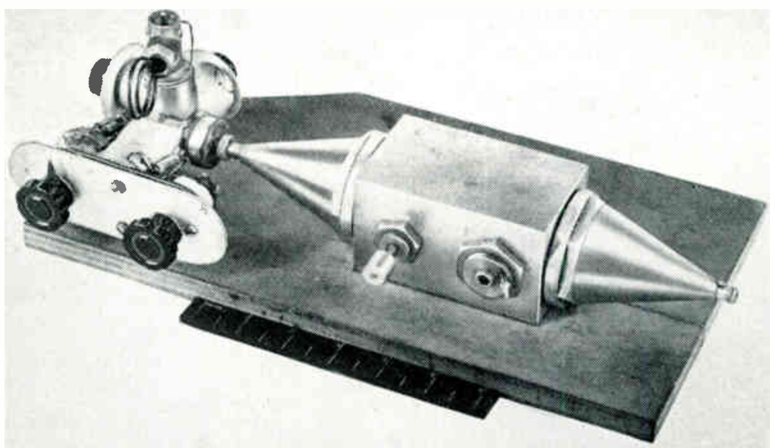


Fig. 2. *Impedometer and Standards Unit.*

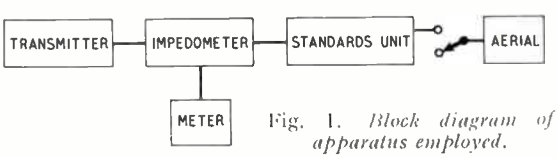


Fig. 1. *Block diagram of apparatus employed.*

MS accepted by the Editor, June 1952.

3. Apparatus

The apparatus consists of a mutilated impedometer or reflectometer (see references 1-5) driving into a standards unit to which may be attached the aerial whose impedance is to be measured. Fig. 1 shows the schematic of the arrangements and Fig. 2 shows a photograph of the impedometer and the standards unit mounted on a wooden board.

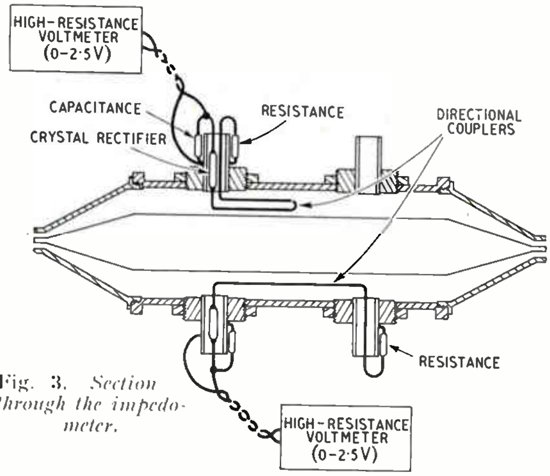


Fig. 3. *Section through the impedometer.*

(a) *The Impedometer*

The apparatus shown in sectional view in Fig. 3 is called the impedometer (after a similar instrument described in reference 3 for measuring impedances).

In its present form, it consists of an opened-out coaxial section, housing directional couplers which are so arranged that they only measure the reflected wave. The ratio, diameter of inner conductor to the internal diameter of the shell, is kept constant to minimize irregularities in the field system.

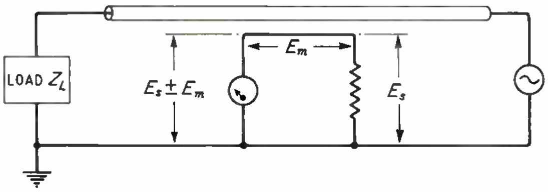


Fig. 4. *This diagram illustrates the action of a directional coupler. E_s and E_m are the voltages produced by the e.s. and e.m. waves.*

The directional coupler, see Fig. 4, is a device for picking up voltages due to the electrostatic (e.s.) and electromagnetic (e.m.) waves and adding them together. For one direction of flow, the voltages due to these waves will be antiphase while, for the other, the voltages will be in phase. The coupler is so adjusted that the voltages induced in it by the e.s. and e.m. waves are equal, and thus it only measures waves travelling in one direction.

The 'pick up' of the directional coupler increases 6 db each time the applied frequency is doubled. While the directional coupler must be large enough to pick up sufficient power to operate a detector satisfactorily, it must not distort the wave system appreciably. To cover a range of frequencies, therefore, it is necessary to employ directional couplers with differing degrees of coupling into the field system. Fig. 5 shows two directional couplers. The left-hand coupler covers the frequency band 30-60 Mc/s, while the other covers 50-100 Mc/s.

To disconnect a coupler from the field system it is only necessary to remove the resistance across which the voltage due to the electrostatic waves is developed. The impedometer shown in Fig. 3, therefore, functions normally over the top half of the frequency band covered by the couplers if the resistance for the larger coupler is removed when measurements are being taken on the smaller. It is not necessary to remove the resistance for the smaller coupler when measurements are taken on the larger, as the smaller coupler will not distort the field at frequencies below its own useful band.

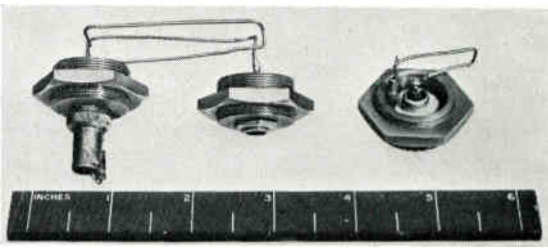


Fig. 5. *Directional couplers for 30-60 Mc/s (left) and 50-100 Mc/s (right).*

(b) *The Standards Unit*

The standards unit consists of parallel elements of resistance, capacitance and inductance. A circuit diagram and photograph of the unit are given in Figs. 6 and 7.

Four controls are available, two for coarse and fine adjustments of resistance and two for coarse and fine adjustments of capacitance. A coil is built on a Pye T junction and is a replaceable item, depending on the frequency used.

The value of the inductance does not need to be known as the capacitance controls are 'centre zero' types and can measure positive and negative capacitances.

The settings of the resistance controls are marked in terms of millimhos, and the capacitance controls in terms of picofarads.

This unit has ordinary carbon resistances which were selected on a d.c. bridge. For accurate work, certain of these resistances have to be checked once a week. Sample resistances were measured in a d.c. bridge, and checked on a Wayne Kerr v.h.f. bridge up to 60 Mc/s. Up to that frequency, the two sets of measurements showed an agreement of better than 1%.

4. Accuracy of Measurement

The percentage accuracy of the admittance of a component under test will vary as the admittance of that component; i.e., the lower the admittance of the component, the less will be the percentage accuracy.

where the ticks (′) indicate values for minimum reflected wave with the aerial connected and their absence denotes values without the aerial. The accuracy of the conductance component will be treated first.

A resistance $1/G_0 = 80$ ohms is left in the circuit, and slow changes with time in its value cannot detract from the accuracy of the result. Changes in value during the reading will impair

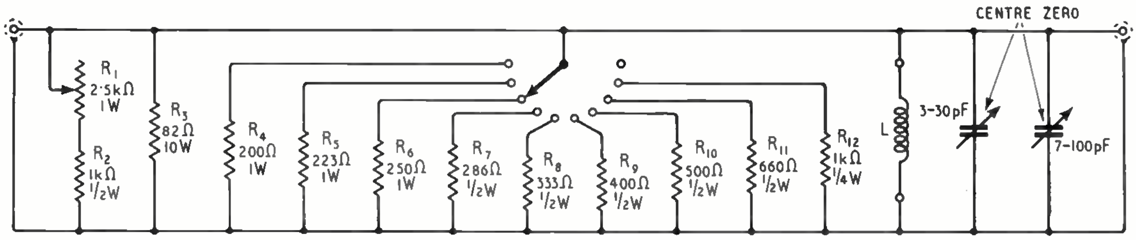


Fig. 6. *Circuit of Standards Unit.*

There are two principal limitations to the accuracy of this instrument—

(a) The sensitivity of the detecting device attached to the directional coupler.

(b) Variations in the values of the components in the standards unit.

(a) An accuracy of ±10% in 1 millimho is possible if a ±½% reflected wave can be detected. A directional coupler feeding into a meter via a rectifier is capable of an accuracy of about this, but a d.c. amplifier between detector and meter should so raise the accuracy that it would be only dependent on the second factor; i.e., on the component variations in the standards unit.

An alternative method of detection using an amplitude-modulated input carrier is worth mentioning. The output from the transmitter (see Fig. 1) is amplitude modulated with a sine wave at some audio frequency. The output from the directional coupler is detected in the usual way employed in a radio receiver. The audio output is then amplified and fed to a loudspeaker or telephones. Absence of a reflected wave will be indicated by zero signal from the amplifier.

Very great sensitivity is possible using this method, so that a small input—less than a watt—is quite adequate. Such a small input would help to keep the variations in the values of the standards components to a minimum.

(b) Variations in the component values in the standards unit due to heat dissipated in them and due to 'ageing' will affect the accuracy of the result.

The formula for aerial admittance [equation (1)] may be rewritten in full (referring to schematic Fig. 8) as follows:

$$GA + jBA = (G_0 + G_1 + G_2) - (G_0 + G_1' + G_2') + j(B_0 + B_1 + B_2) - (B_0 + B_1' + B_2') \quad ..(2)$$

the accuracy, and this must be about ½ of 1%, to achieve an accuracy of ±4% in 1 millimho. This small increase in resistance can be achieved using sufficient carbon resistances in parallel.

Both the slow 'ageing' and rapid variations during the reading time of the resistances composing $1/G_1$ will affect the accuracy. Resistances composing $1/G_1$ are mounted on a rotating turret (as Fig. 9 shows) to keep the shunt capacitance a constant, and this construction also provides very good cooling. The accuracy demanded of the resistances in this unit varies inversely as the resistance on the unit; i.e., the lowest value, 200Ω, must be held to ±1% corresponding to an accuracy of ±5% in 1 millimho.

An accuracy of ±1% in the carbon resistance

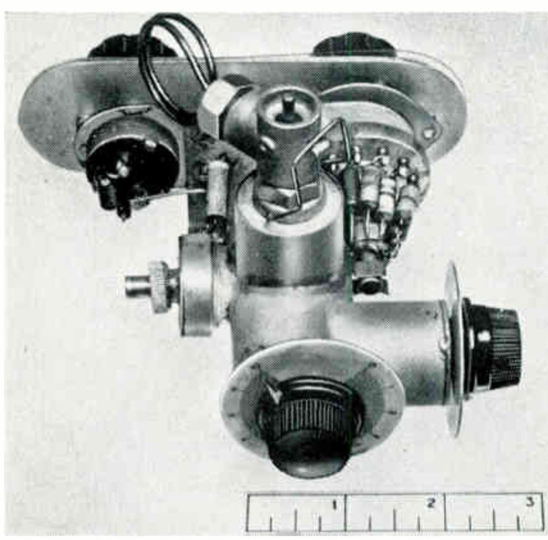


Fig. 7. *Interior of Standards Unit.*

used proved impossible to maintain over any long period of time, and the apparatus was checked weekly, and components which had altered were changed. High-stability carbon resistances would get rid of this necessity

The variations with time of the resistance comprising $1/G_2$ are very slight, and $\pm 2\%$ was allowed for it corresponding to $\pm 1\%$ in 1 millimho. The total variation with heating is $\pm 10\%$ for 1 millimho in the conductance component.

The stray capacitance and the inductance of the coil are contained in the term B_0 of equation (2).

The coil is such a size and Q that it does not warm up when measurements are being taken and change its size and inductance. All the other components in the standards unit have been designed so that the stray capacitance is kept constant. The total variation in B_0 is estimated as being no more than $\pm 0{\cdot}1$ pF during a reading. Any slow variations caused by damage to the coil or any other cause will not impair the accuracy.

The accuracy obtained for the susceptance term is $\pm 0{\cdot}1$ pF. This corresponds to $\pm 4\%$ in 1 millimho at 60 Mc/s.

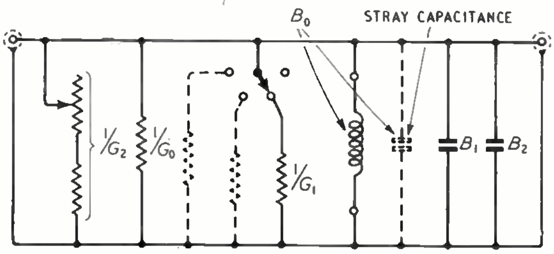


Fig. 8. *Values in Standards Unit in admittance terms.*

The frequency range over which such an instrument could be used is fairly clearly defined. The upper limit is set by the dimensions of the leads in the standards unit and the lower limit is set by the size of the directional coupler. Without undue precautions being taken, it is reasonable to expect a frequency coverage of about 7-100 Mc/s.

Conclusions

An apparatus has been designed for measuring aerial impedance in the range 30-60 Mc/s. This range could be extended to 7-100 Mc/s. The main advantage this method has over previous ones is the detection, which is unaffected by the proximity of a radiating aerial.

The accuracy is inversely proportioned to the size of the aerial impedance and is not worse than $\pm 10\%$ for 1,000 ohms.

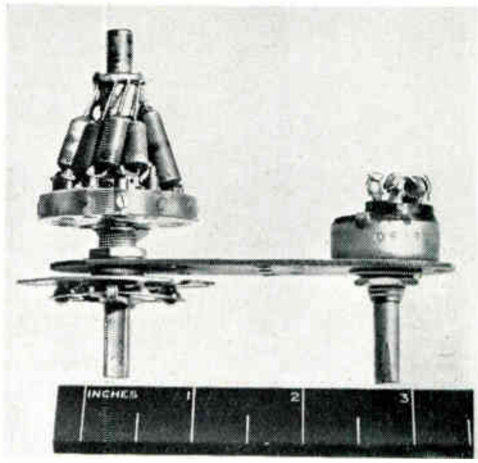


Fig. 9. *Turret mounting of resistors.*

Acknowledgments

I wish to thank Mr. D. H. C. Scholes of The Plessey Company, Ilford, and Mr. E. Forster of the Signals Research and Development Establishment for permission to publish this work.

REFERENCES

[1] H. C. Early. "A Wideband Wattmeter for Wave Guide," *Proc. Inst Radio Engrs*, Oct. 1946, p. 803.

[2] H. C. Early. "A Wideband Directional Coupler for Wave Guide," *Proc. Inst. Radio Engrs*, Nov. 1946, p. 883.

[3] B. Parzen and A. Yalow. "Theory and Design of the Reflectometer," *Electrical Communication*, March 1947, p. 94.

[4] H. R. Allen and C. D. Curling. "The Reflectometer," *Proc. Instn elect. Engrs*, (Part 3), Jan. 1949, p. 25.

[5] B. Parzen. "Impedance Measurements with Directional Couplers and Supplementary Voltage Probe," *Proc. Inst. Radio Engrs*, Oct. 1949, p. 1208.

I.E.E. MEETINGS

9th March. Discussion on "Is the Presentation of Technical Literature Adequate?", to be opened by Professor M. G. Say, Ph.D., M.Sc.

11th March. "Low-Level Modulation Vision Transmitters with special reference to the Kirk o' Shotts and Wenvoe Stations", by E. McP. Leyton, E. A. Nind, B.Sc. (Eng.), and W. S. Percival, B.Sc.(Eng.).

23rd March. Debate—"That Broadcasting Hours should be Drastically Curtailed".

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

BRIT. I.R.E. MEETING

10th March. Discussion on "The Standardization of Symbols and the Arrangement of Electronic Circuit Diagrams", to be opened by L. H. Bainbridge-Bell, M.C., M.A., at 6.30 at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, W.C.1.

INSTITUTION OF ELECTRICAL ENGINEERS

The Council of the Institution has made the 31st Award of the Faraday Medal to Colonel Sir A. Stanley Angwin, K.B.E., D.S.O., M.C., T.D., B.Sc.(Eng.), "for his outstanding contributions to the development of telecommunication in Great Britain and in the international and intercontinental fields".

Sir Stanley was born at Penzance in 1883 and joined the Post Office Engineering Department in 1906. He became Engineer-in-Chief in 1939, but resigned in 1946 to become chairman of Cable & Wireless, Ltd. In 1951 he became chairman of the Commonwealth Telecommunications Board.

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.

Formulae for Ladder Filters

SIR,—In connection with the history of ladder networks, it is interesting to note that the maximally-flat terminated network was given by W. R. Bennett in U.S. Patent 1,849,656 (March 1932). This early and remarkable work foreshadowed the insertion-loss method, developed later by Norton and Darlington.

A. T. STARR.

Cottage Laboratories, Ltd.,
Cobham, Surrey.
13th January 1953.

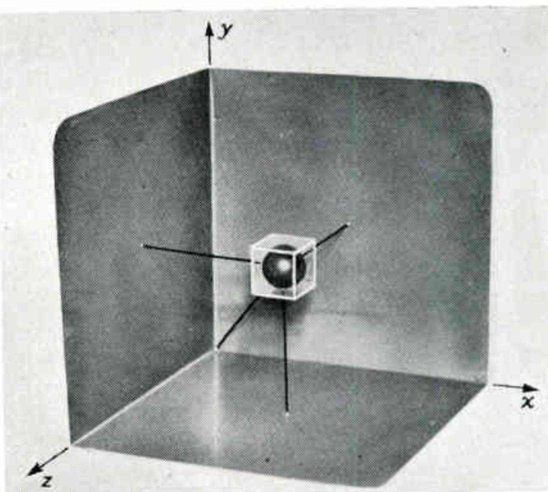
"Ideal Coding" versus "Redundancy"

SIR,—"Shannon's Law" or the "Revised Hartley Law" for the maximum communication capacity of a channel, $C = W \log (1 + P/N)$ is now widely known, but it was remarkable that at the 1952 Symposium on Applications of Communication Theory[1] hardly any attention was given to methods of achieving *this rate with negligible risk of error*, as Shannon[2] states is possible. (With a proviso as to the fidelity criterion for continuous waveforms.) The importance of the topic was, however, indicated by P. H. Blundell, in his paper "On the Definition of Rate of Information in the Presence of Noise," where it was pointed out that the communication rate of a channel cannot readily be defined unless coding be admitted.

It is obvious that the certainty of communication can be increased by employing redundancy, but this does not necessarily increase the *rate* of communication because it requires an increase in either time or bandwidth: the rate will be increased only if the greater TW product is off-set by the higher proportion of correctly received signals. (Incidentally, the result would then depend on whether one accepted the usual convention of counting errors as of zero value, or whether one attached a negative or 'penalty' value to errors: there are circumstances in which the result of an erroneous signal may be far more damaging than the mere deletion of the signal.)

Theorem II in Shannon's paper[2] surely prescribes the following condition. Let M bits per second be the maximum rate of generation of information from a certain source. Let C bits per second be the communication rate of a given channel, according to the Shannon formula quoted above. Then provided $M \leqslant C$ the output of the given information source can (with suitable coding) be transmitted through the given channel with negligible risk of error. This process cannot be said to involve redundancy, because one must ask, "Redundant with respect to what?" The only available standard is the measure M bits per second of the original message, and if the measure C bits per second of the signal in the communication channel is equal to M there is a one-to-one correspondence between message and signal, and therefore no redundancy.

Since Shannon's derivation involves the geometry of non-Euclidean space, the writer prepared the three-dimensional model shown in the accompanying illustration, in the hope that it would encourage further study of Shannon's theorem. It is impossible to review all the implications within this letter, but one point is immediately obvious. The model is of the representational space for signals consisting of three pulses ('digits'), each one of which may initially have any amplitude. Then the point (x_1, y_1, z_1) represents a specific message. Now say that on grounds of noise, only discrete steps of co-ordinate amplitude will be allowed to distinguish different messages—say $\pm\ \delta n$ is the range associated with each variable for a given message. Then the message which initially (i.e., for a noise-free system) was represented by a *point* (x_1, y_1, z_1) now occupies a *domain* $(x_1 \pm \delta n, y_1 \pm \delta n, z_1 \pm \delta n)$ which is a cube of side $2\delta n$ and is represented by the transparent cube in the model. If, for example, the noise has Gaussian amplitude distribution, and the limits $\pm\ \delta n$ are set at $\pm\ 3\sigma$ there is a chance of 2 in 1,000 that the pulse representing each digit in turn will fall outside the δn limit, thus shifting the point into the domain of a different message and causing errors with an average frequency of 6 per 1,000 for the complete 3-digit signals. This is the state of affairs in the absence of 'Shannon-coding.'



Geometric Representation of Three-Dig Signal

The black lines normal to the co-ordinate planes represent the three ordinates corresponding to the three digits of the signal. The transparent cube represents the region of uncertainty if a certain tolerance must be allowed on each digit separately. The black sphere represents schematically the region of uncertainty if the total noise in the whole received signal is considered. The black line from the origin represents the message radius-vector, the square of which is proportional to the sum of the squared digit amplitudes and therefore to the average power associated with this particular message.

But now let the system be so organized that a received-signal point (x, y, z) is associated with a specific message point (x_1, y_1, z_1) not by observing whether each co-ordinate in turn lies within $x_1 \pm \delta n$, $y_1 \pm \delta n$, $z_1 \pm \delta n$ but by investigating the *total distance* $r = \sqrt{\{(x - x_1)^2 + (y - y_1)^2 + (z - z_1)^2\}}$. It seems to the writer that this is the essential point of Shannon-coding. If the noise amplitudes were of equal magnitude (3σ in our example) on each digit (though not necessarily of the same sign) the received-signal point corresponding to the message-point in the model would be shifted along a diagonal of the cube to one of its corners. But if the chance of *one* digit having this noise amplitude or greater is 2 in

1,000, the chance that *three digits consecutively* will exceed this level of noise amplitude is only 8 in 10^9. Therefore the corners of the cube are comparatively free from received signals, as indicated schematically by the black sphere within the transparent cube. Clearly this (a) reduces the risk of total error for a given cubic array of message points, and (b) suggests re-arranging the message-points so as to give a closer packing of message-points in the available space.

The quantitative development of Shannon's theorem depends on the use of hyper-space with the number of dimensions tending to infinity; but it is hoped that this three-dimensional model will suffice to persuade engineers that Shannon-coding involves a principle other than redundancy, and that it is worth considering the derivation of the channel-capacity as set out in Shannon's *I.R.E.* paper[2], the result only being given in his *B.S.T.J.* papers and book. Shannon-coding should in principle be far more powerful than redundancy as a means of combating noise, though the cost of putting it into practice may make it uneconomic in some applications.

D. A. Bell.

Electrical Engineering Department,
University of Birmingham.
26th January 1953.

