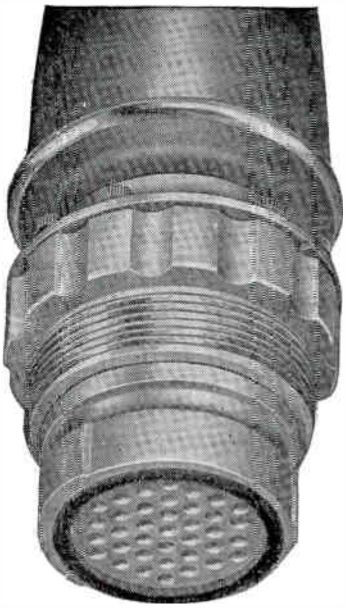
ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Three shillings and sixpence

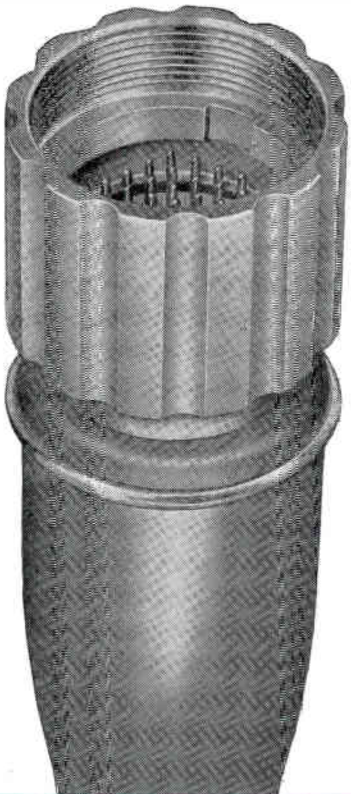
FEBRUARY 1959 Vol 36 *new series* No 2



37-way moulded-on polypole coupler.

BICC

'custom-built'