[1] "Applications of Communication Theory," Butterworth's Scientific Publications, Ltd., 4-6 Bell Yard, Temple Bar, London, W.C.2. To be published shortly.

[2] C. E. Shannon, "Communication in the Presence of Noise," *Proc. Inst. Radio Engrs*, 1949, Vol. 37, p. 10.

NEW BOOKS

Die Ionosphäre (ihre Bedeutung für Geophysik und Radioverkehr)

By Dr. K. Rawer. Pp. 189 with 67 illustrations. P. Noordhoof N.V., Groningen, Holland.

The author of this interesting book is the Scientific Director of the French Ionospheric Prediction Service (S.P.I.M.). His aim is to review the present state of knowledge of the ionosphere and radio-wave propagation from the theoretical and practical points of view.

In the beginning, the general principles of the magneto-ionic theory are summarized. This theory applies to the propagation of radio waves in the ionosphere in the presence of the earth's magnetic field, and it takes into account the observed refraction and polarization effects. The author then describes various techniques of exploration of the ionosphere, such as radio pulse soundings, rocket flights, meteoric observations, spectroscopic analyses, etc. This is followed with an account of theories of formation of ionospheric layers and of relevant ionization processes.

The remaining half of the book is taken up with the description of the regular and irregular variations of the ionospheric layers and the practical problems of short-wave radio communication. Prediction methods and field-strength calculations are dealt with, including a comparison of techniques used by various national organizations.

This book is very readable, as the text is well written and illustrated by many excellent diagrams, but the absence of a classified index is to be regretted. There is also practically no mention of radio astronomy. To summarize, this is a valuable and up-to-date publication for an informed reader generally interested in radio propagation and the ionosphere. Opinion may be, however, divided as to whether a practical user in this country should use it as a textbook. The chapters that have a direct application to his problems are rather brief and some of the procedures described necessitate the use of complex calculating machines. J. P.

Elements of Radio Engineering

By H. I. F. Peel, M.Sc., A.M.C.T., A.M.I.E.E. Pp. 232 + vii. Cleaver-Hume Press Ltd., 42a South Audley Street, London, W.1. Price 10s. 6d.

Covers the syllabi of Radio I, and Telecommunications (Principles) I and II of the City and Guilds' five-year course in Telecommunications.

Applied Electronics Annual 1952

Edited by R. E. Blaise, A.M.Brit.I.R.E. Pp. 240. British Continental Trade Press Ltd., 222 Strand London, W.C.2. Price 40s.

The Use of Radar at Sea

Edited by Capt. F. J. Wylie, R.N.(ret.). Pp. 279 + xv. Produced by the Institute of Navigation. Hollis & Carter, 25 Ashley Place, London, S.W.1. Price 30s.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for January 1953

Date 1953 January	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 1029-1130 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1*	− 0·2	+ 4	NM
2**	—	+ 4	—
3*	NM	+ 4	NM
4	− 0·1	+ 4	+ 2·3
5	− 0·1	+ 5	+ 2·3
6	− 0·2	+ 5	+ 3·3
7*	− 0·2	+ 4	+ 4·2
8*	− 0·1	+ 4	+ 5·2
9*	− 0·1	NM	+ 7·3
10	− 0·1	+ 5	NM
11	NM	NM	+ 6·9
12	− 0·1	+ 4	+ 7·3
13	− 0·1	+ 4	+ 7·9
14*	0·0	+ 4	+ 8·1
15*	+ 0·1	+ 5	+ 9·1
16*	0·0	+ 5	+ 9·0
17	0·0	− 2	+ 9·2
18	0·0	− 1	+ 9·8
19	+ 0·1	− 1	+ 10·0
20*	0·0	− 2	+ 10·7
21*	+ 0·1	− 2	+ 10·1
22	+ 0·1	− 1	+ 10·2
23*	+ 0·1	− 1	+ 10·4
24	+ 0·1	0	+ 10·5
25	+ 0·1	NM	+ 10·6
26**	—	0	—
27*	+ 0·1	− 1	+ 9·7
28**	—	0	—
29*	+ 0·1	− 1	+ 8·8
30**	—	0	—
31	+ 0·2	+ 1	+ 7·9

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

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

NM = Not Measured.

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

** = No MSF Transmission at 1029 or 1429 G.M.T.

ABSTRACTS and REFERENCES

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

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

ACOUSTICS AND AUDIO FREQUENCIES

534.15 599

Analysis of Air Vibrations in a Pipe with Internal Discontinuities.—J. Guittard. (*Acustica*, 1952, Vol. 2, No. 5, pp. 231–236. In French.) A method of investigating the effect of various types of discontinuity, such as cross-section variation, using fine powder as indicator.

534.26 + 535.43 600

On a Multiple Scattering Theory of the Finite Grating and the Wood Anomalies.—Twersky. (See 695.)

534.26 601

The Diffraction of Sound Pulses by an Oscillating Infinitely Long Strip.—F. J. Berry. (*Quart. J. Mech. appl. Math.*, Sept. 1952, Vol. 5, Part 3, pp. 324–332.) A solution is obtained for the case of a plane pressure pulse incident normally on an infinitely long strip of finite width, capable of motion as a spring-supported rigid body. Numerical results are given for the case of a sharp-fronted pulse of constant unit pressure, with graphs showing how the mobility of the strip affects the pressure distribution on the back of it.

534.26 602

The Diffraction of a Sound Pulse by a Non-rigid Semi-infinite Plane Screen.—F. J. Berry. (*Quart. J. Mech. appl. Math.*, Sept. 1952, Vol. 5, Part 3, pp. 333–343.) Analysis is given for the two-dimensional diffraction of a plane-fronted pulse, incident normally on the screen. Two methods of finding the pressure change across the screen are described, both involving successive approximations. Graphs illustrate the results obtained by one method for several types of screen material.

534.321.9 603

Aspects of the Concentration of Ultrasonic Energy.—A. Barone. (*Acustica*, 1952, Vol. 2, No. 5, pp. 221–225.) For another account see 3309 of 1952.

534.321.9 : 534.845 604

Ultrasonic Absorption and Relaxation Mechanism.—A. K. Dutta. (*Indian J. Phys.*, June 1952, Vol. 26, No. 6, pp. 279–282.) Discussion leading to the conclusion that the absorption of elastic waves in liquids is mainly due to a frictional relaxation mechanism.

534.614 605

The Velocity of Sound in Air.—J. M. A. Lenihan. (*Acustica*, 1952, Vol. 2, No. 5, pp. 205–212.) Pulses derived from a 13·5-kc/s transmitter and a microphone receiver were displayed together on a double-beam c.r.o. Coincidence of the pulses could be obtained by movement of the transmitter toward or away from the microphone, the motion being effected by means of an accurate screw of length 1·8 m. Transmitter positions were noted for the 1st–5th and the 41st–45th coincidences, to obtain five readings for the value of 40 wavelengths. Corrections were applied for the effects of temperature, humidity, and other less important factors. The final value deduced for the velocity of 13·5-kc/s sound waves in dry air at 273·16°K and 1 013·2 mb is 331·45 ± 0·04 m/s.

534.84 606

Review of Methods and Apparatus for Room-Acoustics Measurements.—L. Keidel. (*Arch. tech. Messen*, Sept. 1952, No. 200, pp. 193–196.)

534.84 : 621.396.712.3 607

The 'Pierre Bourdan' Low-Frequency [broadcasting] Centre, Paris.—Conturie. (See 825.)

534.845.1/.2 608

Sound Insulation by means of Rubber and Steel Springs.—M. L. Exner. (*Acustica*, 1952, Vol. 2, No. 5, pp. 213–221. In German.) Good agreement was obtained between theory and experiment in measurements on combinations of rubber and steel springs. The high internal damping of the rubber, whose loss factor is about 10 times that of steel springs, results in good insulation over the whole frequency range investigated, 20 c/s–2 kc/s. Mechanical filters, consisting of masses with intervening springs, were also investigated; they were found better than simple systems above a certain critical frequency, but had disadvantages at low frequencies.

534.845.1/.2 609

The Mechanism of Sound Transmission through Single-Leaf Partitions, investigated using Small-Scale Models.—A. Schoch & K. Fehér. (*Acustica*, 1952, Vol. 2, No. 5, pp. 189–204.) A summary is given of Cremer's results (2904 of 1951) for transmission through plates of infinite area, and the effect of the boundaries is discussed qualitatively for the case of finite plates. Measurements are reported, using small-scale models with corresponding high-frequency plane waves in an echo-free box of dimensions 150 × 80 × 80 cm. Results for plates of various materials, for both normal and oblique incidence, are shown graphically and discussed.

534.846 610
Auditorium Specifically Designed for Technical Meetings.—D. M. Beard & A. M. Erickson. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1952, Vol. 59, No. 3, pp. 205–211.) Description of the auditorium, seating 550, at the Naval Ordnance Laboratory, White Oak, Md. The effects of polycylindrical sections, absorbent plaster, serrated rear wall, and padded seats, combine to give excellent acoustic characteristics. Additional facilities include 24 microphones distributed about the room, controlled lighting, optical-projection booth, and an adequate telephone communication system.

534.861.4 : 621.395.623.7 611
The Environment of High-Quality Reproduction.—F. H. Brittain. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, pp. 2–5.) A survey of reproducer circuit requirements and auditorium conditions to be satisfied in order to obtain the full benefit of good loudspeaker performance.

621.395.6 : 621.396.712 612
Speech Input Systems for Broadcast Transmitters.—Hill. (See 824.)

621.395.61 613
On the Directivity of Spherical Microphones.—W. Kuhl. (*Acustica*, 1952, Vol. 2, No. 5, pp. 226–231.) The directivity patterns of microphones with plane circular diaphragms on the surface of a sphere were measured for 10 frequencies and compared with the patterns computed by Schwarz (2992 of 1945) for point microphones. Uniform response for all angles of incidence can be obtained by adding the voltages derived from two point microphones located at opposite ends of a diameter of a sphere. A frequency-response curve is calculated for a capacitor microphone with spherical diaphragm.

621.395.623.7 614
Friction-driven Loudspeaker.—(*Wireless World*, Jan. 1953, Vol. 59, No. 1, pp. 27–28.) A loudspeaker whose operation depended on the attraction between a metal surface and a poor conductor (e.g., agate) was designed by Johnsen & Rahbek over 30 years ago. A modern commercial public-address loudspeaker using the same principle is described, in which a metal band attached to the diaphragm is held in contact with a rotating cylinder coated with a semiconductor. The power output is comparable with that of an amplifier-driven loudspeaker of similar size.

621.395.623.8 615
P.A. [public-address] **Systems in Generating Plants.**—S. C. Bartlett. (*Radiotronics*, Oct. 1952, Vol. 17, No. 10, pp. 159–164.) The special conditions encountered in power stations are discussed, and the design is considered of suitable equipment to provide adequate speech coverage to all personnel, with reply or break-in from any point of the system.

789.983 616
The Chord Organ.—A. Douglas. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 562–566.) Description of a Hammond instrument which produces, if desired, a chord when any single note is pressed. Tone controls are also provided.

AERIALS AND TRANSMISSION LINES

621.315.2 .3 617
New Cables and Conductors of the I.K.A. [Verwaltung für Installationen, Kabel, Apparate].—H. Göttlich & H. J. Franz. (*Deutsche Elektrotech.*, Sept. 1952, Vol. 6, No. 9, pp. 429–432.) Data and illustrations of various cables, including flexible types, for counter-circuit, television, h.f. and u.h.f. applications.

621.392 + 621.315.212].018.44 618
Mathematical Theory of Laminated Transmission Lines: Part 2.—S. P. Morgan, Jr. (*Bell Syst. tech. J.*, Nov. 1952, Vol. 31, No. 6, pp. 1121–1206.) Part 1 (25 of January) dealt mainly with Clogston-1 lines; the present paper deals mainly with Clogston-2 lines, which are composed entirely of laminated material; both parallel-plane and coaxial lines are considered. Formulae are also derived for the general case of a line with arbitrary fractions of space occupied by the main dielectric and the laminations. Analysis is given for the principal and higher modes, assuming infinitesimally thin laminae; the effect of finite thickness is considered subsequently. The influence of nonuniformity of the laminations is examined; to achieve an attenuation constant of the order of a tenth that of a conventional line, nonuniformities must be smaller than a few parts in 10 000. Dielectric and magnetic losses are discussed; their magnitude is directly proportional to frequency provided the loss tangents do not vary with frequency.

621.392.21 619
Calculation of Transmission-Line Constants.—R. O. Kapp. (*Engineering, Lond.*, 5th Sept. 1952, Vol. 174, No. 4519, pp. 315–316.) Approximate formulae are derived for the three line constants, usually represented by the symbols **A**, **B** and **C**. Errors involved in the use of these formulae are $<0{\cdot}7\%$ for either the real or the imaginary component of any of the constants for line lengths up to about 600 miles. The formulae do not involve hyperbolic functions.

621.392.21.09 620
Application Possibilities of a Surface-Wave Mode.—W. F. Gunn. (*Marconi Rev.*, 4th Quarter 1952, Vol. 15, No. 107, pp. 145–166.) A summary is given of recent work on surface waves, together with a short account of experiments carried out in 1951 on a go-and-return system of total length about 130 ft. Relevant analysis from various sources is included in appendices.

621.392.21.029.64 621
Microstrip — A New Transmission Technique for the Kilomegacycle Range.—D. D. Grieg & H. F. Engelmann. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1644–1650.) A general description and background theory are given for transmission lines comprising a single wire conductor arranged parallel to a ground plane and for flat-strip lines of the type dealt with in 2705 of 1952 (Barrett). Practical methods of making such lines include printing and embossing.

621.392.21.029.64 622
Simplified Theory of Microstrip Transmission Systems.—F. Assadourian & E. Rimai. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1651–1657.) An analysis is made of TEM-mode propagation in a line comprising a wire or a finite-width strip immersed in a uniform dielectric and arranged parallel to a ground plane. Characteristic impedance, power flow and losses are considered. Numerical calculations based on the theory indicate the practicability of lines of this type at microwave frequencies.

621.392.21.029.64 : 621.317.3 623

Microstrip Components.—J. A. Kostriza. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1658–1663.) Standing-wave detectors are described suitable for making measurements on lines of the single-conductor-and-ground-plane type dealt with in 621 above. The deviations from pure TEM-mode propagation for different constructions are assessed on the basis of the measured values of guide wavelength. Dispersion and r.f. impedance are also discussed. Voltage s.w.r. measurements are reported on components for effecting transitions to coaxial lines. Attenuator pads and loads, crystal mounts, directional couplers, etc., using the same constructional basis are described.

621.392.26 624

Calculation of the Propagation Constants of an Inhomogeneously-Filled Waveguide.—J. A. Bradshaw: L. G. Chambers. (*Brit. J. appl. Phys.*, Oct. 1952, Vol. 3, No. 10, pp. 332–333.) Comment on 2114 of 1952 and author's reply.

621.392.26 625

Diffraction of Guided Waves at Plane Diaphragms.—R. Müller. (*Z. Naturf.*, Nov. 1950, Vol. 5a, No. 11, pp. 617–621.) The problem of diffraction at a plane diaphragm is formulated for the general case of a waveguide of arbitrary cross-section. For the special case of a coaxial aperture in a circular cylinder, a wave suffers no transformation from E to H mode or vice versa if incident parallel to the axis, but does suffer transformation if incident in any other direction.

621.392.26 626

The Concept and Measurement of Impedance in Periodically Loaded Wave Guides.—E. T. Jaynes. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1077–1084.) The theory of the node-shift method of investigating transmission-line problems is reviewed. A definition of impedance in a form applicable to the disk-loaded waveguide is developed from ordinary circuit theory by expressing the e.m. field in terms of a set of independent component fields. An extension of the node-shift technique for measuring impedance is described which involves a determination of the parameters of a coupling system.

621.396.67 627

Mutual Radiation Resistance of Aerials and Arrays.—L. Lewin. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 24–25.) Comment on 32 of January (Knudsen).

621.396.67 628

Theory of Electrically Short Transmitting and Receiving Antennas.—R. King. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1174–1187.) Centre-driven cylindrical aerials of length $\leqslant \lambda/\pi$ are considered; complete quantitative solutions are obtained by determining the distributions of current that actually satisfy the integral equations. Components of current in phase and in quadrature with the driving voltage are evaluated, together with the impedance, the effective length and the gain. Quite accurate results are obtained even in a first-order solution when the King-Middleton method of solving Hallén's integral equation by iteration is applied correctly. The newly determined values are combined with the King-Middleton second-order results to obtain more accurate values for aerials of length $\leqslant 1{\cdot}4\,\lambda/\pi$.

621.396.67 629

Cylindrical Aerials.—R. W. P. King. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, p. 24.) Comment on 2715 of 1952 (Storm).

621.396.67 : 621.317.336 630

An Antenna Impedance-Measuring Instrument.—J. F. Cline. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1686–1689.) An indicating instrument is described which has a small capacitive loading effect when connected directly to the terminals of an aerial. This is achieved by isolating those conductive components which are not part of the r.f. circuit and operating the aerial as a receiving aerial, so that the signal generator is some distance away and not incorporated in the measuring instrument.

621.396.67 : 621.397.6 631

The WJZ-TV Auxiliary Antenna.—J. Preston. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 38–39.) The emergency WJZ television aerial array consists of four asymmetrical corner reflectors, with $\lambda/2$-dipole feed, uniformly spaced round the conical portion at the top of the Empire State Building and set askew in order to prevent the occurrence of deep nulls in the radiation pattern.

621.396.67 : 621.397.6 632

Aerials of Modern High-Power Television Stations.—G. Rutelli. (*Alta Frequenza*, Aug./Oct. 1952, Vol. 21, Nos. 4/5, pp. 215–216.) Short discussion of the radiation characteristics of the aerials at the Sutton Coldfield and Holme Moss stations, with reference to the theory of that type of aerial previously published by the writer (3331 of 1941 and 436 of 1942).

621.396.67.001.11 : 517.948.32 633

Difficulties with Present Solutions of the Hallén Integral Equation.—S. H. Dike. (*Quart. appl. Math.*, Oct. 1952, Vol. 10, No. 3, pp. 225–241.) Dike & King (2716 of 1952) have found serious discrepancies between experimental values of broadside absorption gain and back-scattering cross-section and those calculated by the King-Middleton modification (1453 of 1946) of Hallén's first-order solution for a cylindrical aerial. These discrepancies are here discussed in detail and the problem is re-examined, reference being made to the published results of many investigators. A theory which lends itself to practical computation of the complete characteristics of a simple dipole aerial does not at present seem to exist. It is significant that the results of Van Vleck et al. (3035 of 1947) for the back-scattering cross-section of a shorted dipole agree more closely with experiment than the first-order solutions of Hallén, King & Middleton, or Gray (1931 of 1944). Variational methods of solving Hallén's equation have given results which are not satisfactory in some respects. It is considered that it might be worth while to follow up a suggestion made by Brillouin (790 of 1945) that the known function and the kernel of the integral equation be expanded in Fourier series with known coefficients, and that the unknown function for the current be expanded likewise with unknown coefficients. Term-by-term integration would then lead to a set of simultaneous equations for determining the coefficients. Results of such an approach do not appear to have been published.

621.396.676 : 623.74 634

Flush-Mounted Antennas for Military Aircraft.—(*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 58–59, 111.) Illustrations of zero-drag types of aerial developed for jet-driven fighters and high-speed bombers and operating at frequencies from 100 Mc/s to 1·25 kMc/s. Their functions include distance measurement, communications, landing approach, navigation, and interrogation.

621.396.677 635

Approximate Determination of Aerial Gain.—S. Giustini. (*Alta Frequenza*, Aug./Oct. 1952, Vol. 21, Nos. 4/5, pp. 204–214.) When side lobes are negligible, the field-strength distribution of a directive aerial can be represented by a single lobe generated by rotation of the polar curve $E = E_0e^{-h\phi^2}$, where E_0 is the maximum field strength and h a matching parameter. The following approximate expressions are derived for the gain: $G \approx 8h : G \approx 2{\cdot}75/(\phi')^2$, where ϕ' is the angle (in radians) corresponding to a field strength of $E_0/\sqrt{2}$.

621.396.677.029.62 **636**

U.S.W. Wide-Band Directive Aerial.—H. Bosse. (*Fernmeldetech. Z.*, Oct. 1952, Vol. 5, No. 10, pp. 437–439.) Description of an aerial system consisting of a vertical stack of four pairs of horizontal $\lambda/2$ dipoles. The dipoles of each pair are arranged in line, and behind the whole system at a distance of about 0·3 λ is a polarizing reflector grid of 20 horizontal rods. Effective bandwidth is 27 Mc/s, centred on 54·5 Mc/s.

621.396.67 **637**

Antennas: Theory and Practice. [Book Review]—S. A. Schelkunoff & H. T. Friis. Publishers: J. Wiley & Sons, New York, 1952, 593 pp., $10. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, p. 1742.) A complete text, intended for students, radio engineers and applied mathematicians and physicists; the mathematics does not go beyond the calculus.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.014.1 : 537.311.2 **638**

Ohm's Law for Build-up Phenomena.—V. Kussl. (*Funk u. Ton*, Oct. 1952, Vol. 6, No. 10, pp. 527–533.) In the case of 2-pole networks, the Laplace transformation is applied in the form of a Fourier integral of the input waveform. Doetsch's symbolic notation for asymmetrical transformations (3450 of 1948) is used, and the equality sign of Ohm's law is replaced by the Doetsch transformation sign. The variation with time of a current through a complex impedance is given by the Laplace transform product of the susceptance and the applied-voltage spectrum. Application of the transformation in quadripole theory is described. By formulating the quadripole equation in chain-matrix form, transient-response parameters can be roughly estimated. Application of the theory to the determination of transient distortion in communication systems and to the stability testing of amplifiers is outlined.

621.3.015.7 : 621.387.4 **639**

Pulse-Amplitude Analysis in Nuclear Research.—A. B. Van Rennes. (*Nucleonics*, July–Oct. 1952, Vol. 10, Nos. 7–10, pp. 20–27, 22–28, 32–38 & 50–56.) Various voltage-discrimination techniques are discussed in detail and descriptions are given of a simple type of analyser and a moderate-precision and a high-precision analyser. Analysers are also discussed in which pulse-height selection is effected either by mechanical means or by use of diode valves, trigger circuits, sorting-ladder circuits, or beam-deflection techniques. Other types of analyser described include those in which height selection is accomplished (*a*) by an expander-amplifier driving a chain of discriminators, (*b*) by conversion of pulse amplitude to pulse duration, (*c*) by means of information-storage devices. 55 references.

621.314.25 **640**

Low-Cost Variable Phase Shifter.—S. Wald. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 168 . . 180.) The basis of the circuit is a linear resistance potentiometer divided by tappings into four equal sections, neighbouring sections being fed, by a pair of transformer secondaries, with voltages phase-separated by 90°. The value of the potentiometer resistance is not critical, providing it is large compared with the impedance of the transformer secondaries.

621.314.7 : 621.396.6 **641**

Dynamics of Transistor Negative-Resistance Circuits.—B. G. Farley. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1497–1508.) A general method is presented for calculating approximately the characteristics of nonlinear circuits. The region of operation is divided into subregions, within each of which the circuit may be considered as nearly linear. The method is applied to (*a*) analysis of a high-speed switching circuit using a point-contact transistor, (*b*) discussion of negative-resistance relaxation oscillations, (*c*) calculation of waveforms and rise times of a regenerative transistor amplifier [769 below (Felker)].

621.314.7 : 621.396.6 : 512.831 **642**

Matrix Representation of Transistor Circuits.—J. Shekel. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1493–1497.) Transistor circuits are discussed in terms of measurable quantities only. Once the admittance matrix of the grounded-base transistor has been determined, the matrices of the grounded-emitter and grounded-collector transistors can easily be derived. Matrix representation also provides a direct method for analysis of stages in cascade.

621.314.7 : 621.396.6 : 621.396.822 **643**

Transistor Noise in Circuit Applications.—H. C. Montgomery. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1461–1471.) Problems of linear circuits involving multiple noise sources can be handled by familiar methods with the aid of certain noise-spectrum functions, which are described. Several theorems of general interest in circuit work, dealing with noise spectra and noise correlation, are derived. The noise characteristics of transistors can be described in terms of the spectrum functions for simple but arbitrary configurations of equivalent noise generators. From these, the noise figure can be calculated for any external circuit. Numerical results for a number of *n-p-n* transistors are given in a table and many curves.

621.314.7.012.8 **644**

Junction-Transistor Equivalent Circuits and Vacuum-Tube Analogy.—L. J. Giacoletto. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1490–1493.) A comparison is made between the operating characteristics of a *p-n-p*-junction transistor and a triode valve, using a Π network to represent the transistor.

621.316.8 + 621.318 + 621.319.4].001.8 **645**

Nonlinear Circuit Elements in High-Frequency and Low-Frequency Technology.—H. E. Hollmann. (*Arch. elekt. Übertragung*, Oct.–Dec. 1952, Vol. 6, Nos. 10–12, pp. 434–440, 478–486 & 520–531.) A review of the properties and applications of nonlinear inductors, capacitors and resistors of many different types. See also 3039 of 1952.

621.316.842 : 621.316.7 **646**

New Commercial Barretters.—J. Sommer. (*Funk u. Ton*, Oct. 1952, Vol. 6, No. 10, pp. 520–526.) The characteristics of two tungsten-coil barretters are given. The resistances when cold are respectively 50Ω and 100Ω; a current of about 5 mA doubles the resistance in each case. Tests to determine the cooling characteristics and the effect of ambient-temperature changes are described.

621.318.4 **647**

Coil Winding Data.—L. Knight. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, p. 22.) Charts give the number of turns required in conjunction with various values of capacitance for tuning over the frequency ranges 2–70 Mc/s and 70 kc/s–3 Mc/s, using standard formers with dust cores.

621.318.57 **648**

New Bistable High-Speed Multi-Purpose [switching] **Device.**—Y. Druet. (*C. R. Acad. Sci., Paris*, 25th Aug. 1952, Vol. 235, No. 8, pp. 494–496.) Operating conditions for an Eccles-Jordan circuit using two Type-ECH42 triode-hexode valves are noted. With suitable diode limiting, consistent operation is maintained at pulse rates $> 2 \times 10^6$/sec.

621.318.57 : 621.3.015.7 **649**

Arithmetical Counters for Pulses.—A. Dauphin. (*Onde élect.*, Nov. 1952, Vol. 32, No. 308, pp. 459–463.) A list is given of 51 relevant publications, with short notes indicating their scope.

621.318.57 : 621.314.7 **650**

A Transistor Reversible Binary Counter.—R. L. Trent. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1562–1572.) The counter is built of elementary transistor packaged units. The mechanism used to achieve reversibility and the circuit for each type of building block are described.

621.318.57 : 621.314.7 **651**

Transistor Trigger Circuits.—A. W. Lo. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1531–1541.) Analysis is presented for transistor trigger circuits which permits prediction as to whether operation will be monostable, bistable or astable (oscillatory) and also of the amplitude and waveform of the output. Practical pulse circuits for various purposes are described which are so designed that their operation is not affected by reasonable variations of circuit parameters, bias voltages, transistor characteristics, or ambient temperature.

621.318.57 : 621.314.7 **652**

Transistors in Switching Circuits.—A. E. Anderson. (*Bell Syst. tech. J.*, Nov. 1952, Vol. 31, No. 6, pp. 1207–1249. *Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1541–1558. Correction, *ibid.*, Dec. 1952, Vol. 40, No. 12, pp. 1732–1733.) Analysis of transistor trigger circuits is based on an approximate representation of the negative-resistance characteristic by three straight lines. Circuits using point-contact transistors for waveform generation, level restoration, delay, storage, and counting are described, and their properties and limitations are discussed in detail.

621.318.57 : 621.314.7 : 518.4 **653**

Graphical Analysis of Some Transistor Switching Circuits.—L. P. Hunter & H. Fleisher. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1559–1562.) Methods are described for generating the entire input characteristics for the various terminals of a transistor, and graphical methods of analysis are applied to (*a*) a base-input amplifier, (*b*) a collector-to-emitter direct-coupled switching circuit, (*c*) a collector-to-base direct-coupled circuit.

621.318.572 **654**

Electronic Switch.—K. R. Sturley. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, pp. 11–14.) A switching circuit for two-waveform display with a single-beam c.r.o. uses a conventional multivibrator for generating the switching voltages.

621.392 **655**

Network Analysis by Repeated Voltage Superposition.—J. E. Parton. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 570–574.) A method of analysis is described, with worked-out examples, which reduces considerably the number of simultaneous equations to be solved for an *m*-mesh network. The method essentially involves successive applications of Thévenin's theorem, each application reducing by one the number of meshes in the network considered, with a corresponding reduction of the number of simultaneous equations to be solved. A similar method has been described by Tasny-Tschiassny (3365 of 1948), who used 'residual' current generators at the final nodes instead of residual voltage generators in the final branches, and whose method differs from the present method in other details.

621.392 **656**

Synthesis of Cascaded Three-Terminal *RC* Networks with Minimum-Phase Transfer Functions.—P. F. Ordung, F. Hopkins, H. L. Krauss & E. L. Sparrow. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1717–1723.) For the realization of a particular transfer function including complex zeros, the method of synthesis presented yields a network with fewer elements, simpler configurations and higher level of transmission than previous methods (e.g., 1605 of 1950).

621.392.1 **657**

The Practical Significance of Complex Frequencies in Electrical Communication Engineering.—J. Peters. (*Arch. elekt. Übertragung*, Oct. 1952, Vol. 6, No. 10, pp. 401–413. Correction, *ibid.*, Dec. 1952, Vol. 6, No. 12, p. 514.) A concise general introduction to the subject. Complex frequencies are complex quantities whose real part represents the gain of a system and whose imaginary part represents the frequency in the usual sense. The application of the Laplace transform in analysis of transmission problems is explained and the properties of the complex plane and its poles and zeros are described. The use of the poles and zeros for representation of the transmission characteristics of a linear network is considered. Analogues of the complex plane, such as that obtainable with a stretched elastic membrane, are discussed and their applications illustrated.

621.392.4/.5 **658**

Anode-Follower Derivatives.—A. W. Keen. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 5–9.) "Low output-impedance stages characterized by anode output and feedback of the entire output voltage are derived from the basic 'anode-follower' by substitution of a valve impedance for the shunt resistor of the feedback path, or of a comparator stage for the entire input-feedback potential divider, and by replacement of the output valve by a series-connected push-pull pair. These developments suggest the possibility of obtaining an anode-follower analogue of each cathode-follower derivative, thereby increasing the number of circuit variants available for practical use."

621.392.4/.5 : 512.972 **659**

Applications of Tensor Theory to Linear Electronic Circuits.—A. Kaufmann. (*Radio tech. Dig., Édn franç.*, 1952, Vol. 6, Nos. 2, 3 & 4, pp. 67–76, 157–168 & 199–209.) An explanation of tensor concepts and their general application to valve circuits, with examples illustrating the determination of input impedance and gain of feedback, cathode-follower, and grounded-grid circuits.

621.392.5 **660**

On the Approximation Problem in Network Synthesis.—A. D. Bresler. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1724–1728.) A method of synthesis is presented in which a desired frequency-response characteristic is replaced by an approximation consisting of a sequence of rectilinear segments. The method is illustrated by application to the design of attenuation equalizers with constant-resistance ladder sections.

621.392.5 **661**

The Parallel-T Resistance-Capacitance Network.—L. G. Cowles. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1712–1717.) Analysis is given for the general case of finite source and load resistances. When these two resistances are appropriately related the network loss is the same at low and high frequencies; the transfer characteristic is then a circle in the complex plane. This 'symmetrical' network is equivalent to a simple series-resonant circuit as regards its transfer characteristic.

621.392.5 **662**

RC Cathode-Follower Feedback Circuits.—S. C. Dunn. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 10–19.) When certain *RC* circuits are associated with a cathode-follower circuit, a voltage gain can be obtained over a band of frequencies. A number of prototype and derived *RC* circuits which can be used in this way are analysed and their response curves are determined. The matrices of the derived networks can be formed easily from those of the corresponding prototypes. The effect of circuit termination on response is discussed and the use of these circuits as feedback elements in amplifiers is considered in some detail. Experiments in confirmation of the theory are mentioned.

621.392.5 **663**

Resistance-Capacitance Networks with Over-Unity Gain.—W. Bacon & D. P. Salmon. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 20–23.) Longmire (2702 of 1947) and Epstein (2940 of 1951) have described *RC* circuits with greater output than input voltage. A method of increasing the voltage gain by feeding the output voltage of one network into a second network is described. The process cannot be extended indefinitely, the voltage gain attainable being limited by the impedance increase necessary at each stage. Experiments confirmed the theory. Such networks can be used with a cathode follower to construct oscillators in which the valve gain is less than unity. An experimental circuit is described.

621.392.5 **664**

Response of a Linear Network to an Input with Linearly Variable Frequency as obtained in Sweep-Frequency Testing.—H. Ekstein & T. Schiffman. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 454–471.) Application of an input voltage with linearly increasing frequency to a linear network produces an output curve which, for very slow frequency variation, is an image of the admittance plotted as a function of frequency. When the rate of frequency variation is not negligible, the response curve is a distorted image of the admittance. The nature of this distortion is investigated. A quantitative approximation method is presented which uses the theory of functions for evaluation of an integral. The result is expressible in terms of elementary functions. Explicit expressions are given for the dynamic corrections to be applied to the 'observed' resonance frequencies and peak admittances. The method is applied to several simple circuits in addition to the *LCR* circuit which has been previously treated by other authors.

621.392.52 **665**

Fundamentals of Filter Theory and Technique.—K. H. Haase. (*Funk u. Ton*, Oct. 1952, Vol. 6, No. 10, pp. 505–519.) Description of the application of the wave theory and the operating-parameter theory in the design and calculation of different basic types of filter.

621.392.52 **666**

Formulae for Ladder Filters.—H. J. Orchard. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 3–5.) Four related sets of explicit formulae for the elements of a basic low-pass filter network are presented with a common notation. Three of the sets have been published previously [2900 of 1937 (Norton); 1543 of 1952 (Bosse); 1541 of 1952 (Belevitch)], the fourth being new; all have reference to the image-parameter theory due to Norton and Darlington (1361 of 1940). If corresponding formulae could be found for a general type of filter with response depending on Jacobian sn elliptic functions, they would represent a considerable contribution to filter-design technique.

621.392.52 **667**

A Nonlinear Statistical Filter.—A. W. Sullivan & J. M. Barney. (*Proc. nat. Electronics Conf., Chicago*, 1951, Vol. 7, pp. 85–91.) A method is described for using the statistical differences between a periodic rectangular pulse (the wanted signal) and the envelope of fluctuation noise (the interfering signal), for the purpose of discriminating between the two signals. Expressions are derived for the density distribution probability of noise and of signal plus noise. A description is given of a practical filter which was checked experimentally under conditions when the signal was completely masked by the noise, both aurally and visually, but could be reliably detected by means of the statistical filter. Possible application to secret-transmission systems is discussed.

621.392.52 : 518.12 **668**

Numerical Calculation of Filter Circuits with Tchebycheff Characteristics after the Method of W. Cauer.—V. Fetzer. (*Arch. elekt. Übertragung*, Oct. 1952, Vol. 6, No. 10, pp. 419–431.) Formulae for Tchebycheff-type filters are derived from those previously given for filters with generalized parameters (1545 of 1952). Three types of antimetrical filter are considered. The formulae necessary for numerical calculation of both symmetrical and antimetrical normalized filters are derived and applied to examples. By suitable frequency transformation the normal type of Tchebycheff filter can be dealt with. A complete set of curves is provided for evaluation of the required parameters.

621.392.52 : 621.315.212 **669**

Coaxial Transmission-Line Filters.—D. E. Mode. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1706–1711.) Analysis for coaxial-line band-pass filters is given based on TEM-mode transmission. The influence of the nature and spacing of the obstacles on the bandwidth is discussed. Measurements on filters of various constructions support the theoretical results. A TE-mode high-pass filter is also discussed. For narrow-band applications this type of filter is inferior to the cavity type.

621.392.52 : 621.396.49 **670**

Nonlinear Filtering and Waveshape Multiplexing.—R. E. Scott, S. Fine & A. Macmullen. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 146, 148.) A method of providing two channels on one carrier consists of differentiating the combined signal, clipping to remove one component, and re-integrating to restore the other. The removed component is recovered by subtracting the retained component from the original input. Experimental circuits and waveforms obtained are illustrated, the two components being respectively a sine wave and a square wave.

621.394/.396].6 : 003.63 **671**

Functional Circuit Diagrams.—C. E. Williams. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, pp. 19–21.) Shortened version of paper abstracted in 63 of January.

621.396.6 : 621.314.7 : 629.13 672

Transistors in Airborne Equipment.—O. M. Stuetzer. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1529–1530.) Discussion of the advantages of using transistors instead of valves in aircraft equipment.

621.396.6 : 621.317.755 673

Slow-Speed Circular Timebase.—(*Radio tech. Dig., Édn franç.*, 1952, Vol. 6, No. 4, pp. 179–193.) French version of paper by Hardie & Thomas (2755 of 1952) supplemented with references and a note on earlier types of sine-cosine potentiometer.

621.396.6.002.2 : 621.314.7 674

Printed Circuitry for Transistors.—S. F. Danko & R. A. Gerhold. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1524–1528.) The auto-sembly technique [355 of February (Danko)] is suggested as a simple and effective method for the production of compact transistor circuits.

621.396.611.1 675

Resonance Curves.—'Cathode Ray'. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, pp. 29–33.) Deviations encountered in practical circuits from the ideal conditions dealt with in elementary theory are simply explained.

621.396.611.1 : 621.396.822 676

Spontaneous Voltage Fluctuations in a Resonant Circuit.—E. Paolini. (*Alta Frequenza*, Aug./Oct. 1952, Vol. 21, Nos. 4/5, pp. 199–203.) The r.m.s. value of voltage fluctuations, of thermal origin, at the terminals of a parallel RLC circuit, is determined for the unusual case where the frequency range in question is not negligible in comparison with the bandwidth of the resonant circuit. Curves are given to facilitate numerical calculation.

621.396.615 + 621.396.645] : 621.314.7 677

A Junction-Transistor Tetrode for High-Frequency Use.—R. L. Wallace, Jr, L. G. Schimpf & E. Dickten. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1395–1400.) Application of suitable bias to a fourth electrode connected to the *p*-type central section of an *n-p-n*-junction transistor causes a considerable reduction of the base resistance. This reduction enables the transistor to be used at frequencies over ten times the normal limit. Circuits are given for a variable-frequency tuned amplifier, a band-pass amplifier, and a sine-wave oscillator for 40–115 Mc/s.

621.396.615 : 621.314.7 678

Transistor Oscillators.—W. Herzog. (*Arch. elekt. Übertragung*, Oct. 1952, Vol. 6, No. 10, pp. 398–400.) The characteristics of transistor oscillators are established by analysis in which the conductance form of the transistor equations is used. The following circuits are considered: (*a*) transistor with frequency-determining quadripole and overall feedback; (*b*) as in (*a*) but with transformer coupling between transistor and quadripole; (*c*) transistor oscillator with a Π-arrangement of three parallel-connected tuned circuits.

621.396.615.029.4 : 621.314.7 679

Low-Drain Transistor Audio Oscillator.—D. E. Thomas. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1385–1395.) Description of the design and performance of a 130-c/s oscillator using a Western Electric Type-A1768 point-contact transistor supplied from a single 6-V battery, the power drain from which does not exceed 35 mW.

621.396.615.11 680

A Low-Frequency Function Generator.—R. H. Brunner. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 114–117.) Description of a relaxation oscillator capable of giving a constant-amplitude output of sine, square or triangular waveform, in the frequency range 0·01 c/s–1 kc/s.

621.396.615.12.078 681

Automatic Tuning Control of H.F. Generators with Varying Load.—H. Hertwig. (*Electronic Applic. Bull.*, Jan. 1952, Vol. 13, No. 1, pp. 9–18.) The phase relation between the primary voltage and the secondary current of the output transformer of a crystal-driven 40-Mc/s 500-W generator is used for automatic control of the tuning of the output circuit under varying load conditions. Control is effected via a bridge type of phase-measurement unit using a pair of Type-EQ80 enneodes, which acts on a control unit using a pair of Type-PL21 thyratrons. Circuit details, photographs and performance figures are given.

621.396.615.17 : 621.314.7 682

Pulse Duration and Repetition Rate of a Transistor Multivibrator.—G. E. McDuffie, Jr. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1487–1489.) Expressions are derived for the pulse duration and repetition rate of an astable (oscillatory) multivibrator circuit using a point-contact transistor. The formulae are confirmed experimentally for repetition rates from 200 to 10 000 per sec and pulse durations from 30 to 900 μs.

621.396.619.13 : 621.318.4 683

Variable Inductance and its Application in Frequency-Sweep Oscillators.—W. Lange. (*Funk u. Ton*, Oct. 1952, Vol. 6, No. 10, pp. 534–540.) The modulation principle described consists in varying the voltage across an inductor L_1 while the current through it is maintained constant. Alternations of the control-grid voltage of a pentode cause anode and screen-grid currents to vary in magnitude, developing a voltage variation across an anode inductor L_2, coupled to the inductor L_1 in the cathode circuit. The current through L_1 is maintained constant by an alternating voltage applied to the suppressor grid. A frequency shift of $\pm 20\%$ around 470 kc/s is attainable. The use of the method for f.m. and frequency-sweep tuning is indicated.

621.396.622.7 : 621.396.619.13 684

Double-Counter F.M. and A.F.C. Discriminator.—J. J. Hupert, A. Przedpelski & K. Ringer. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 124–125.) A frequency discriminator with large bandwidth and good stability comprises a pair of pulse-counter circuits, each preceded by mixer and pulse-forming circuits, with their outputs connected in series.

621.396.645 + 621.396.615 685

Vacuum-Tube Circuits without Plate Supplies.—P. B. Clark. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 192 . . 199.) The existence of thermal e.m.f. and contact potential makes circuit operation possible without application of voltage to valve anodes. A limiter, a low-gain amplifier, a multivibrator and an oscillator circuit are described.

621.396.645 686

Volume Compression and Expansion.—B. D. Corbett: G. J. Pope. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, p. 580.) Comment on 365 of February and author's reply.

621.396.645 : 621.313.2.026.441/.442]-9 687

Calculations for a Power Amplifier for a D.C. Motor.—J. Zakheim. (*Onde élect.*, Nov. 1952, Vol. 32, No. 308, pp. 455–458.) The design is considered of valve circuits for driving low-power motors (5–15 W), such as are often used in control or telemetry equipment.

621.396.645 : 621.314.7 688

Transistor Operation: Stabilization of Operating Points.—R. F. Shea. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1435–1437.) The provision of optimum emitter and collector bias currents for a transistor amplifier is discussed, and a relation between the values of resistors and voltages is derived which should be satisfied if stable operation is to be achieved.

621.396.645 : 621.314.7 689

Transistor Amplifier Cut-Off Frequency.—D. E. Thomas. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1481–1483.) The effect of the positive feedback associated with the internal base resistance of a transistor on its cut-off frequency is analysed. Expansion of the expression for current gain α in a Taylor series shows that only the phase shift in α is important in reducing the cut-off frequency.

621.396.645 : 621.314.7 690

Frequency Variations of Current-Amplification Factor for Junction Transistors.—R. L. Pritchard. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1476–1481.) In a grounded-emitter or grounded-collector connection of a junction-type transistor, the effective current amplification is proportional to $1/(1 - \alpha)$, where α is the current gain of the transistor. As a result of the phase shift associated with α, the value of $1/(1 - \alpha)$ decreases rapidly with increasing frequency, so that the upper frequency limit of the grounded-emitter or grounded-collector arrangements may be considerably lower than has been expected. Results of measurements of the frequency variation of α for several fused-impurity *p-n-p*-junction transistors developed by Saby (877 below) are shown graphically.

621.396.645 : 621.387.4 691

The Reproduction of Voltage Pulses by means of a Proportional Amplifier.—U. Cappeller. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 330–343.) Analysis of an amplifier for nuclear-physics investigations is based on Laplace transformations. A characteristic transmission function is introduced which completely describes the transmission properties of the amplifier and represents the combined pulse-distorting influences of the individual circuit elements. The transmission functions of single-stage and multistage amplifiers are derived and a detailed exposition is given of the distortion experienced by a typical exponential-decay pulse. Optimum reproduction of such a pulse requires a particular relation between pulse duration and the time constants of the amplifier circuits. Improvement of time-resolution involves closer restriction of this relation and greater complexity of circuits. Use of negative feedback offers advantages; a two-stage feedback amplifier is a useful unit in a large amplifier system. The transmission function can alternatively be determined from the response of the amplifier to steady alternating voltage; a suitable method of measurement is described.

621.396.645.371.029.45 692

A Photocell Amplifier for Infra-Red Spectroscopy.—D. A. H. Brown. (*J. sci. Instrum.*, Sept. 1952, Vol. 29, No. 9, pp. 292–294.) Description of a highly linear amplifier for use at a chopping frequency of 800 c/s. Gain variation is $< \frac{1}{2}\%$ for 10% variation of supply voltage.

621.396.822 693

Symposium on Noise. General Introduction.—H. B. G. Casimir. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 199–206.) Discussion of the mathematical representation of noise.

GENERAL PHYSICS

534.01 + 538.56] : 621.319.55 694

A Coherent Theory of Relaxation Phenomena.—E. Hiedemann & R. D. Spence. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 109–123.) The theory presented is based on the theory of functions. General formulae are derived and the particular case of elastic relaxation oscillations is discussed. Conditions for the occurrence of discrete and of continuous relaxation spectra are determined and a formula is derived for the distribution function of the continuous spectrum.

534.26 + 535.43 695

On a Multiple Scattering Theory of the Finite Grating and the Wood Anomalies.—V. Twersky. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1099–1118.) The problem is treated by applying the solution previously obtained (2685 of 1952) for multiple scattering from an arbitrary configuration of parallel cylinders. Both the transmission grating and the reflection grating comprising semi-cylindrical bosses on a perfectly conducting plane are considered. The complete expression for the scattered wave is given for the case of grating width small compared with distance to observation point and cylinder spacing large compared with λ. The case of radii $\ll \lambda$ is investigated in detail; bright and dark bands in the intensity curves, similar to the grating anomalies found by Wood in 1902, are related to the magnitudes and phases of the various orders of scattering. The theory is extended to gratings with elements other than cylinders.

535.42.001.11 696

Removal of an Inconsistency in the Theory of Diffraction.—D. S. Jones. (*Proc. Camb. phil. Soc.*, Oct. 1952, Vol. 48, Part 4, pp. 733–741.) In certain cases the integral equations involved in analysis of the diffraction of small-amplitude acoustic or e.m. waves give a solution which does not satisfy the boundary conditions imposed, but which agrees with the solution found by other means. This inconsistency is removed if different boundary conditions are imposed in the formulation of the problem; this is illustrated by discussion of two-dimensional diffraction of a plane e.m. wave by a perfectly conducting semi-infinite plane.

537.12 697

The Rydberg Constant and the Atomic Mass of the Electron.—E. R. Cohen. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, pp. 353–360.) Re-evaluation of the spectroscopic data pertinent to the Rydberg constant and the mass of the electron yields the values $R_\infty = 109737{\cdot}326 \pm 0{\cdot}014$ cm^{-1} and $m = (54{\cdot}895 \pm 0{\cdot}008) \times 10^{-5}$ atomic mass units. From microwave measurements $m = (54{\cdot}8785 \pm 0{\cdot}0019) \times 10^{-5}$ a.m.u. and $R_\infty = 109737{\cdot}311 \pm 0{\cdot}012$ cm^{-1}. Houston's and Chu's data can be brought into agreement with those of Drinkwater, Richardson & Williams by assuming differences in the wavelength standards used.

537.226.2/.3 : 541.135 698

Dielectric Dispersion in Pure Polar Liquids at Very High Radio Frequencies: Part 3 — The Effect of Electrolytes in Solution.—J. A. Lane & J. A. Saxton. (*Proc. roy. Soc. A*, 9th Oct. 1952, Vol. 214, No. 1119, pp. 531–545.) A description is given of measurements, at mm and cm wavelengths, of the absorption of electrolyte solutions, of concentrations up to 3N, water and methyl alcohol being used as solvents. The results obtained are analysed in terms of Debye's theory of dispersion in a polar dielectric, and modifications of the theory necessary to take account of the ionic conductivity of an electrolyte are indicated. Measurements on aqueous solutions of

NaCl are discussed in relation to Hückel's theory of electrolytic solutions. Part 2: 3400 of 1952.

537.228.1 : 548.0 699
Piezoelectricity, Ferroelectricity, and Crystal Structure.—A. von Hippel. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 158–173. In English.) By visualizing polar crystals as a network of permanent dipole moments, the piezoelectric and ferroelectric properties of dielectrics may be deduced from considerations of molecular symmetry. This approach is used to clarify the relation between the sphalerite (cubic) and wurtzite (hexagonal) structures, the ferroelectric feedback effect in $BaTiO_3$, aspects of domain formation, and the interrelation between ferroelectricity and piezoelectricity.

537.311.1 : 538.632 700
Carriers of Electricity in Metals exhibiting Positive Hall Effects.—S. Brown & S. J. Barnett. (*Phys. Rev.*, 15th Aug. 1952, Vol. 87, No. 4, pp. 601–607.) Measurements on samples of Mo and Zn, made by an inertia method, showed the sign of the charge/mass ratio of the carriers to be negative, the mean value of over 100 measurements being within 3% of the value for free electrons.

537.523/.527 701
Electrical Discharges in Gases.—F. L. Jones. (*Nature, Lond.*, 11th Oct. 1952, Vol. 170, No. 4328, pp. 601–603.) A summarized account of six papers, with discussion, presented at a meeting of Section A (Mathematics and Physics) of the British Association, Belfast, September 1952.

537.523/.525] : 546.292 : 538.56.029.63 702
High-Frequency Breakdown in Neon.—A. D. MacDonald. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, p. 420.) Preliminary report of the results of measurements of breakdown fields for cylindrical cavities of heights 0·317 and 0·634 cm, over the pressure range 0·5–300 mm Hg. A detailed report is to be published in the *Canadian Journal of Physics*.

537.525 703
An Explanation of the Extremely Low Normal Running Potential of a High-Frequency Discharge between Plane Plates.—F. Schneider. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 324–325.) A tentative explanation is provided by taking account, in the equation of electron motion, of a restoring force indicated by plasma diffusion theory.

537.533.75 704
Chromatic Losses of Electrons in Passage through Matter.—G. Möllenstedt. (*Optik, Stuttgart*, 1952, Vol. 9, No. 10, pp. 473–480.) An account of investigations of the velocity distribution of electrons after passage through various gases and solid materials.

537.533.8 705
The Theory of Secondary Emission.—J. F. Marshall. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, pp. 416–417.) Preliminary note; to be treated in detail in a forthcoming paper.

538.113/.114 706
Antiferromagnetism.—R. Ochsenfeld. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 350–360.) A comprehensive review of the subject, with 42 references.

538.3 707
A Simplified Form of the Relativistic Electromagnetic Equations.—N. W. Taylor. (*Aust. J. sci. Res., Ser. A*, Sept. 1952, Vol. 5, No. 3, pp. 423–429.) Maxwell's equations are expressed in the form of a single four-vector density equation, in which the field tensor has only three distinct complex components. The number of equations is reduced, but all the usual classical formulae may be obtained by separating the real and imaginary parts.

538.521 708
The Induction of Electric Currents in a Uniformly Conducting Circular Disk by the Sudden Creation of Magnetic Poles.—A. A. Ashour. (*Quart. J. Mech. appl. Math.*, Sept. 1952, Vol. 5, Part 3, pp. 379–384.) Using toroidal coordinates, a Riemann space of two regions is constructed, as in Sommerfeld's method of multiform potentials, and the Riemann potential for a single magnetic pole is determined. Jeans' treatment of uniform finite plane current sheets is applied to the case of the circular disk, numerical results being given for a particular example.

538.56 + 535.14 709
Hertzian Waves and Photons.—L. de Broglie. (*Onde élect.*, Oct. 1952, Vol. 32, No. 307, pp. 393–396.) Discussion of the implications, in physical theory, of an uncertainty relation between (*a*) the number of particles of the Bose-Einstein type associated with a wave and (*b*) the phase of that wave.

538.566 : 535.42 710
Diffraction of Electromagnetic Waves by Apertures in Plane Conducting Screens.—J. P. Vasseur. (*Ann. Phys., Paris*, July/Aug. 1952, Vol. 7, pp. 506–563.) See 2184 of 1952.

538.566 : 535.42 711
Diffraction by an Edge and by a Corner.—D. S. Jones. (*Quart. J. Mech. appl. Math.*, Sept. 1952, Vol. 5, Part 3, pp. 363–378.) Conditions are given which are sufficient to ensure that the current density normal to an edge is zero at the edge and that there is no line distribution of charge on the edge. An extra condition is given which makes the components of the field parallel to the edge finite. The solution is then shown to be unique. Simpler conditions are given for two-dimensional fields. The agreement of various known solutions with the conditions here determined is discussed. Certain simple types of current and charge distribution lead to a unique solution for the diffraction by corners formed by flat surfaces.

538.566 : 537.562 712
Propagation of Electric Waves in Stratified and Continuously Variable Plasmas.—W. O. Schumann. (*Z. Naturf.*, Nov. 1950, Vol. 5a, No. 11, pp. 612–617.) For a plasma of sandwich structure, with the denser medium inside, there are two possible frequency ranges for propagation in which the phase velocity falls from the value c at the lower limiting frequency to zero at the upper limiting frequency. When the density varies linearly towards the interior to an arbitrarily high value, the wave is strongly concentrated at the plane for which $\epsilon = 0$ and the phase velocity tends to zero. For a plasma of uniform density and natural frequency ω_0, with outer layers in which density falls off linearly, propagation is possible only when $\omega^2 < \omega_0^2/2$. The more the wave is concentrated at the region $\epsilon = 0$, and the smaller its phase velocity, the nearer ω^2 approaches to $\omega_0^2/2$.

538.632 713
Hall Effect.—O. Lindberg. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1414–1419.) Discussion of the Hall, Ettingshausen, Nernst, and Righi-Leduc effects.

539.234 : 537.311.1 714
Mean Free Paths of Electrons in Evaporated Metal Films.—F. W. Reynolds & G. R. Stilwell. (*Phys. Rev.*,

15th Oct. 1952, Vol. 88, No. 2, pp. 418–419.) Estimates of the mean free paths of the conduction electrons in Cu and Ag films, based on resistivity measurements on films of thickness from 100 to 1 500 μ, are found to be in good agreement with values calculated from Dingle's theory (2189 of 1950).

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.746/.75 **715**

Sunspot Areas, Flares and Filaments observed at the Stockholm Observatory in the Years 1950 and 1951.—Y. Öhman & L. O. Lodén. (*Tellus*, Aug. 1952, Vol. 4, No. 3, pp. 241–248.)

523.755 : 523.78 **716**

Comparison of Photographs of the Corona obtained at the Eclipse of 1952, February 25, with Simultaneous Observations by Lyot Coronagraphs.—H. von Klüber. (*Observatory*, Oct. 1952, Vol. 72, No. 870, pp. 207–209.) Photographs of the solar corona taken at Khartoum are reproduced and discussed in relation to routine estimates of the intensity of the green corona line λ5303. Possible correlation of particular features of the corona with geomagnetic effects is noted.

523.85 : 621.396.822 **717**

The Positions of Six Discrete Sources of Cosmic Radio Radiation.—B. Y. Mills. (*Aust. J. sci. Res., Ser. A*, Sept. 1952, Vol. 5, No. 3, pp. 456–463.)

523.854 : 621.396.822.029.63 **718**

A Preliminary Survey of 1420-Mc/s Line Emission from Galactic Hydrogen.—W. N. Christiansen & J. V. Hindman. (*Aust. J. sci. Res., Ser. A*, Sept. 1952, Vol. 5, No. 3, pp. 437–455.) Report of measurements on a radiation source having the form of a band of varying intensity along the galactic equator.

551.510.52 **719**

The Ionic Equilibrium of the Lower Atmosphere and Recombination Theories.—B. Vitale. (*Ann. Geofis.*, April 1952, Vol. 5, No. 2, pp. 257–271.)

551.510.535 **720**

A Note on Ionospheric Wind Measurements at 150 kc/s.—G. H. Millman. (*Ann. Géophys.*, Oct./Dec. 1951, Vol. 7, No. 4, pp. 272–274. In English.) Report of measurements made in Pennsylvania, using a three-receiver technique similar to that of Mitra (96 of 1950).

551.510.535 **721**

The Ionization of the E Layer: its Measurement and Relation to Solar Eruptions.—K. Bibl. (*Ann. Géophys.*, Oct./Dec. 1951, Vol. 7, No. 4, pp. 208–214.) Criteria for distinguishing between normal and abnormal layers are discussed. Relatively thin layers are classed as abnormal; their ionization distribution corresponds to a power law with index > 2. A definition of f_0E is given which remains valid for complex layers; f_0E is the highest critical frequency of a normal layer preceding or coinciding with the first discontinuity between the E and F echoes. Application of this definition to the evaluation of ionosphere observations made at Freiburg during 1950 and 1951 leads to greater constancy of the daily variations and monthly means of f_0E. Examination of the values of f_0E for three summer months indicates that all deviations $> 0{\cdot}2$ Mc/s above the monthly mean are attributable to sudden ionospheric disturbances.

551.510.535 : 523.854 : 621.396.822 **722**

The Diffraction of Galactic Radio Waves as a Method of Investigating the Irregular Structure of the Ionosphere.—A. Hewish. (*Proc. roy. Soc. A*, 9th Oct. 1952, Vol. 214, No. 1119, pp. 494–514.) An account of an investigation of ionospheric characteristics by observation of changes in the diffraction pattern of radio waves from the galaxy. Ionospheric irregularities cause irregular changes of phase in the galactic waves passing through them. Observations at two stations about 1 km apart indicate that such ionospheric irregularities may have a lateral extent of 2–10 km and a variation of electron density of about 5×10^9 e/cm³; their height is about 400 km, they are most pronounced about midnight, and they show little annual variation. Such portions of the ionosphere have a wind-like motion with velocities of the order of 100–300 m/s. The velocity decreases after midnight, and large velocities are associated with periods of geomagnetic disturbances.

551.510.535 : 546.21-1 : 537.56 **723**

Production of the E Layer in the Oxygen-Dissociation Region in the Upper Atmosphere.—D. C. Choudhury. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, pp. 405–408.) The probable value of the absorption cross-section of O_2 for the ionization causing the E layer is calculated, using the O_2 height-distribution data obtained by Moses & Wu (129 of 1952). Discussion indicates that the pre-ionization by solar radiation in the range 900–1 000 Å suggested by Nicolet (420 of 1947) and ionization by high-energy photons emitted from the solar corona [Hoyle & Bates (388 of 1949)] are both operative in producing E-layer ionization. The former produces the normal E layer and the latter is responsible for the fine structure of the E layer recently reported by Naismith & Bramley (473 of 1952).

551.510.535 : 621.3.087.4 **724**

A New Ionosphere Sounder.—K. Bibl. (*Ann. Géophys.*, Oct./Dec. 1951, Vol. 7, No. 4, pp. 265–267.) Frequency-sweep apparatus covering the range 1–16 Mc/s is described. The aerial circuit is eliminated, the high-impedance rhombic aerial being directly matched to the power valves. The frequency range is covered in the other circuits in two wavebands each with a 1 : 4 ratio; quick waveband changing is achieved by means of carbon brushes rotating with the variable capacitors. The transmitter and receiver have a common oscillator. The power-supply arrangements are particularly described; the unit is of small size.

551.510.535 : 621.3.087.4 **725**

Improvements to the Berkner-Wells-Seaton Type of Ionosphere Sounder.—E. Harnischmacher. (*Ann. Géophys.*, Oct./Dec. 1951, Vol. 7, No. 4, pp. 262–264.)

551.510.535 : 621.317.083.7 **726**

High-Altitude Research.—E. Burgess. (*Engineer, Lond.*, 12th & 19th Sept. 1952, Vol. 194, Nos. 5042 & 5043, pp. 338–340 & 370–373.) An account of methods and equipment used with rockets for investigations at heights up to 250 miles; physical phenomena observed include the ion and electron densities in the ionosphere. A description is given of a 23-channel telemetering system transmitting at 1 kMc/s with a peak power of 1 kW.

551.510.535 : 621.396.11.029.55 **727**

Ionospheric Measurements at Oblique Incidence over Eastern Australia.—Strohfeldt, McNicol & Gipps. (See 807.)

551.594.5 : 621.396.9 **728**

Localization of Aurorae with 10-m High-Power Radar Technique, using a Rotating Antenna.—G. Hellgren & J. Meos. (*Tellus*, Aug. 1952, Vol. 4, No. 3, pp. 249–261.) Radar equipment with a peak pulse power of 100 kW, pulse width 40 μs, and 3-element Yagi aerial rotating at

2 r.p.m., has been used since May 1951 at Kiruna Geophysical Observatory for the location of aurorae. Preliminary results of the observations indicate good correlation between auroral activity, geomagnetic activity, and the appearance of the N_1 layer, a special type of sporadic-E ionization often appearing in connection with magnetic bays. The distribution in range and bearing of the recorded aurorae agrees with the simple theory that most of the radio-wave scattering comes from those points where the radar beam is perpendicular to the surface of the auroral discharges. The calculated height distributions of the reflection centres have maxima around 120 km.

551.594.6 : 538.566.029.45/.51 **729**

Propagation of Very Long Electromagnetic Waves, and the Wave Spectrum of Lightning.—Schumann. (See 802.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 **730**

Commercial Radar System.—W. F. Johnson. (*Elect. Mfg*, Feb. 1951, Vol. 47, No. 2, pp. 72–77.) General description of navigational radar equipment for merchant ships. Both 3-cm and 10-cm wavelengths are catered for and a standard 16-in. television c.r. tube provides a satisfactory display. Accessibility of the various units to facilitate maintenance work is a special feature of the design. The reflectors for the aerial systems consist of expanded metal attached to a suitable framework.

621.396.9 : 621.396.677.088.22 **731**

Scanning Aberrations of Radio Lenses.—T. C. Cheston & D. H. Shinn. (*Marconi Rev.*, 4th Quarter 1952, Vol. 15, No. 107, pp. 174–184.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.472.21 **732**

Photoelectric Measurements on ZnS Single Crystals.—J. Krumbiegel. (*Naturwissenschaften*, Oct. 1952, Vol. 39, No. 19, p. 447.) Measurements on a number of specimens, using monochromatic irradiation of wavelength ranging from 450 to 2 000 mμ in steps of 50 mμ, indicated two peaks of photoconductivity, at about 750 and 1 150 mμ respectively. Time lag of the photocurrent was observed in all cases.

535.215 : 546.883 **733**

Photoconduction in Anodic Ta_2O_5.—L. Apker & E. A. Taft. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, pp. 58–59.) Observations show that fundamental optical absorption occurs for $h\nu > 4{\cdot}6$ eV. The resultant photoconduction between electrolyte and Ta substrate shows a quantum yield near 0·5 when the field in the film is of the order of 10^7 V/cm.

535.376 **734**

Light Emission and Destruction of Phosphors due to Electron and Ion Impact.—W. Hanle & K. H. Rau. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 297–308.) Measurements were made of the light emission from ZnS-Ag, Zn_2SiO_4-Mn and $MgWO_4$ phosphors under excitation by H_2, He, Ne, Ar and Xe canal rays. The emission depends only to a very slight extent on the energy of the ions and is independent of the ion density. With increase of ion mass the emission decreases, with a jump in passing from He to Ne ions. Destruction of the three phosphors by canal-ray bombardment was found to increase with ion mass and to follow a simple law; it was about 10 times more rapid for ZnS than for the other two phosphors. Organic phosphors were found to undergo deterioration when bombarded by electrons, the effect being correspondingly smaller than in the case of ion bombardment.

537.228.1 **735**

Voltage Measurements on Rochelle Salt under Compressive Stress.—B. Püschel. (*Arch. tech. Messen*, Sept. 1952, No. 200, pp. 197–198.) A 50-g weight was dropped from different heights on to a Rochelle-salt crystal contained between brass plates 1 mm thick. Voltages up to 600 V were developed.

537.311.33 **736**

Electronic States in Crystals under Large Over-All Perturbations.—P. Feuer. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, pp. 92–100.) Solutions of the three-dimensional Schrödinger equation are discussed for a potential which is the sum of a potential with the periodicity of the crystal lattice plus a disturbing potential. General theory developed for large perturbations is applied to a one-dimensional crystal in a uniform electric field, using the narrow-band approximation; the probability for an electron to cross a forbidden energy band is calculated. The results are considered in relation to the observations of McAfee et al. (164 of 1952) on Ge *p-n* junction.

537.311.33 **737**

On the Electrical Properties of Porous Semiconductors.—E. B. Hensley. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1122–1129.) Calculations are made of the component of conductivity due to the presence of an electron gas in the pores of a semiconductor, such as the oxide coating of a cathode, on the basis of simplified pore models. The conditions are indicated for which this component can become an appreciable part of the total conductivity. The thermoelectric power is also investigated.

537.311.33 : [546.28 + 546.289 **738**

Properties of Silicon and Germanium.—E. M. Conwell. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1327–1337.) The latest available information on the fundamental properties of Ge and Si is presented in tables and curves. Electrical properties, especially carrier density and mobility, are treated in greatest detail.

537.311.33 : 546.289 **739**

***P-N* Junctions by Impurity Introduction through an Intermediate Metal Layer.**—L. D. Armstrong. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1341–1342.) Indium is melted on to a gold-plated area on the surface of a Ge *n*-type crystal and diffuses into the Ge to form a *p-n* junction with a well-defined area.

537.311.33 : 546.289 **740**

Lifetime of Injected Carriers in Germanium.—D. Navon, R. Bray & H. Y. Fan. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1342–1347.) Carrier lifetime was determined from measurements of the variation of the conductance of the test piece after production of excess carriers by applying a voltage pulse. The effect of heat treatment was also investigated.

537.311.33 : 546.289 **741**

Measurement of Minority-Carrier Lifetime in Germanium.—L. B. Valdes. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1420–1423.) Equipment is described in which the carriers are liberated optically on a flat face of a crystal, their conçentration then being measured as a function of distance from the point of liberation. Lifetimes from a few to several hundred microseconds can be determined.

537.311.33 : 546.289 742

Activation Energy of Heat-Treatment-Introduced Lattice Defects in Germanium.—R. M. Baum & C. S. Hung. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, pp. 134–135.) Hall-coefficient and resistivity measurements are reported on two single-crystal samples converted from *n*-type to *p*-type by heat treatment. The values deduced for the thermally produced acceptors, viz. 0·058 and 0·047 eV respectively, are considerably higher than for chemically produced acceptors; this result is in general agreement with that of Dunlap (*Phys. Rev.*, 1st July 1952, Vol. 87, No. 1, p. 190).

537.311.33 : 546.289 743

***p-n* Junctions produced by Growth-Rate Variation.**—R. N. Hall. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, p. 139.) More than 100 uniformly spaced *p-n* junctions have been formed in an ingot of Ge by periodically varying the rate of growth of the crystal from the melt. The process depends on the presence in the melt of two impurities of opposite type, e.g. Sb (*n* type) and Ga or In (*p* type), whose segregation constants vary at different rates with rate of growth.

537.311.33 : 546.289 : 548.55 744

Preparation of Germanium Single Crystals.—L. Roth & W. E. Taylor. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1338–1341.) Description of two methods, using a vacuum furnace with h.f. induction heating. In the first method the crystal is grown in the crucible by controlling the rate of cooling; in the second, a seed crystal is gradually withdrawn from the surface of the melt.

537.311.33 : 621.314.632 745

Cadmium-Sulfide Crystal Rectifiers.—(*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 189, 192.) Long hexagonal *n*-type CdS crystals were tested for rectifier properties; the influence of orientation, mounting, and impurities was investigated. Point-contact rectifiers were obtained, optimum rectification occurring at 7 V. Transistor action was not obtained, but is thought to be possible.

537.311.33 : 621.314.7 746

Transistor Electronics: Imperfections, Unipolar and Analog Transistors.—W. Shockley. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1289–1313.) A detailed discussion of the physics of transistors, particularly as regards the properties of holes and electrons and the effects of impurities, together with a description of transistor action and of a new class of unipolar transistors, some of which, from their similarity to valves, are termed analogue transistors.

537.311.33 : 621.396.822 747

Current Noise in Semiconductors. A Re-examination of Bernamont's Data.—D. A. Bell. (*Phil. Mag.*, Oct. 1952, Vol. 43, No. 345, pp. 1107–1111.) The results obtained by Bernamont (1715 of 1937) on thin metal films, here regarded as semiconductors, indicated that current noise varied approximately inversely with frequency over the range 96 c/s–162 kc/s. On adjusting his results to make allowance for Johnson noise and for the supposed loss of a factor of 10 in the calculations, the noise from his resistor 'B' closely follows the inverse-frequency law for all current densities.

537.311.33 : 621.396.822 748

Current Noise in Semiconductors.—D. A. Bell. (*Wireless Engr*, Jan. 1953, Vol. 30, No. 1, pp. 23–24.) Discussion of the law relating the frequency spectrum of the noise voltage with the current in a semiconductor.

538.114 749

Magnetic Domains.—H. J. Williams. (*Bell Lab. Rec.*, Oct. 1952, Vol. 30, No. 10, pp. 385–396.) The domain structure of magnetic materials is shown by numerous microphotographs and diagrams, and electron-spin changes and domain-boundary movement under the action of a magnetic field are illustrated by diagrams and models.

538.221 750

Applications and Properties of Ferroxcube.—(*Electronic Applic. Bull.*, March/April 1952, Vol. 13, Nos. 3/4, pp. 44–58.) Tables and numerous curves are given which show the properties and operating characteristics of the various available types and grades of ferroxcube, so that the material most suitable for a particular application can easily be selected.

538.221 751

Ferromagnetic Properties of Hexagonal Iron-Oxide Compounds with and without a Preferred Orientation.—G. W. Rathenau, J. Smit & A. L. Stuyts. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 250–260. In English.) An account of ferroxdure, a ceramic material with the composition $BaO.6Fe_2O_3$, has previously been given [2824 of 1952 (Went et al.)]. Materials of about this composition are here discussed. The variation of the magnetization with the strength of the applied field is shown graphically. The formation of Bloch walls results in irregularities in the crystal structure. With regular crystal distribution and crystals above a certain critical size, demagnetization by Bloch-wall displacement can occur for positive field strengths of about $4\pi I_s$, where I_s is the saturation magnetization. When crystal-orientation processes are applied to ferroxdure materials, BH_{max} values of 3×10^6 gauss.oersted can be reached. An improvement of the texture of these ceramics is obtained by increase of grain size.

538.221 752

A New Method of Melting Ferromagnetic Semiconductors. $BaFe_{18}O_{27}$, a New Kind of Ferromagnetic Crystal with High Crystal Anisotropy.—H. P. J. Wijn. (*Nature, Lond.*, 25th Oct. 1952, Vol. 170, No. 4330, pp. 707–708.) An account is given of the h.f. method used in melting a mixture of Fe_3O_4 and $BaCO_3$ in an atmosphere of N_2; an alumina crucible was used, with a lid of Rh or Ir to give additional surface heating. From one melt a crystal with the composition $BaFe_{18}O_{27}$ was obtained. This has a structure resembling that of $BaFe_{12}O_{19}$ [2824 of 1952 (Went et al.)] and exhibits very high crystal anisotropy.

538.221 : 548.0 753

Coercive Force and Crystal Energy.—W. Gerlach. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 286–290.) Sintered Ni rods show an increase of coercive force with decreasing temperature proportional to the square root of the crystal energy. Wires of solid Ni annealed for a long time at a moderate temperature show a similar effect, but sintered Fe and Co rods exhibit anomalous effects.

538.221 : 548.1.023 754

Crystal Structure of $BaFe_{18}O_{27}$.—P. B. Braun. (*Nature, Lond.*, 25th Oct. 1952, Vol. 170, No. 4330, p. 708.) Results are tabulated of an X-ray investigation of the crystal mentioned by Wijn (752 above).

538.221 : 621.392.26 755

Magnetic Double Refraction at Microwave Frequencies.—M. T. Weiss & A. G. Fox. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, pp. 146–147.) Magnetic double refraction has been observed at 24 and 9 kMc/s, evidenced by a conversion from linear to elliptical polarization of a dominant *TE* wave in a circular waveguide filled with ferrite and subjected to a d.c. magnetic field transverse to the direction of propagation and at 45° to the initial direction of polarization. The double refraction is attributed to the difference between the r.f. permeability parallel to and transverse to the field.

538.23 **756**

A Relation between Hysteresis Coefficient and Permeability.—M. Kornetzki. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 343–345.) Assuming for simplicity that the magnetization curves of various materials, measured in closed magnetic circuits, are distinguished from one another only by different scale factors, it is shown that the relative hysteresis coefficient is proportional to the initial permeability and inversely proportional to the saturation induction. Measurements on Ni-Zn ferrites with initial permeabilities of 30–1 500 indicate hysteresis coefficients differing among themselves by a factor of 50, though the ratio of hysteresis coefficient to initial permeability varies only by a factor of 3.

538.242 **757**

A New Gyromagnetic Effect in Permalloy and Iron.—S. J. Barnett & L. A. Giambomi. (*Phys. Rev.*, 1st Oct. 1952, Vol. 88, No. 1, pp. 28–37.) Experiments were made with long cylinders of compressed powder material magnetized axially nearly to saturation and subjected to a weak transverse alternating magnetic field of period much greater than the relaxation time. A transverse magnetization was observed perpendicular to the applied alternating field. Measured values of the effect are in substantial agreement with results of earlier experiments.

539.234 **758**

The Contamination in Evaporated Films by the Material of the Source.—O. S. Heavens. (*Proc. phys. Soc.*, 1st Oct. 1952, Vol. 65, No. 394B, pp. 788–793.) A microchemical and a radioactive-tracer method were used to investigate contamination of Ag and Ge films by the boats or filaments used in the vaporization process. One or two parts of W or Mo in 10^7 can be detected by the tracer method. Minimum contamination attainable was of the order of a few parts in 10^6.

546.431.824-31 : 539.11 **759**

Domain Properties in $BaTiO_3$.—W. J. Merz. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, pp. 421–422.) Short account of investigations of the arrangements and movements of the ferroelectric domains in single crystals of $BaTiO_3$ under the influence of an electric field. Only so-called 90° walls have hitherto been reported, but the smaller 180° walls constituting the true domain boundaries have now been observed. The average width of these antiparallel domains, as measured by means of a microscope, is from $0{\cdot}1\,\mu$ to about $10\,\mu$. Observations were made of domain movements corresponding to each step of a hysteresis loop. More detailed studies of the number of domains, their velocities and relaxation times, are being made by applying very short rectangular electrical pulses.

621.3.042.143 : 538.221 **760**

Ultrathin Magnetic Alloy Tapes with Rectangular Hysteresis Loops.—M. F. Littmann. (*Elect. Engng, N.Y.*, Sept. 1952, Vol. 71, No. 9, pp. 792–795.) Text of paper presented at A.I.E.E. General Meeting, June 1952. Report of an experimental study of high-permeability and high-resistivity tapes, suitable for cores for h.f. applications. Thicknesses range from ⅛ to 1 mil.

621.314.2.042.143 **761**

Core Materials for Small Transformers.—C. C. Horstman. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 40–42, 90.) Discussion of the reduction in transformer weight and electrical losses effected by the use of hypersil cores produced from strip material of thickness 1–5 mil. A recent development is the forming of longitudinal grooves in the thin strip prior to winding into a core. This results in increased rigidity and higher stability under temperature cycles. Improvements in Ni-alloy materials are also noted, such as the production of strip materials of thickness only ⅛ mil, with retention of high permeability and low coercive force.

621.318.1.042.15 **762**

The Production and Application of Magnetic Powders.—G. R. Polgreen. (*G.E.C. J.*, July 1952, Vol. 19, No. 3, pp. 152–169.) A description is given of modern methods and equipment for the manufacture of Fe and Cu-Ni-Fe-alloy (Gecalloy 111) powders; the chemical methods used result in reduced losses and improved stability. The properties of these powders and of the resulting cores are compared with corresponding properties of commercially available sheet materials and ferrites.

MATHEMATICS

517.511 **763**

The Decomposition of Functions.—J. W. Head. (*Proc. Camb. phil. Soc.*, Oct. 1952, Vol. 48, Part 4, pp. 742–743.) Discussion of functions of the step-voltage type. If $f(t)$ approaches asymptotically the value $t^n e^{-t}$ when the time scale is suitably chosen, $f(t)$ can be decomposed into a series of Laguerre functions which are mutually orthogonal over the range 0–∞. The coefficients in this series are here obtained in terms of the various Laguerre functions and of $(f)t$ and $dL_n(t)/dt$ when the Laguerre function $L_n(t)$ has a zero. An explicit formula is derived involving functions which can be calculated and tabulated.

517.942.82 **764**

General Rules for Laplace Transformation.—U. Kirschner. (*Funk u. Ton*, Oct. 1952, Vol. 6, No. 10, pp. 541–547.) A statement of general relations applying in different mathematical operations.

681.142 **765**

MONECA — A New Network Calculator for Motor Performance Calculations.—C. G. Veinott. (*Elect. Engng, N.Y.*, Sept. 1952, Vol. 71, No. 9, pp. 795–801.) Text of paper presented at A.I.E.E. General Meeting, January 1952. Description of an analogue computer for calculations on single-phase induction motors.

681.142 : 621.314.7 **766**

An Optical Position Encoder and Digit Register.—H. G. Follingstad, J. N. Shive & R. E. Yaeger. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1573–1583.) Transistor circuits are used in a small unit which performs the operations of 6-digit photoelectric encoding, pulse regeneration, digit storage, reflected-to-natural binary translation, and digit shifting.

681.142 : 621.314.7 **767**

A Transistor Shift Register and Serial Adder.—J. R. Harris. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1597–1602.) Equipment is described which can store a pair of binary numbers, add them, and produce the sum a digit at a time. The equipment is constructed from primary transistor units including a flip-flop circuit, pulse amplifiers with and without delay, and diode gate circuits.

681.142 : 621.395.625.3 **768**

On Two Problems in Potential Theory and their Application to the Design of Magnetic Recording Heads for Digital Computers.—A. D. Booth. (*Brit. J. appl. Phys.*, Oct. 1952, Vol. 3, No. 10, pp. 307–308.) Both theory and experiment indicate that the field outside the gap depends very little on the particular shape of pole-piece.

681.142 : 621.396.645 : 621.314.7 **769**

Regenerative Amplifier for Digital Computer Applications.—J. H. Felker. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1584–1596.) An amplifier using a point-contact transistor is used to regenerate digital information at a rate of 10^6/sec and to develop pulses with rise times $< 0{\cdot}05\ \mu$s.

MEASUREMENTS AND TEST GEAR

531.76 **770**

A Combined Timer and Cycle Counter.—P. Huggins. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 578–579.) Description, with full circuit details, of a unit which combines the functions of timer and cycle counter (2841 of 1952) within the range 0·02–2·64 sec. Two 12-position dekatron valves are used which provide visual indication of the time, in steps of $\frac{1}{50}$ sec, as the apparatus operates. Conversion from timer to cycle counter is effected by a simple switch.

531.76 : 621.318.57 : 621.317.755 **771**

A Dekatron C.R.O. Time Marker.—J. H. L. McAuslan. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 567–569.) Description, with detailed circuit diagram, of equipment using Type-GC10B dekatrons to provide longer time-markers on the screen of a c.r.o. at each tenth pulse, with additional negative markers for each hundredth pulse.

621.316.8(083.74) **772**

Gold-Chromium Standard Resistors.—A. Schulze & H. Eicke. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 321–324.) Experiments during the last 15 years have shown standard resistors made of Au-Cr alloy to be superior to those made of manganin. New constructions are described in which the resistance coil is housed in an argon-filled glass envelope; these types are suitable for measurements of the greatest precision, and are designated principal standard resistors.

621.317.328.089.6 **773**

Calibration of Commercial Field-Strength Meters.—C. C. Cook. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 44–46..99.) An account of the service provided by the National Bureau of Standards, with an outline of the methods of calibration of meters operating in the ranges 10 kc/s–30 Mc/s and 30–300 Mc/s.

621.317.336.029.64 **774**

The Determination of Impedance with a Double-Slug Transformer.—R. C. Ellenwood & E. H. Hurlburt. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1690–1693.) Formulae are derived by means of which the required impedance can be determined from a knowledge of the length, spacing, position and 'effective' dielectric constant of the slugs. A method of determining the 'effective' dielectric constant experimentally is described. Results of accuracy comparable with those given by a precision slotted-line method have been obtained.

621.317.337 : 621.396.611.1 **775**

***Q*-Factor Measurement.**—A. G. Wray. (*Marconi Instrumentation*, Oct. 1952, Vol. 3, No. 7, pp. 118–123.) The *Q* factors of single reactive components and of complete oscillatory circuits are discussed. The principles of measurement methods are outlined, and resistive, inductive and capacitive methods of injecting the necessary small test e.m.f. are considered.

621.317.351 : 534.442.2 **776**

Audio-Frequency Spectrum Analysis.—W. Sagajllo: S. V. Soanes. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, p. 581.) Comment on 3171 of 1952 and author's reply.

621.317.352 : 621.395 **777**

Reference-Level Test Equipment with Direct Indication, and its Importance for the Improvement of Telephony.—K. Braun & H. Koschel. (*Fernmeldetech. Z.*, Oct. 1952, Vol. 5, No. 10, pp. 447–455.) Description of apparatus, including artificial mouth and artificial ear, suitable for tests on complete telephone circuits or on individual items of equipment.

621.317.7.029.6 : 621.396.615.141.2 **778**

Microwave Devices for Magnetron Production Testing.—M. Nowogrodzki. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 36–37..111.) Outline description of (*a*) a cavity-resonator type of wavemeter using a neon lamp for visual indication of resonance, (*b*) equipment for measurement of output power, using a water load, (*c*) a r.f. power monitor for use in life tests.

621.317.7.029.62/.63 : 621.392 **779**

Helical Measuring Line for Microwaves.—F. Tischer. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 345–350.) A line of considerably reduced length and of characteristic impedance 50 Ω comprises a helical conductor wound in a groove in an insulating sleeve on an axial conductor, the whole being enclosed in a coaxial tube with a slot, through which projects an inductive probe of a type previously described (3048 of 1951). When the parameters are correctly chosen, phase and amplitude fluctuations along the line are $< 1\%$ over the frequency range 250 Mc/s–2 kMc/s. The accuracy of measurements made with this line is equal to that with a medium-quality straight line.

621.317.725.088.22 : 621.385.2 **780**

Diode-Valve-Voltmeter Errors.—G. D. Morgan. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 575–577.) Discussion, based on experience with a voltmeter using a Type-VR78 (Mazda-D.1) diode, of the magnitude of the possible errors in diode voltmeters.

621.317.733 : 621.317.374 **781**

New Method for the Measurement of Dielectric Loss Angle.—L. Schnell. (*Elektrotechnika, Budapest*, Sept. 1952, Vol. 45, No. 9, pp. 264–268. Discussion, pp. 268–269.) A bridge arrangement is described in which the voltage across the meter diagonal is a nearly linear function of tan δ.

621.317.733.029.51/.63 : 621.317.335 **782**

A Measurement Bridge for 0·1-1 000 Mc/s.—H. Voigt. (*Arch. elekt. Übertragung*, Oct. 1952, Vol. 6, No. 10, pp. 414–418.) The four capacitors constituting the arms of the bridge are formed by a set of parallel disks within a cylindrical screen. A flat test sample of a material whose dielectric properties are to be determined is introduced between one end plate and the removable end of the screen. The other end plate has a micrometer adjustment for balancing the bridge. A rectifier and valve voltmeter serve as balance indicator. A cylindrical type of construction is also shown that is suitable for measurements on liquid dielectrics. Typical results for the variation of dielectric constant and loss angle of

PVC with the amount of softening agent and with frequency are shown in diagrams.

621.317.737.088.22 : 621.3.012.3 **783**

***Q*-Meter Correction Chart for *Q*-Voltmeter Loading.**—R. Lafferty. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, p. 43.) An abac is given for correcting errors due to the shunting effect of the input resistance of the valve voltmeter used in the measurements.

621.317.755 : 621.314.7.012 **784**

Oscilloscopic Display of Transistor Static Electrical Characteristics.—N. Golden & R. Nielsen. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1437–1439.) Description, with schematic circuit diagram, of c.r.o. equipment permitting rapid comparison with a standard.

621.317.755 : 621.396.6 **785**

Slow-Speed Circular Timebase.—(See 673.)

621.317.78.029.6 **786**

A Laboratory Power Meter.—E. F. Schelisch. (*Marconi Rev.*, 4th Quarter 1952, Vol. 15, No. 107, pp. 167–173.) Two cartridge-type Si-crystal diodes are connected as shunt elements $\lambda_0/4$ apart in a coaxial line, a matching transformer being fixed between them. Power measurement to within 0·1 db is achieved over a range of 40 db or more in the microwave band. Possible uses of the device as a f.m. discriminator, an a.m. demodulator, or a frequency multiplier, are illustrated and discussed.

621.396.645.35 **787**

A High-Sensitivity Direct-Voltage Amplifier with High Input Resistance.—W. Kroebel. (*Z. Phys.*, 15th Sept. 1952, Vol. 133, Nos. 1/2, pp. 30–40.) A new type of contact breaker is described which uses a flexural type of double quartz plate as its vibrator. A type of construction is used in which the contact gap is completely screened from the exciting voltage applied to the crystal. Application is made to the amplification of direct voltages from sources of very high internal resistance. Full circuit details of an amplifier are given with which, for a bandwidth of 1 c/s and input resistance of 100 MΩ, a power of about 5×10^{-20} W can be measured.

621.396.822.029.64 : 621.327.3 **788**

The Design of Microwave-Noise Generators.—P. M. Ratcliffe. (*Marconi Instrumentation*, Oct. 1952, Vol. 3, No. 7, pp. 124–127.) Mumford (929 of 1950) showed that an ordinary Hg-vapour discharge lamp mounted in a waveguide acts as a good noise source for the 6-cm band. Similar sources have now been produced for the S and X bands at 10 cm and 3·2 cm. The new tubes are about 9 in. long and $\frac{5}{8}$ in. in diameter, and are filled with a Hg-vapour and Ar mixture at a pressure of 30 mm Hg, the d.c. power consumption for reliable operation being about 10 W. A filament is fitted at each end, one being heated by a.c. to assist in starting the discharge. Matching of the source to the waveguide is effected for the X-band mounting by insertion of the tube in the E plane of the waveguide at an angle of 10° to the waveguide axis. For the S-band mounting an H-plane fitting is used with the tube at right angles to the narrow walls of the waveguide. The generators deliver a noise-power output about 15 db above zero level.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

53.087.55 : 771.4 **789**

Photographic-Exposure Timers providing Compensation for Supply-Voltage Variations.—R. J. Hercock & D. M. Neale. (*Proc. Instn elect. Engrs*, Part II, Oct. 1952, Vol. 99, No. 71, pp. 507–515.) Description of a circuit giving exposure times, at the nominal supply voltage, continuously adjustable from 1 sec to 1 min or more, with an intensity-time product constant to within ±5% for supply-voltage variations from +15% to −20%.

534.1.08 **790**

Vibration Measurements.—R. Winslade. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 553–557.) Description of a pickup unit, with spring-supported coil moving in a magnetic field, for measuring vibration amplitudes, velocities and accelerations.

621.315.3.001.41 **791**

Continuous Measurement of Cable Diameters.—E. C. R. Scarfe. (*Elect. Times*, 4th Sept. 1952, Vol. 122, No. 3174, pp. 399–401.) The extruded cable runs between a pivoted tungsten-carbide stylus and a rotating anvil. Movement of the stylus varies the air gap between a fixed quartz crystal and one of its electrodes. The frequency variations derived from the capacitance change modulate a 100-kc/s carrier. The direct voltage obtained via conventional discriminator and output circuits is applied to a meter, calibrated in thousandths of an inch, which shows deviations from a preset nominal diameter.

621.316.7 : 621.314.7 **792**

Control Applications of the Transistor.—E. F. W. Alexanderson. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1508–1511.) The possibility is examined of using transistors for control functions at present performed by magnetic amplifiers, amplidynes and thyratrons. A transistor controlled by auxiliary transistors can function in a manner similar to that of a phase-controlled rectifier, and has certain definite advantages.

621.317.083.7 : 551.510.535 **793**

High-Altitude Research.—Burgess. (See 726.)

621.384.6 : 621.317.083.7 **794**

A Telemetering System for a Large Electrostatic Accelerator.—C. W. Johnstone, J. F. Kalbach & H. J. Lang. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1664–1674.) Description of the 16-channel pulsed-light-beam system used for monitoring and controlling the ion source, focusing and belt charging in the 12-MeV e.s. accelerator nearing completion at Los Alamos.

621.385.833 **795**

Electron-Optical Properties of Electrostatic Lenses.—W. Lippert & W. Pohlit. (*Optik, Stuttgart*, 1952, Vol. 9, No. 10, pp. 456–462.) Results of experimental investigations are presented as families of curves from which the electron-optical properties of an e.s. lens of the symmetrical 3-electrode type can be determined from the parameters of the electrode system.

621.387.4 **796**

Self-Quenching Parallel-Plate Vapour-Filled Counters with Operating Voltages below the Static Breakdown Field Strength.—J. Christiansen. (*Z. angew. Phys.*, Sept. 1952, Vol. 4, No. 9, pp. 326–329.)

621.387.424 **797**

Geiger Counter Tubes.—N. B. Balaam. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 558–561.) Description of the construction and characteristics of a series of counter tubes for various specific purposes. All are of the cylindrical gas-filled type, with halogen or organic quenching agent.

621.387.424 : 537.525.92 **798**

Remark on the Space-Charge Sheath of the Geiger Counter.—D. H. Wilkinson. (*Rev. sci. Instrum.*, Sept. 1952, Vol. 23, No. 9, pp. 463–464.) The popular notion of the Geiger-counter space-charge sheath as a thin expanding shell requires modification. Owing to the considerable charge transported in the sheath, it grows in thickness as it crosses to the cathode and may occupy more than half the total volume of the counter. The attendant wide spread in arrival time of the positive ions at the cathode may explain, to some degree, the time distribution of spurious counts.

621.387.424 : 539.26 **799**

A [Geiger-Müller] **Counter Arrangement for X-Ray Interference Measurements.**—R. Berthold & A. Trost. (*Schweiz. Arch. angew. Wiss. Tech.*, Sept. 1952, Vol. 18, No. 9, pp. 277–282.)

621.791.3 : 534.321.9 **800**

Ultrasonic Tinning Techniques for Aluminum.—A. E. Crawford. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 102–105.) The ultrasonic iron described uses a self-driving longitudinally excited magnetostriction element made of Co-Fe alloy laminations and operated as a $\lambda/2$ resonator. The frequency is about 20 kc/s. A proposed plant for continuous tinning of Al wire is also described.

621.791.3 : 534.321.9 **801**

Ultrasonic Tinning of Aluminium.—P. Wenk & H. Boljahn. (*Z. Metallkde*, Sept. 1952, Vol. 43, No. 9, pp. 322–324.) Description of an ultrasonic soldering tool using a 20-kc/s Ni magnetostriction oscillator, with illustrations of tests on sheet Al, using pure Sn as solder.

PROPAGATION OF WAVES

538.566.029.45/.51 : 551.594.6 **802**

Propagation of Very Long Electromagnetic Waves, and the Wave Spectrum of Lightning.—W. O. Schumann. (*Naturwissenschaften*, Oct. 1952, Vol. 39, No. 20, pp. 475–476.) Regarding lightning as a Dirac current pulse, a formula derived for the wavelength of maximum intensity (λ_m) shows that λ_m increases as the square of the distance from the point where the flash occurs. The impulse received at a distance from the flash is the Fourier integral of the spectral components of the wave, a formula for which is given. The actual pulse shape for a lightning flash probably favours the lower frequencies in the spectrum. A detailed account of the investigation is to be published in *Z. angew. Phys.*

538.566.2 **803**

A Method of Solving the Wave Equation in a Region of Rapidly Varying Complex Refractive Index.—J. J. Gibbons & R. L. Schrag. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1139–1142.) The equation for wave propagation in an ionized medium is transferred into an integro-differential equation with only one real dependent variable, thus avoiding the need to solve two simultaneous differential equations. The solution yields one of the two wave functions directly, the second being derived from the first by direct integration. The method is illustrated by computing the reflection coefficient for a region where the refractive index passes through a sharp peak.

621.396.11 **804**

Scattering of Electromagnetic Energy in a Randomly Inhomogeneous Atmosphere.—H. Staras. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1152–1156.) First-order perturbation theory is used to derive an integral representing the scattered power at a receiver resulting from random inhomogeneities in the propagating medium. The expression obtained corresponds with that used by Booker & Gordon (1757 of 1950), but instead of the space-correlation function of refractive index used by them a time-correlation function is introduced which permits evaluation of the time average of the scattered power; this time-correlation function is directly measurable. For small-scale turbulence the average scattered power is not affected by the particular model of atmospheric turbulence chosen; for large-scale turbulence the results depend on the particular time-correlation function chosen and on particular assumptions regarding the scattering bodies.

621.396.11 **805**

Effect of Magnetic Field in Oblique Propagation over Equatorial Region.—B. Chatterjee. (*Indian J. Phys.*, June 1952, Vol. 26, No. 6, pp. 297–312.) Curves of the Booker type (422 of 1939) are presented for the case of propagation across an equatorial region, and the phenomenon of lateral deviation is discussed. Propagation curves are also given for the particular transmission paths Calcutta-Bandoeng and Calcutta-Bombay. The calculations were made for the case of a flat stratified layer over a flat earth, so that correction factors are required to take account of the earth's curvature.

621.396.11.029.51 : 551.510.535 **806**

The Polarization of Vertically Incident Long Radio Waves.—J. M. Kelso, H. J. Nearhoof, R. J. Nertney & A. H. Waynick. (*Ann. Géophys.*, Oct./Dec. 1951, Vol. 7, No. 4, pp. 215–244. In English.) Analytical expressions relating the distribution of electron concentration and collision frequency in the ionosphere to the polarization characteristics are derived, the wave being treated as a single magneto-ionic component. Measurements made on a frequency of 150 kc/s are reported and interpreted. A model of the D and E layers is assumed which gives theoretical results in good agreement with experimental observations. See also 517 of February (Gibbons & Nertney).

621.396.11.029.55 : 551.510.535 **807**

Ionospheric Measurements at Oblique Incidence over Eastern Australia.—M. Strohfeldt, R. W. E. McNicol & G. de V. Gipps. (*Aust. J. sci. Res., Ser. A*, Sept. 1952, Vol. 5, No. 3, pp. 464–472.) An account of attempts to identify night-time ionosphere reflecting layers by measuring apparent path lengths of pulsed transmissions on 5·8 Mc/s over a baseline of 763 km, using responder technique. The characteristics of beacon triggering are discussed in relation to the type of echo received. Correlation was established between occurrences of E_s observed at oblique incidence and at vertical incidence near the mid-point of the trajectory. Unusual records of Pedersen rays are shown, and sudden height increases and diffuseness of F_2 echoes are discussed. A check on the oblique-incidence theory, using a Millington transmission curve in conjunction with vertical-incidence $h'f$ records, yielded reasonable agreement between measured and deduced reflection heights. A rough analysis of oblique-incidence penetrations showed that the average frequency separation of the ordinary-ray and the extraordinary-ray m.u.f. was about half the gyromagnetic frequency.

621.396.81 **808**

A Comparison of C.W. Field Intensity and Backscatter Delay.—W. L. Hartsfield & R. Silberstein. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1700–1706.) The relation between back-scatter and skip phenomena was investigated by comparing the intensity of the 15-Mc/s WWV signals, received at White Sands in New Mexico, with the recorded delay times for the back-scatter received at Sterling, Virginia, from a pulse transmitter operating at about the same frequency. Rapid variations in the back-scatter records for disturbed days correspond in order of magnitude with previously observed motions of ionosphere irregularities.

621.396.81.029.62 **809**

Field Strengths Recorded on Adjacent F.M. Channels at 93 Mc/s over Distances from 40 to 150 Miles.—G. S. Wickizer & A. M. Braaten. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1694–1699.) Analysis of records of the field strength received from stations KE2XCC (93·1 Mc/s, Alpine, N.J.) and WBZ-FM (92·9 Mc/s, Boston, Mass.) over a period of more than a year at two places on Long Island. For the evening hours there is a seasonal trend towards higher intensities in summer. The overall variations were larger for the longer transmission paths. The hourly distribution curves are discussed in relation to possible modes of propagation.

RECEPTION

621.396.62 **810**

Receiver Production of the VVB-RFT.—A. Blaha. (*NachrTech.*, Sept. 1952, Vol. 2, No. 9, pp. 261–264.) Short descriptions of the special features of some of the best types of receiver produced by this East German nationalized industry in 1952.

621.396.62(083.7) **811**

Standards on Receivers: Definitions of Terms, 1952.—(*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1681–1685.) Standard 52 IRE 17.S1.

621.396.621 : 621.396.662 **812**

A Method of Band-Spreading.—C. A. Parry. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1952, Vol. 13, No. 10, pp. 365–369.) Analysis of a capacitive type of circuit for use in communication receivers shows that it is possible to obtain frequency bands of equal width and constant gain with simple circuits, certain requirements being imposed on the tuning inductors used on each range. Errors due to stray capacitance increase with frequency and cannot be neglected beyond a certain limit, but this limit can be made to lie above the highest operating frequency by suitable circuit design. A suggested design procedure is outlined.

621.396.621.54 : 621.385.5 **813**

Application of the DK 92 Tube on 30 Mc/s.—H. H. van Abbe & J. Jager. (*Electronic Applic. Bull.*, Jan. 1952, Vol. 13, No. 1, pp. 1–7.) Full circuit details are given of a frequency changer for the ranges 3·3–10 Mc/s and 8·8–30 Mc/s. A single Type-DK92 heptode is used; this is a 50-mA miniature heptode with a variable-μ characteristic making it suitable for a.g.c.

621.396.621.54 : 621.396.82 **814**

Microphony in Superhet Oscillators.—H. Stibbé. (*Wireless World*, Dec. 1952, Vol. 58, No. 12, pp. 504–506 & Jan. 1953, Vol. 59, No. 1, pp. 35–38.) Microphony in the oscillators of superheterodyne receivers is caused by a frequency-discriminator action of the i.f. amplifier plus detector when the tuning is not quite correct. It can be prevented in some circumstances by using an over-critically coupled i.f. amplifier, thus permitting a greater degree of detuning before the discriminator action starts. A numerical calculation is made of the effect of vibration of tuning-capacitor plates for a typical case; the figures indicate that the highest possible degree of rigidity and symmetry are required in the assembly of this component. Methods are described for mounting it with good mechanical insulation. Measurements of the microphony-output of a receiver are also described.

621.396.822 : 621.396.62 **815**

Noise in Receivers and Amplifiers.—S. Gratama. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 207–247.) The various causes of noise in receivers are surveyed. The physical mechanisms of shot effect, induced grid noise and total-emission noise in valves are explained. Practical use can be made of the correlation which exists between induced grid noise and shot effect to reduce the noise caused by the latter. 59 references.

621.396.822 : 621.396.621 **816**

A Note on the Approach of Narrow-Band Noise after a Nonlinear Device to a Normal Probability Density.—G. R. Arthur. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1143–1144.) From the integral equation derived by Kac & Siegert (3645 of 1947) for the output of a low-pass filter preceded by a nonlinear device excited by a noise source, the first three central moments of the probability density of the output signal are obtained and the approach of these moments to those of a Gaussian density is demonstrated.

621.396.822.029.62 **817**

V.H.F. Radio Noise.—E. G. Hamer. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, p. 43.) Total-noise measurements were made at typical sites, during the latter part of 1950, at frequencies of 77 and 172 Mc/s. The noise level was much less in the country and in residential areas than in industrial areas; there was no noticeable difference between the levels obtaining with horizontal and with vertical polarization. The variation of noise level with distance from a main road was also investigated. The general atmospheric noise level was greater in all cases than that due to thermal noise alone, but was less at the higher frequency; in an industrial area the net result might be appreciably better reception at 172 than at 77 Mc/s.

621.396[.397].828 **818**

Radio Interference Suppression. [Book Review]—G. L. Stephens. Publishers: Iliffe & Sons, London, 2nd edn 1952, 132 pp., 10s. 6d. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, p. 585.) A practical guide to the various methods of eliminating interference with radio and television reception.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 : 519.272 **819**

Contribution to the Statistical Study of Communications.—S. Malatesta. (*Alta Frequenza*, Aug./Oct. 1952, Vol. 21, Nos. 4/5, pp. 163–198.) An introduction to the application of statistical methods in network theory, making use of the spectral-density and correlation functions and a criterion of network efficiency based on distortion. The statistical method is applied to the optimum noise filter of Wiener.

621.395.521.3 : 621.396.97 **820**

A Variable Equalizer for Broadcast-Programme Circuits on Trunk Lines.—C. M. Hall. (*Telecommun. J. Aust.*, Oct. 1951, Vol. 8, No. 5, pp. 311–313.) Description of equipment designed to facilitate the work involved in equalizing trunk routes in Queensland, where several circuits are over 800 miles in length and two exceed 1 000 miles. The equalizer finally adopted includes three units, one for correction of the low-frequency slope of the response curve, the next for the response hump usually occurring at about 150 c/s, and the last section for the high-frequency response. Plug-in attenuation pads are provided so that any reasonable degree of equalization can be obtained in each section.

621.396.333 + 621.396.5] : 621.396.71 **821**

Navy V.L.F. Transmitter will radiate 1 000 kW.—T. D. Hobart. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 98–101.) Description of transmitter sited at Jim Creek Valley, near Arlington, Washington, and planned to provide both c.w. and frequency-shift teleprinter communication with ships throughout the Pacific area, including submerged submarines. The twin 500-kW power amplifiers use a push-pull arrangement of Type-5831 water-cooled triodes, with 6-V thoriated tungsten filaments taking about 13 kW of heating power, and 11·5 kV anode voltage; each amplifier feeds half of the aerial. The tuning range is 14·5–35 kc/s. The aerial is suspended between two 3 000-ft mountain ridges, and comprises a horizontal zig-zag of 10 spans, arranged in two groups of five, with a 900-ft down-lead at the mid-point of each span. Facilities for a 46-man staff are provided on the 7 000-acre site.

621.396.4 **822**

A Beam Radio System with Pulse-Phase Modulation for 12- and 24-Channel Telephony Transmission.—E. Hölzler & H. Holzwarth; H. Holzwarth & W. Arens; E. Schulz, G. Piefke & E. Seibt; W. Wild, U. v. Kienlin & H. Simon. (*Fernmeldetech. Z.*, Sept. & Oct. 1952, Vol. 5, Nos. 9 & 10, pp. 397–405 & 456–467.) Section A, Survey, by Hölzler & Holzwarth, outlines the general characteristics of the system, whose 4-Mc/s frequency band can be selected at will in the range 2·45–2·7 kMc/s.

Section B, Modulation Equipment, by Holzwarth & Arens, describes in detail the modulation and demodulation arrangements for the two 12-channel groups.

Section C, High-Frequency Equipment, by Schulz, Piefke & Seibt, gives an account of the transmitters and receivers, and of the ring-modulator type of mixer used at the receiver input. One transmitter, with a power of 0·5 W, uses a disk-seal triode, Type 2C40, the other, with an output of 5 W, uses a Type-2C39A valve. Tuning is effected by varying the length of the grid cylinder. Control and monitoring equipment is also noted.

Section D, Aerial System, by Wild, v. Kienlin & Simon, describes the lens-type and parabolic aerials used, and also the filter and feeder arrangements. The radiation diagram of a parabolic aerial 3 m in diameter is shown and the operating characteristics of aerials respectively 3 m, 2 m and 1·2 m in diameter are compared.

621.396.5 : 621.396.8 **823**

Comparison of Mobile Radio Transmission at 150, 450, 900, and 3 700 Mc/s.—W. R. Young, Jr. (*Bell Syst. tech. J.*, Nov. 1952, Vol. 31, No. 6, pp. 1068–1085.) An account is given of tests conducted in and around New York City. Sufficient test locations were used to give a statistical indication of the trend of performance with frequency variation. Variations of aerial gain and frequency stability with frequency variation are taken into account. The transmitter power required to achieve the same coverage at different frequencies depends on the variation with frequency of both path loss and strength of signal required to produce satisfactory communication. The combination of these factors fixes a broad optimum-frequency band at about 500 Mc/s; this frequency band is more suitable than all the others for a mobile radio-telephone service. The 900-Mc/s band may be preferable to the 150-Mc/s band if full use is made of aerial gain, but above 900-Mc/s performance falls off rapidly.

621.396.712 : 621.395.6 **824**

Speech Input Systems for Broadcast Transmitters.—S. Hill. (*J. Brit. Instn Radio Engrs*, Oct. 1952, Vol. 12, No. 10, pp. 533–541.) Text of paper presented at 1951 Radio Convention, London. Technical and economic factors involved in the design of the a.f. equipment are considered. Layout and switching requirements are reviewed. Microphones, amplifiers, level indicators, faders, mixers and recording arrangements are discussed.

621.396.712.3 : 534.84 **825**

The 'Pierre Bourdan' Low-Frequency [broadcasting] **Centre, Paris.**—L. Conturie. (*Onde élect.*, Oct. 1952, Vol. 32, No. 307, pp. 397–410.) A detailed account of the general lay-out of the establishment, the various studios and control and recording rooms, with particular reference to the methods of construction giving good sound insulation between the studios and their surroundings, and to the treatment of studio walls, ceilings, etc., to obtain the desired acoustic properties for faithful recording or high-quality transmission of programmes. See also 3318 of 1952 (Pujolle).

621.396.712.3 : 621.396.6 **826**

Broadcasting-Studio Installations and the New S.F.R. Equipment.—J. Cordonnier & M. Bernard. (*Onde élect.*, Oct. 1952, Vol. 32, No. 307, pp. 411–422.) The relative merits of centralized and decentralized installations are discussed and descriptions are given of new equipment units, developed by the Société Française Radioélectrique, which are economical in use and retain all the essential advantages of the mixed type of installation. The units have been designed so that various combinations can be adopted to meet the requirements of different studios; they include microphone and line amplifiers, attenuators, etc., which can be assembled into monobloc programme consoles. Performance data are tabulated and illustrations are given of typical units and assemblies, including sound-pickup consoles and consolettes.

621.396.822 : 621.395.44 **827**

Intermodulation Noise.—J. L. Bordewijk. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 261–279.) Intermodulation noise in multichannel telephony systems increases as the signal level rises, in contrast to noise from other sources, which becomes more noticeable as the signal level decreases. An optimum signal/total-noise ratio generally occurs at the point where intermodulation noise and noise due to other causes are about equal. The intermodulation-noise spectrum can be calculated either from the intermodulation products or by using correlation functions.

621.396.822 : 621.396.619.1 **828**

Signal/Noise Ratio for Various Modulation Systems.—F. L. Stumpers. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 249–260.) Comparison is made between the two broad groups of modulation methods respectively using (*a*) nonquantized and (*b*) quantized signals. The signal/noise ratio in the output of the different systems is related to the signal/noise ratio in the transmission channel and to the required bandwidth. When nonquantized signals are used the effects of noise in the successive stages of the system are cumulative; when quantized signals are used (as in p.c.m. and delta modulation) noise is introduced by the initial quantization process but the system is nearly immune to channel noise. For ratios of useful energy to noise energy in the transmission channel greater than a threshold value of about 20 db, the quantized system is practically fault-free.

621.396.93 **829**

Maritime Distress Frequency.—W. Blow. (*Wireless World*, Jan. 1953, Vol. 59, No. 1, p. 16.) The frequency of 1·65 Mc/s hitherto used in European waters is to be replaced by 2·182 Mc/s from 1st May 1953; this will be a world-wide distress and calling frequency.

SUBSIDIARY APPARATUS

621-526 **830**
Servomechanisms, a Survey.—G. R. Arthur. (*J. Brit. Instn Radio Engrs*, Oct. 1952, Vol. 12, No. 10, pp. 507–516.) Design techniques discussed include frequency and time analysis and statistical methods; an indication is given of problems not yet solved. 55 references.

621-526 **831**
Nonlinear Servomechanism.—J. Loeb. (*Onde élect.*, Nov. 1952, Vol. 32, No. 308, pp. 431–437.) The various factors limiting the application of the linear theory of servomechanisms are discussed, the theory of 'filtered' systems using relays, developed independently by Dutilh (743 of 1951) and Kochenburger (*Elect. Engng, N.Y.*, Aug. 1950), is outlined, and two new criteria applicable to all 'filtered' servomechanisms are established. The first criterion is concerned with the possibility of hunting taking place and includes the criteria of Nyquist and Kochenburger as special cases. The second criterion determines the stability of such oscillations of the system.

621-526.001.11 **832**
A Formula for an Integral occurring in the Theory of Linear Servomechanisms and Control Systems.—H. Bückner. (*Quart. appl. Math.*, Oct. 1952, Vol. 10, No. 3, pp. 205–213.)

621.311.6 : 621.396.615 : 621.314.7 **833**
Application of Transistors to High-Voltage Low-Current Supplies.—G. W. Bryan, Jr. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1521–1523.) A transistor oscillator is used to develop the h.v. required for such devices as Geiger-Müller counters. The oscillations have a sawtooth waveform, the flyback being used for shock excitation of the h.v. transformer.

621.314.632 : 546.289] + 621.314.7 **834**
Power Rectifiers and Transistors.—R. N. Hall. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1512–1518.) Power rectifiers with rectification ratios as high as 10^7 can be made by fusing donor and acceptor contacts to the opposite faces of a Ge wafer. Analysis of the characteristics of such rectifiers gives results in good agreement with experimental values. The properties of transistors prepared in a similar manner, and capable of outputs as high as 100 W, are described. At present the operation of these power units is limited to about 20 kc/s by transit-time effects.

621.314.632 : 546.289 **835**
A High-Voltage, Medium-Power Rectifier.—C. L. Rouault & G. N. Hall. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1519–1521.) A description is given of the operating characteristics of *p-n*-junction rectifiers prepared by fusing impurity metals to Ge wafers. The addition of cooling fins enables higher powers to be handled.

621.314.65 **836**
The 'Nevitron' Mercury-Arc Rectifier.—(*Engineering, Lond.*, 19th Sept. 1952, Vol. 174, No. 4521, p. 373; *Overseas Engr*, Dec. 1952, Vol. 26, No. 300, p. 163.) Description of a new type of rectifier with the Hg pool in a Mo cup with external Al cooling fins. Provided the Mo cup is wetted by the Hg to give a concave meniscus, the cathode spot runs in a continuous line round the edge of the Hg. The cooling system ensures no excessive emission of vapour. An auxiliary electrode, lifted from the Hg by means of a solenoid, serves to start the arc. The voltage drop across the arc is 12·5 V. The weight of a 50-A 500-V Nevitron is only 2·5 lb, excluding the ignition solenoid. Types with grid control have also been tested. The power required in the grid circuit for full control is only one-thousandth of that for a multianode rectifier.

621.316.722.1 : 621.387 **837**
Improved Stabilization from a Voltage-Regulator Tube.—M. D. Armitage. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 568–569.) By using a suitable barretter in place of the resistor usually connected in series with a voltage-regulator tube, a definite improvement in performance is obtained. The barretter type of circuit is most useful when the load current is relatively high.

621.316.722.1.027.3 **838**
Stabilizer for Control of High Direct Voltages.—J. Serny. (*Rev. gén. Élect.*, Sept. 1952, Vol. 61, No. 9, pp. 411–420.) The development and performance of a stabilizer for direct voltages of the order of several thousands of volts are described. The application of the stabilizer to voltages of any value (*a*) using as variable resistor a large number of triode valves in series, (*b*) suppressing the voltage-divider bridge generally used to control the regulator system, is also discussed. Rectifier ripple can be eliminated, so that a simple type of filter can be used.

621.318.435.3 : 621.311.62 **839**
The Transbooster.—A. H. B. Walker. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 546–550.) Description of circuits in which a transductor is used for regulation of the output voltage of a rectifier, fed from a.c. mains, under varying load conditions. If the transductor is connected on the d.c. side of the main rectifier, a smaller transductor can be used, since it will only have to handle the total boost voltage required to cover rectifier regulation and mains-voltage variation.

621.316.7 : 621-526 **840**
Automatic Feedback Control. [Book Review]—W. R. Ahrendt & J. F. Taplin. Publishers: McGraw Hill, New York, 1951, 412 pp., 64s. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, p. 584.) Theory and applications of servo control systems.

TELEVISION AND PHOTOTELEGRAPHY

621.397 **841**
A High-Speed Direct-Scanning Facsimile System.—C. R. Deibert, F. T. Turner & R. H. Snider. (*Elect. Engng, N.Y.*, Sept. 1952, Vol. 71, No. 9, p. 784.) Digest of paper presented at A.I.E.E. General Meeting, January 1952. A description is given of the Western Union system, which handles copy up to 8½ in. by 15 in. at the rate of 2½ in.²/sec. Using d.s.b. transmission, the required bandwidth is 30 kc/s. The transmitter comprises two identical scanning units, which are used alternately to save time while copy is changed. The copy-holding cylinder rotates at 1 800 r.p.m., and the controls are partly electromechanical and partly electronic, with tuning-fork frequency standards for synchronization.

621.397.5 **842**
Television Programme Origination: The Engineering Technique.—D. C. Birkinshaw. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 43–73. Discussion, pp. 174–178.) A comprehensive review of the development of the B.B.C. television service including discussion of apparatus and techniques for studio and outside broadcasts and for telecine and telefilm recording.

621.397.5 **843**

Television Programme Production Problems in Relation to Engineering Technique.—I. Atkins. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 74–81. Discussion, pp. 174–178.) Discussion of production planning, camera technique and lighting problems.

621.397.5 **844**

Determination of the Number of Lines to be chosen for a Television System, as dependent on the Size of the Receiver Screen.—P. Stroobants. (*Onde élect.*, Nov. 1952, Vol. 32, No. 308, pp. 438–444.) See 3245 of 1952.

621.397.5 : 061.4 (443.611) **845**

The Second Television Salon.—(*Onde élect.*, Nov. 1952, Vol. 32, No. 308, pp. 464–466.) Discussion of design trends for receivers, c.r. tubes, aerials and feeders, as exemplified in the equipment on show at the 1952 Salon. For other accounts see *Télévision*, Nov. 1952, No. 28, pp. 255–258; *TSF et TV*, Nov. 1952, Vol. 28, No. 289, pp. 333–335; *Toute la Radio*, Nov. 1952, Vol. 19, No. 170, pp. 398–402; *Radio Télév. prof., Paris*, Oct. 1952, Vol. 21, No. 210, pp. 14–15. .20.

621.397.5 : 534.86 **846**

Problems of Sound in Television Programmes.—R. F. A. Pottinger. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 145–149. Discussion, pp. 174–178.)

621.397.5 : 778.5 **847**

Television Recording.—W. D. Kemp. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 115–127. Discussion, pp. 174–178.) Various methods of photographic recording on film, with intermittent or continuous motion of the film, are discussed and the two methods now used by the B.B.C. are described in some detail.

621.397.5(091) **848**

The History of Television.—G. R. M. Garratt & A. H. Mumford. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 25–40. Discussion, pp. 40–42.) A review of developments in various countries. 45 references.

621.397.6 : 535.317.5 **849**

A 5 : 1 Television Zoom Lens.—H. H. Hopkins. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 109–112. Discussion, pp. 174–178.) Performance requirements of zoom lenses for television are discussed and a lens system satisfying the requirements, designed for the B.B.C., is described. The aberrations of such lens systems are considered and methods of correction are outlined.

621.397.6 : 621.385.832 **850**

The Monoscope.—R. D. Nixon. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17. pp. 132–135. Discussion, pp. 174–178.) Discussion of the operating principles and factors affecting the design of c.r. tubes for generating a stationary-picture signal. The picture is printed on a conducting plate, usually Al, using printing material finally converted to carbon with a low secondary-emission ratio. Since Al has a relatively high secondary-emission ratio at low voltages, scanning of the picture by the c.r. beam results in a picture signal being produced in the lead connecting the Al plate to the final accelerator electrode. Such devices have recently been used by the B.B.C. for test-card transmission. See also 2865 of 1938 (Burnett).

621.397.6 : 778.5 **851**

A Continuous-Motion System for Televising Motion-Picture Films.—H. E. Holman & W. P. Lucas. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 95–108. Discussion, pp. 174–178.) Detailed description of B.B.C. equipment for 35-mm film and of the special equipment required for 16-mm film on account of the increased magnification from film to viewing screen. The flying-spot system is used.

621.397.611 : 778.5 **852**

The Development of a High-Quality 35-mm Film Scanner.—T. C. Nuttall. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 136–144. Discussion, pp. 174–178.) Detailed description of B.B.C. equipment using the flying-spot system.

621.397.611.2 **853**

A Small High-Velocity-Scanning Television Pickup Tube.—J. E. I. Cairns. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 89–94. Discussion, pp. 174–178.) The tube described, scaled down from one of the super-emitron type [756 of 1951 (McGee)], has a superior performance and is two or three times as sensitive as its predecessor. This superiority is due to increased mosaic storage, which reduces the shading signal but increases the picture lag. The lag is unnoticeable except for very rapidly moving objects.

621.397.611.2 **854**

The Influence of Tube Characteristics and other Factors on Camera Design.—L. H. Bedford. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 82–88. Discussion, pp. 174–178.) The principal types of camera tube are discussed and their transfer characteristics are shown graphically. Factors affecting camera design are considered and the principal features of eight modern cameras are tabulated.

621.397.611.2 **855**

An Investigation into the Use of Secondary-Electron Signal Multipliers in Image Iconoscopes.—R. Theile & H. McGhee. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 159–165. Discussion, pp. 174–178.) By using an electron-transmissive screen in front of the first multiplier dynode, complete collection and sufficient acceleration of the secondary electrons leaving the target can be accomplished. Further, with a suitable geometrical arrangement of the multiplier and target, uniform picture generation can be achieved over the whole target area. A practical assembly is described and its performance discussed.

621.397.611.2 **856**

Design Features of a Television Camera with a Single-Lens Optical View-Finder.—T. Worswick & J. L. Bliss. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 166–173. Discussion, pp. 174–178.) Detailed description of a B.B.C. camera, of relatively small size and weight, with an optical view-finder located on top of the camera. A servo focusing-control system is used.

621.397.62 **857**

Intercarrier-Sound Television Receivers.—A. Boekhorst. (*Electronic Applic. Bull.*, Feb. 1952, Vol. 13, No. 2, pp. 21–33.) The basic principles of the intercarrier-sound system are described and the requirements imposed on the transmitter and on the response curve of the receiver are discussed. Three methods of separating the video and intercarrier signals are considered and a detailed description is given of (*a*) the detector and video amplifier, (*b*) the sound channel of a receiver supplying a large signal to the picture tube.

621.397.62 : 621.385.3 **858**

Stable Oscillator for U.H.F. TV Receivers.—Loofbourrow & Morris. (See 900.)

621.397.62 : 621.396.67 **859**

Community Antennas bring TV to Fringe Areas.—J. M. Carroll. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 106–111.) A system representative of many operating in the U.S.A. is described. Signals in the various v.h.f. channels are received by separate high-gain aerials mounted on high towers, and are passed through pre-amplifiers mounted high up the towers. Signals from odd and even channels respectively are combined and transmitted via coaxial cables to the tower base, where they are redistributed to separate-channel amplifiers, shifted from high-band to low-band channels, and fed to subscribers by a transmission-line system. Suitable aerials and amplifiers are discussed.

621.397.621 **860**

Reactive Time Bases.—A. B. Starks-Field. (*J. Brit. Instn Radio Engrs*, Oct. 1952, Vol. 12, No. 10, pp. 519–532. Discussion, p. 532.) Text of paper presented at 1951 Radio Convention, Cambridge. High-efficiency line-deflection circuits for large-screen television tubes are discussed. Various methods are described for recovering the energy in the deflecting field at the end of the scan; this energy can either be returned to the h.v. line or used to boost the voltage of the driver stage. The design of booster circuits is discussed with reference to efficiency, linearity and convenience of operation from h.v. supplies of the order of 200 V. Particular attention is paid to the design of the transformer, which may be either an auto-transformer or a multi-winding type. A system for operating directly on to high-impedance deflection coils is briefly mentioned.

621.397.621.029.63 **861**

One-Channel Converter for U.H.F. Television.—Wen Yuan Pan. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 134–138.) Printed inductors and Ge-diode mixers are used in a commercially available converter for shifting signals from any given u.h.f. channel to a selected channel in the lower v.h.f. band. The whole of the u.h.f. television band is covered by the grounded-anode oscillator circuit.

621.397.621.2 : 621.318.2 **862**

A New Focusing Unit for Television Picture Tubes.—P. van Tilburg & J. A. Verhoef. (*Electronic Applic. Bull.*, March/April 1952, Vol. 13, Nos. 3/4, pp. 37–43.) The properties of ferroxdure which make it particularly suitable for use in magnetic focusing units are described and details are given of a unit which uses two annular ferroxdure magnets. The strength of the focusing field is adjusted by variation of the separation of the two rings, which can be varied from 1 mm to 12 mm to cover a range of beam voltages from 5 kV to 20·5 kV.

621.397.7 : 621.316.7 **863**

Television Control-Room Lay-Out.—R. D. Chipp. (*Tele-Tech*, Oct. 1952, Vol. 11, No. 10, pp. 48–51.) Separate arrangements for audio, video, and direction control result in greater efficiency of operation. Suggested plans for large, medium, and small studios are presented and discussed.

621.397.7 : 628.972 **864**

Television Studio Lighting Equipment.—S. L. Johnson. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 113–114. Discussion, pp. 174–178.)

621.397.7 : 628.972 **865**

Television Lighting Technique.—H. O. Sampson. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 150–158. Discussion, pp. 174–178.) An account of arrangements adopted in B.B.C. studios.

621.397.7 : 628.972 : 621.327.4 **866**

Discharge Lamps for Television Studios.—E. H. Nelson & W. A. Price. (*Proc. Instn elect. Engrs*, Part IIIA, April/May 1952, Vol. 99, No. 17, pp. 128–131. Discussion, pp. 174–178.)

621.397.82 : 621.396.621 **867**

The Maximum Permissible Interference Radiation from U.S.W. Receivers.—(*Radio Tech., Vienna*, Oct. 1952, Vol. 28, No. 10, pp. 425–426.) With the introduction of television in Germany, principally on frequencies in the band 174–216 Mc/s, interference from the second harmonic of the local oscillator of s.w. f.m. broadcasting receivers has been experienced. It has been decided that such interference must be reduced below the level of 30 μV/m at a distance of 30 m from an aerial connected to a receiver. Equipment suitable for making the necessary measurement is noted.

621.397.822 **868**

Noise Measurements on Television Transmissions.—R. Rasch. (*Fernmeldetech. Z.*, Oct. 1952, Vol. 5, No. 10, pp. 440–444.) Noise voltages of amplifiers are usually measured in terms of effective values, while for television picture signals peak values are used. Equipment is described for c.r.o. comparison of peak and effective values of noise voltages. A correction of 15 db should be applied when comparing effective values with peak-to-peak values. A method of determining the permissible noise level in a television transmission system is outlined and illustrations are given of the effect on picture quality of a progressive reduction of the signal/noise ratio.

621.396/.397].828 **869**

Radio Interference Suppression. [Book Review]—Stephens. (See 818.)

TRANSMISSION

621.396.61 : 621.396.611.3 : 621.396.67 **870**

Transmitter Combining Circuits.—A. R. A. Rendall & G. A. Hunt. (*Electronic Engng*, Dec. 1952, Vol. 24, No. 298, pp. 550–552.) Description of typical circuits used at B.B.C. unattended stations for coupling two or more transmitters, operating on a common frequency, to a single aerial. For transmitter powers < 1 kW a special hybrid coil, wound on an iron-dust core, has been developed. The circuits used include both hybrid and bridged-T circuits.

VALVES AND THERMIONICS

621.314.6/.7 : 621.396.822 **871**

On the Theory of Noise in *P-N* Junctions and Related Devices.—R. L. Petritz. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1440–1456.) Noise resulting from fluctuations inherent in the electronic system of a *p-n* junction is investigated and found to be a result of fluctuations of the concentration of the minority carriers. A noise theory based on a lumped-parameter representation of a *p-n* junction is developed, and an equivalent circuit, with appropriate noise generator, is derived. Noise characteristics of *p-n*-junction rectifiers and transistors are analysed. The available noise power of a *p-n*-junction rectifier is voltage dependent, its equation resembling that of Weisskopf for point-contact rectifiers. The noise figures of *p-n*-junction transistors are of the order of unity and are independent of size and current density. A comparison is made, as regards noise, between point-contact and *p-n*-junction rectifiers and transistors, using the Weisskopf noise formula (M.I.T. Rad. Lab. Series, No. 133,

1943) for the point-contact devices. A relation between the noise spectrum and the admittance of a *p-n* junction is obtained. Fluctuation noise constitutes only a part of the measured noise of point-contact and *p-n*-junction rectifiers and transistors. Another source of noise, connected with control of mean current, is required to account for (*a*) the noise figures of *p-n*-junction and point-contact transistors being appreciably greater than unity, (*b*) the large difference between the noise figures of *p-n*-junction and point-contact transistors, (*c*) the $1/f$ law of the frequency spectrum of the measured noise.

621.314.63 : 546.289 **872**

On Some Transients in the Pulse Response of Point-Contact Germanium Diodes.—M. C. Waltz. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1483–1487.) To explain the hole-storage effect noted by Michaels & Meacham (1817 of 1950) a hypothesis is proposed which postulates the presence of traps in the Ge *p* layer near the point electrode. Measurements and calculations indicate trap densities of the order of $10^{16}/cm^3$, trap depths in the energy band of about 0·3 eV, and capture cross-section diameters of about 0·3 Å.

621.314.7 **873**

Present Status of Transistor Development.—J. A. Morton. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1314–1326.) Reprint. See 2651 of 1952.

621.314.7 **874**

Effects of Space-Charge-Layer Widening in Junction Transistors.—J. M. Early. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1401–1406.) Some effects of the dependence of the thickness of the collector barrier on collector voltage are analysed. The thickness of the base layer decreases as collector voltage increases, resulting in an increase of the current-gain factor α and a decrease of the emitter voltage required to maintain any given emitter current. These effects lead to the introduction of two new elements in the small-signal equivalent circuit: (*a*) the collector conductance, which is proportional to emitter current and varies inversely with collector voltage; (*b*) the voltage-feedback factor, which is independent of emitter current but varies inversely with collector voltage.

621.314.7 : 537.311.33 **875**

Transistor Electronics: Imperfections, Unipolar and Analog Transistors.—Shockley. (See 746.)

621.314.7 : 546.289 **876**

A Developmental Germanium *P-N-P* Junction Transistor.—R. R. Law, C. W. Mueller, J. I. Pankove (Pantchechnikoff) & L. D. Armstrong. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1352–1357.) A transistor of the *p-n-p* junction type can easily be made in the laboratory by diffusing indium into opposite faces of a single-crystal *n*-type Ge wafer. The characteristics of such units are illustrated by experimental curves obtained in tests of over 100 units.

621.314.7 : 546.289 **877**

Fused-Impurity *P-N-P* Junction Transistors.—J. S. Saby. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1358–1360.) Transistors of the *p-n-p*-junction type were produced by fusion of acceptor impurities so as to create *p*-type areas on opposite faces of a wafer of *n*-type Ge. The power dissipation of such units can be increased by the addition of metal cooling fins. The current multiplication factor α is nearly constant at about 0·95 up to 120°C and decreases slightly above this temperature. The power gain is high and noise figure low.

621.314.7 : 546.289 **878**

Four-Terminal *P-N-P-N* Transistors.—J. J. Ebers. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1361–1364.) A *p-n-p*-junction transistor and one of the *n-p-n*-junction type can be interconnected in such a way as to have an equivalent circuit identical with that of a *p-n-p-n* transistor. A simplified equivalent circuit is obtained for the case where the *p-n-p-n* transistor is used as a grounded-base hook-collector transistor.

621.314.7 : 546.289 **879**

A Unipolar 'Field-Effect' Transistor.—W. Shockley. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1365–1376.) The field-effect transistor consists of a layer of *p*-type material sandwiched between two layers of heavily doped *n*-type material, termed *n* +. The working current is carried by hole conduction in the *p*-type layer, between terminals consisting of heavily doped *p* + inserts. With reverse bias across the *p-n* junctions, the current flows in a channel of *p*-type material bounded by two space-charge regions with negligible carrier concentration. A theory of the action of such devices is presented, the new terms used for the various electrodes are defined, and design calculations are made for a unit made from Ge.

621.314.7 : 546.289 **880**

Junction Fieldistors.—O. M. Stuetzer. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1377–1381.) A description is given, with theory, of an amplifying device with high-impedance input and low-impedance output, which uses an auxiliary electrode close to a *p-n* junction to control the surface conductivity in the neighbourhood of the junction. The arrangement is similar to those previously described (3198 of 1950). A frequency cut-off in the a.f. range will limit the application of the device in its present form.

621.314.7 : 546.289 **881**

Theory of Alpha for *P-N-P* Diffused-Junction Transistors.—E. L. Steele. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1424–1428.) Equations are developed for the emitter and collector currents for *p-n-p*-junction transistors, and the current-gain factor α is deduced. The l.f. value of α and its h.f. cut-off value are markedly dependent on the thickness of the *n*-type 'base' region, the h.f. characteristics being better when this thickness is small. In grounded-emitter applications the h.f. characteristics depend more directly on the lifetime of holes, and show only second-order dependence on the base thickness; the shorter the lifetime the higher the cut-off frequency.

621.314.7 : 546.289 **882**

Effect of Electrode Spacing on the Equivalent Base Resistance of Point-Contact Transistors.—L. B. Valdes. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1429–1434.) An expression for the equivalent base resistance r_b is derived and is checked experimentally. Electrode spacing, and the thickness and resistivity of the Ge slice, have major effects on the value of r_b.

621.314.7 : 546.289 : 536.49 **883**

Variation of Transistor Parameters with Temperature.—A. Coblenz & H. L. Owens. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1472–1476.) Measurements were made of the variations with temperature of the parameters of 20 Western Electric Type-1698 and Type-1768 transistors over the range 25–85°C. The results are shown graphically and indicate that these transistors can operate satisfactorily for many small-signal applications up to about 60°C, the gain, e.g., being decreased by only about 2 db at this temperature.

621.314.7 : 621.3.016.352 **884**

The Control of Frequency Response and Stability of Point-Contact Transistors.—B. N. Slade. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1382–1384.) Satisfactory stability and frequency characteristics have been obtained by control of the point-contact spacing and the resistivity of the Ge used. By means of the methods outlined, transistors have been produced that can oscillate at frequencies considerably higher than 100 Mc/s, one reaching 300 Mc/s.

621.314.7 + [621.314.632 : 546.289 **885**

Power Rectifiers and Transistors.—Hall. (See 834.)

621.314.7 : [621.396.615 + 621.396.645 **886**

A Junction-Transistor Tetrode for High-Frequency Use.—Wallace, Schimpf & Dickten. (See 677.)

621.314.7 : 621.396.822 **887**

An Experimental Investigation of Transistor Noise.—E. Keonjian & J. S. Schaffner. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1456–1460.) Transistor noise is discussed and methods of measuring it are described. Experimental results give noise figures for point-contact transistors of about 50 db, while for *p-n*-junction transistors the values may be as low as 10 db at 1 kc/s. The noise figure of point-contact transistors was found relatively independent of the d.c. operating point, but for junction transistors the noise figure may vary considerably with collector voltage and to some extent with emitter current.

621.383.5 **888**

New Photoelectric Devices utilizing Carrier Injection.—K. Lehovec. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1407–1409.) The devices described are: (*a*) the photomodulator, which permits modulation of a light beam by the change in absorption due to injected carriers; (*b*) the graded-seal junction, which is prepared by fusing together two materials at a temperature intermediate to their melting points, with subsequent slow cooling. A theory of the phenomenon of electro-luminescence [1341 of 1951 (Payne et al.)] is proposed which is based on the injection of minority carriers.

621.383.5 **889**

Properties of the M-1740 *P-N* Junction Photocell.—J. N. Shive. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1410–1413.) Description of a cell evolved from the practical unit described by Pietenpol (2302 of 1951) and the work of Goucher et al. (1669 of 1951). The cell is only $\frac{3}{8} \times \frac{1}{4} \times \frac{3}{16}$ in., has low dark current, low noise and high sensitivity.

621.383.5 : 546.289 : 621.397.611.2 **890**

Use of the Flying-Spot Scanner to Study Photosensitive Surfaces.—J. I. Pantchechnikoff, S. Lasof, J. Kurshan & A. R. Moore. (*Rev. sci. Instrum.*, Sept. 1952, Vol. 23, No. 9, pp. 465–467.) Variations of photosensitivity over the surface of a large-area Ge photocell [2656 of 1952 (Pantchechnikoff)] are investigated by scanning with a flying spot from a c.r. tube and using the output of the photocell, after amplification, to control the beam intensity of a second c.r. tube.

621.385 : 621.396.822 **891**

A New Method of Calculating Microwave Noise in Electron Streams.—J. R. Pierce. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, pp. 1675–1680.) The approach to the problem is essentially the same as that of North (3420 of 1940). Linearized equations are used to calculate a frequency component of the noise excited in an electron beam by a charge having a velocity different from the mean velocity of the beam. Noise maxima and minima are found for a beam traversing a drift space. Results agree with values calculated by the Rack-Llewellyn-Peterson method (see 'Traveling-Wave Tubes', J. R. Pierce, Chap. 10).

621.385.032.21 : 061.3(47) **892**

Conference on Cathode Electronics.—I. Dykman. (*Zh. tekh. Fiz.*, Jan. 1952, Vol. 22, No. 1, pp. 175–182.) Summaries are given of the papers read at a conference held in Kiev on 4th–9th June 1951. The papers are grouped under the following headings: (*a*) general questions on the operation and structure of cathodes, (*b*) photoelectric effect, (*c*) secondary electron emission, (*d*) thermoelectron emission, (*e*) cathodes under discharge conditions or ionic bombardment.

621.385.032.213 : 546.431.221 **893**

The Electronic Properties of Barium Sulfide.—W. Grattidge & H. John. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1145–1151.) Results are reported of measurements made of the electron emission from BaS used as cathode coatings in planar diodes. For the most active of the samples used, the emission at temperatures of 900°K and over was comparable with that from pure BaO. Work function, conductivity and thermoelectric power were determined and the effect of Fe as an impurity was studied. The evaporation rate was found to be much less than that of BaO.

621.385.032.216 : 537.311.32 **894**

Conductivity of Oxide Emitters.—R. C. Hughes & P. P. Coppola. (*Phys. Rev.*, 15th Oct. 1952, Vol. 88, No. 2, pp. 364–368.) Measurements of the electrical conductivity of a (BaSrCa)O emitter over a range from room temperature to 1 100°K indicate the existence, in well activated cathodes, of a low-temperature conduction mechanism with an activation energy as low as 0·05 eV. A high-temperature conduction mechanism of 1-eV activation energy is noted for the temperature at which appreciable electron emission can be drawn. Exposure to Xe at a pressure of 25 atm causes a marked decrease of conductivity in the high-temperature range. These results are considered to confirm Loosjes & Vink's hypothesis (3208 of 1950) that in the high-temperature range conduction takes place mainly through space currents in the pores in the material. The low activation energy of the low-temperature conduction mechanism indicates that this conduction probably occurs in a monolayer of Ba on the surface of the oxide.

621.385.032.216 : 537.533.8 **895**

Secondary Electron Emission from Barium Oxide.—J. Woods & D. A. Wright. (*Brit. J. appl. Phys.*, Oct. 1952, Vol. 3, No. 10, pp. 323–326.) Report of an experimental investigation of the influence on the secondary-emission coefficient, δ, of the methods of preparing and operating the BaO layers. Variation of δ with temperature is small between room temperature and 600°C. For evaporated films under steady bombardment at 240 V the value of δ is about 3 for thickness 10^{-6} cm and about 2 for thickness 10^{-4} cm; with pulsed operation δ is considerably larger. For sprayed coatings the value of δ is about 2 for both steady and pulsed operation. Over a long period of operation the value of δ falls and decomposition of the oxide occurs.

621.385.032.216 : 546.431/.432]-31 **896**

Electrical Conductivity and Thermoelectric Power of (BaSr)O and BaO.—J. R. Young. (*J. appl. Phys.*, Oct. 1952, Vol. 23, No. 10, pp. 1129–1138.) Report of an investigation of the temperature dependence over the range 1 100°–300°K of the properties of oxide-cathode coatings at different states of activation and with Ni bases of different purities. Details are given of experimental procedure. No significant differences were found between the properties of BaO and (BaSr)O. Thermo-electric-power/temperature curves confirm theory developed by Hensley (737 above). Results cannot be explained on the basis of a simple one- or two-level semiconductor model, but give general support to the pore-conduction theory.

621.385.032.216 : 621.386 **897**

A Study of the Oxide-Coated Cathodes by X-Ray Diffraction Method: Part 1.—E. Yamaka. (*J. appl. Phys.*, Sept. 1952, Vol. 23, No. 9, pp. 937–940.) See 1158 of 1952.

621.385.032.216.2 **898**

Latest Disc-Cathode Developments.—(*Electronics*, Nov. 1952, Vol. 25, No. 11, pp. 236 . . 252.) C.r.-tube cathodes are described in which a ceramic disk is used as insulator between cathode and first grid. Improvements introduced include the use of more efficient alloys for cathode caps, techniques for maintaining critical spacings constant during long production runs, reduction of electron leakage across the ceramic disk and between heater and cathode, and elimination of heater shrinkage caused by damage during insertion.

621.385.032.24 : 537.533 **899**

Origin of Thermal Grid Emission and Investigations on its Elimination.—H. Köppen. (*NachrTech.*, Aug. 1952, Vol. 2, No. 8, pp. 246–247.) The results of investigations of grid currents in valves with grids and anodes of various materials and constructions show that such currents can be largely reduced by using grid materials with a high work function, by adopting a form of construction in which grid heating by radiation from the cathode is avoided as far as possible, and by choice of a suitable cathode-activation process.

621.385.3 : 621.397.62 **900**

Stable Oscillator for U.H.F. TV Receivers.—K. E. Loofbourrow & C. M. Morris. (*Electronics*, Dec. 1952, Vol. 25, No. 12, pp. 118–121.) The Type-6AF4 acorn oscillator triode is described. The influence of construction details on operating parameters is discussed. Power output is increased by silver-plating the leads; an increase from about 90 to 150 mW is observed for samples operating at about 1 kMc/s. Use of the valve in a receiver with an i.f. of 41·25–45·75 Mc/s is discussed; the required oscillator frequency range (in the region of 930 Mc/s) is obtained by using the valve with a $\lambda/2$ external line. Circuits are suggested capable of holding the drift to 500 kc/s in intercarrier-sound receivers.

621.385.832 **901**

The Optimum Space-Charge-Controlled Focus of an Electron Beam.—D. L. Hollway. (*Aust. J. sci. Res., Ser. A*, Sept. 1952, Vol. 5, No. 3, pp. 430–436.) A theoretical investigation is made of the defocusing of a beam of circular cross-section due to space charge; expressions for the condition of optimum focus are derived from the equations of the beam profile. Over a wide range of values of spot radius the optimum-focus expressions may be replaced by simpler formulae useful for dealing with electron-beam design problems.

621.387 **902**

The Thyratron as Switching and Control Valve and its Industrial Application Possibilities.—R. Hübner. (*Bull. schweiz. elektrotech. Ver.*, 20th Sept. 1952, Vol. 43, No. 19, pp. 760–764.)

621.387.032.212 **903**

Inertia Effects in Cold-Cathode Tubes.—M. O. Williams. (*Strowger J.*, July 1952, Vol. 8, No. 3, pp. 106–117.) The type of discharge in cold-cathode valves is examined both for the current-growth and current-decay periods. Measurement methods are outlined and typical oscillograms of current rise and decay with recurrent pulses are shown. Investigations with small-amplitude a.c. superimposed on the d.c. glow discharge reveal inertia effects of considerable magnitude and also complex-impedance effects. Results obtained on several types of valve are given in graph form; they show surprisingly high values of apparent inductance and appreciable values of effective resistance. The origin of the quadrature current in such valves is discussed.

621.396.615.141.2 **904**

Oscillations in a Nonslotted Magnetron in connection with Amplification by Space-Charge Waves.—R. Warnecke, H. Huber, P. Guénard & O. Doehler. (*C. R. Acad. Sci., Paris*, 25th Aug. 1952, Vol. 235, No. 8, pp. 493–494.) A formula for the frequency of oscillations in a whole-anode magnetron is confirmed experimentally. From a 'diocotron' oscillator of this type in which the frequency can be changed within wide limits by adjustment of the anode voltage, about 50 mW power has been obtained over a frequency band of more than an octave around 800 Mc/s.

621.396.622.6 : 546.28 **905**

Silicon *P-N* Junction Alloy Diodes.—G. L. Pearson & B. Sawyer. (*Proc. Inst. Radio Engrs*, Nov. 1952, Vol. 40, No. 11, pp. 1348–1351.) Acceptor or donor impurities are alloyed with *n*- or *p*-type Si to produce *p-n*-junction diodes with reverse currents as low as 10^{-10} A, rectification ratios as high as 10^8 at 1 V, stable Zener voltage which can be fixed, during the production process, at a value between 3 V and 1 kV, and ability to operate at temperatures as high as 300°C.

621.396.822 **906**

Symposium on Noise. General Introduction.—Casimir. (See 693.)

621.396.822 : 621.385.13 **907**

Valve Noise at Very High Frequencies.—G. Diemer. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 281–301.) At frequencies $>10^8$ c/s, for which electron transit times are not negligible, the finite duration and particular shape of current pulses induced by the individual electrons give rise to additional noise; total-emission noise increases considerably and space-charge smoothing becomes less effective. The correlation between the various causes of noise is emphasized, and the difference between the transmission process through the valve for signal and for noise is indicated. Various triode circuits are discussed in relation to noise factor, the value of which is affected by the feedback in the circuit.

MISCELLANEOUS

621.396.822 **908**

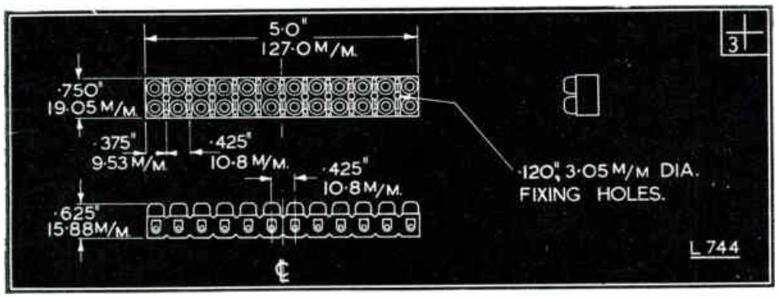
Symposium on Noise. Historical Introduction.—J. L. van Soest. (*Tijdschr. ned. Radiogenoot.*, Sept./Nov. 1952, Vol. 17, Nos. 5/6, pp. 197–198.) The subject is traced from Brown's investigations in 1827 of the movements of particles. In present-day electrical engineering the greatest emphasis is laid on signal/noise ratio, on account of its importance in relation to the transmission of information.

621.396/.397 **909**

Electronics for Communications Engineers. [Book Review]—J. Markus & V. Zeluff. Publishers: McGraw-Hill, New York, 1952, 601 pp., $10. (*Proc. Inst. Radio Engrs*, Dec. 1952, Vol. 40, No. 12, p. 1741.) A collection of papers published in *Electronics* during the past five years on the design of communication, broadcasting, television and radar equipment.


Grinding Mullard 'Ticonal' permanent magnets to required size.

The "Belling-Lee" page for Engineers



LIST No. L.744

CAPTIVE TERMINAL SCREWS

FLEXIBLE

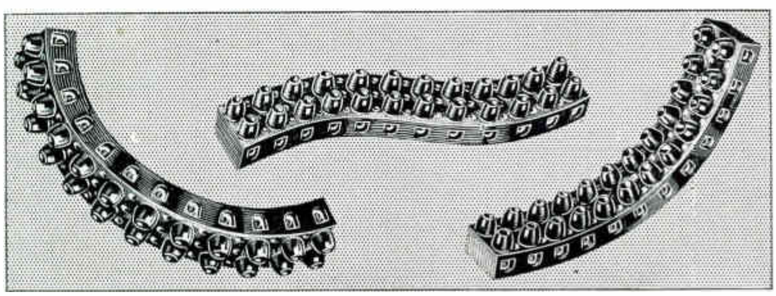
MECHANICALLY SHOCK-PROOF

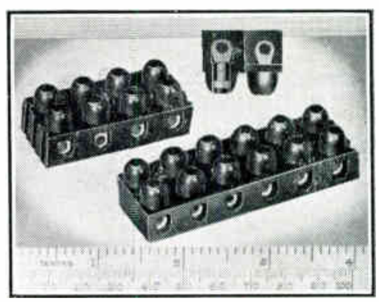
EASILY SECTIONALISED

A UNIQUE 5-AMP FLEXIBLE TERMINAL BLOCK

Engineers, designers, and those in close touch with radio and electronic instrumentation, have been quick to appreciate the numerous possibilities and applications of this versatile component, which represents an entirely new departure in terminal block design.

Moulded in P.V.C., and in consequence very flexible, the moulding securely grips the terminal screws, which are so dimensioned that they cannot fall out, *even if totally unscrewed from their inserts and the block mounted upside down.*





FULLER DETAILS ON REQUEST

PLEASE QUOTE P378/WE

A very useful feature of this component is its great flexibility in both planes, which enables it to be snugly secured to irregular surfaces or over uneven contours.

The block is also mechanically shock-proof, the nature of the moulding rendering it safe against breakage due to vibration. It can easily be sub-divided with an ordinary knife, to make smaller groups of terminal units. Each pair of terminals is provided with three fixing holes.

BELLING & LEE LTD
CAMBRIDGE ARTERIAL RD., ENFIELD, MIDDX., ENGLAND

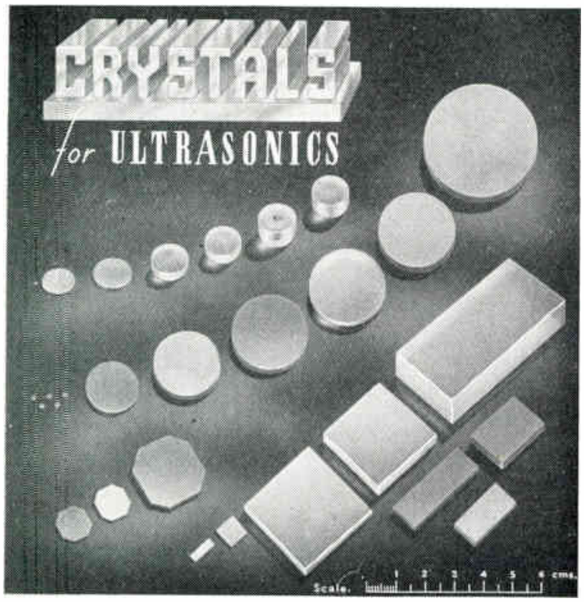
Capacitors by Wego

for
C W
V H F
RADAR
X-RAY
SMOOTHING
ENERGY STORAGE
IMPULSE GENERATORS
AND OTHER ELECTRONIC APPLICATIONS

Full information will be supplied on request.

WEGO CONDENSER CO. LTD.

BIDEFORD AVENUE, PERIVALE, GREENFORD, MIDDLESEX. Telephone : Perivale 4277.



Quartz Crystals of any shape and size cut and ground precisely to specification and coated, if required, with Gold, Silver, Aluminium or Rhodium, etc.

BROOKES CRYSTALS LTD.

Suppliers to Ministry of Supply, Home Office, etc.

10, STOCKWELL STREET, GREENWICH, LONDON, S.E.10
Phone: Greenwich 1828
Grams: Xtals Green, London. *Cables:* Xtals, London

★ ★ ASSOCIATED ILIFFE TECHNICAL BOOKS ★ ★

SOUND RECORDING AND REPRODUCTION

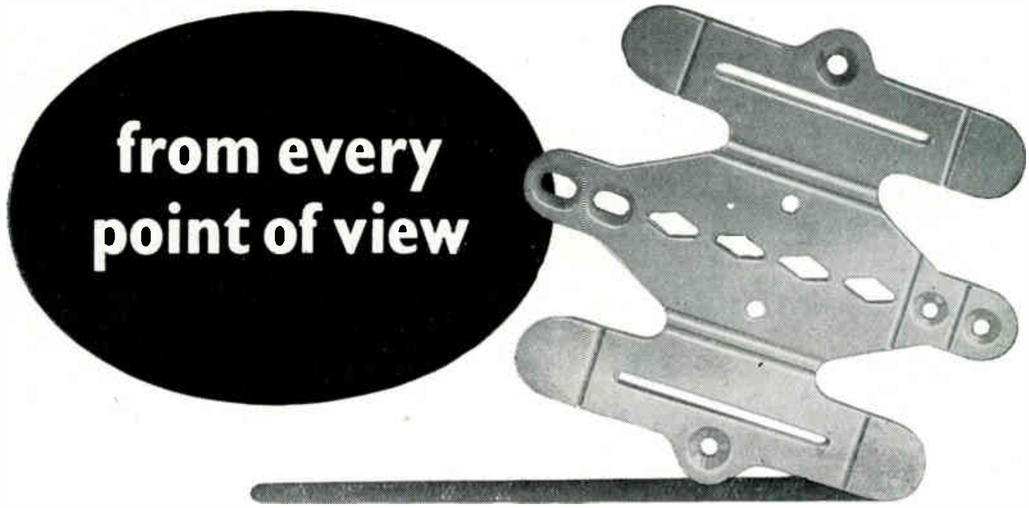
By J. W. Godfrey and S. W. Amos, B.SC., A.M.I.E.E., in collaboration with the B.B.C. Engineering Division. Will prove invaluable to all interested in the technique of high-quality sound recording. It covers in detail the theory and practice of disc, magnetic and film recording, special reference being made to equipment used by the B.B.C. 30*s. net.* *By post* 30*s.* 8*d.*

ADVANCED THEORY OF WAVEGUIDES

By L. Lewin. Sets out various methods of treating waveguide problems—the author selecting for discussion a number of topics as representative of the field in which the centimetre-wave engineer is now engaged. Many of these examples are concerned with the rectangular waveguide, but the reader will be able to apply the general principle to most of the cases encountered in practice.

30*s. net.* *By post* 30*s.* 7*d.*

Obtainable from all booksellers or by post from :—
DORSET HOUSE, STAMFORD ST., LONDON, S.E.1


from every
point of view

APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or the employer is excepted from the provisions of The Notification of Vacancies Order, 1952.

The Universal AVO MINOR

Precision ELECTRICAL TESTING INSTRUMENTS

A dependably accurate instrument is indispensable for testing and rapid fault location. For economy of time and expense, a multi-range instrument is to be recommended. Where compactness is desirable in the size of the meter, no more suitable instrument is available.

A small but highly accurate instrument for measuring A.C. and D.C. voltage, D.C. current, and also resistance. It provides 22 ranges of readings on a 3-inch scale, the required range being selected by plugging the leads supplied into appropriately marked sockets. An accurate moving-coil movement is employed, and the total resistance of the meter is 200,000 ohms.

The instrument is self-contained for resistance measurements up to 20,000 ohms and, by using an external source of voltage, resistance ranges can be extended up to 10 megohms. The ohms compensator for incorrect voltage works on all ranges. The instrument is suitable for use as an output meter when the A.C. voltage ranges are being used.

D.C. Voltage	A.C. Voltage
0—75 millivolts	0—5 volts
0—5 volts	0—25 ,,
0—25 ,,	0—100 ,,
0—100 ,,	0—250 ,,
0—250 ,,	0—500 ,,
0—500 ,,	
D.C. Current	**Resistance**
0—2.5 milliamps	0—20,000 ohms
0—5 ,,	0—100,000 ,,
0—25 ,,	0—500,000 ,,
0—100 ,,	0—2 Megohms
0—500 ,,	0—5 ,,
	0—10 ,,

Size: $4\frac{5}{8}$*in.* × $3\frac{7}{8}$*in.* × $1\frac{7}{8}$*in.*
Nett weight: 18 ozs.

£10 : 10 : 0

Complete with leads, interchangeable prods and crocodile clips, and instruction book.

Fully descriptive leaflet available from the sole Proprietors and Manufacturers:

The AUTOMATIC COIL WINDER & ELECTRICAL EQUIPMENT CO. LTD.
WINDER HOUSE • DOUGLAS STREET • LONDON • S.W.1 *Telephone: VICTORIA 3404/9*

On STAND 13 MUREX LTD will be showing

R.E.C.M.F EXHIBITION
GROSVENOR HOUSE • PARK LANE • W1
APRIL 14—16 1953

'SINCOMAX' SINTERED COMPOSITE MAGNETS

SINTERED PERMANENT MAGNETS • *Also* TUNGSTEN MOLYBDENUM, TANTALUM & ZIRCONIUM METALS

. . . and fabricated instrument & valve components

MUREX LIMITED *(Powder Metallurgy Division)* RAINHAM • ESSEX • *Telephone: Rainham, Essex 3322*
LONDON SALES OFFICE: CENTRAL HOUSE, UPPER WOBURN PLACE, W.C.1. *Telephone: EUSton 8265*

Index to Advertisers

Printed in Great Britain for the Publishers, Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1, at The Baynard Press by Sanders Phillips & Co., Ltd., Chryssell Road, London, S.W.9.

FOCUS ON LEADERSHIP

"LECTROPACKS"

Safeguarding the TV serviceman

The Service engineer depends upon the reliability of the components he uses to safeguard his own reputation: he must be certain they will give trouble-free service. The fact that so many Service men use only T.C.C. Condensers is evidence of their supremacy.

T.C.C. "Lectropack" Dry Electrolytic Condensers are robust yet compact and employ ALL-ALUMINIUM non-corrosive internal construction. The range below is a useful guide to the types available, which in the majority of cases will be found the exact replacement.

Capacity µF.	D.C. Volts		Ripple Current Max. M A	Dimensions in inches		T.C.C. Type No.	List Price
	Wkg.	Surge		L	D		
60 - 100	275	325	450	$4\frac{1}{2}$	$1\frac{3}{8}$	CE 37 HE	16/-
60 - 250	,,	,,	530	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 HE	28/-
100 - 200	,,	,,	650	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 HEA	28/-
100	350	400	450	$2\frac{3}{4}$	$1\frac{3}{8}$	CE 10 LE	13/6
200	,,	,,	770	$4\frac{1}{2}$	$1\frac{1}{2}$	CE 36 LE	24/-
60 - 100	,,	,,	450	$4\frac{1}{2}$	$1\frac{3}{8}$	CE 37 LEA	21/-
60 - 100	,,	,,	500	$4\frac{1}{2}$	$1\frac{1}{2}$	CE 36 LEB	23/-
64 - 120	,,	,,	530	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 LE	24/-
60 - 250	,,	,,	500	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 LEB	34/-
100 - 100	,,	,,	550	$4\frac{1}{2}$	$1\frac{1}{2}$	CE 36 LEA	26/-
100 - 200	,,	,,	700	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 LEA	33/-
100 - 200	,,	,,	700	$4\frac{1}{2}$	2	CE 15 LE	34/-
60	450	550	450	$3\frac{1}{4}$	$1\frac{3}{8}$	CE 38 PE	14/-
60 - 100	,,	,,	500	$4\frac{1}{2}$	$1\frac{3}{4}$	CE 60 PE	29/-

T.C.C. TECHNICAL BULLETIN No. 28 IS NOW AVAILABLE
detailing full ranges of TV condensers

SPECIALISTS
IN
CONDENSERS
SINCE 1906

THE TELEGRAPH CONDENSER CO. LTD

RADIO DIVISION

NORTH ACTON · LONDON · W.3 · Telephone: ACORN 0061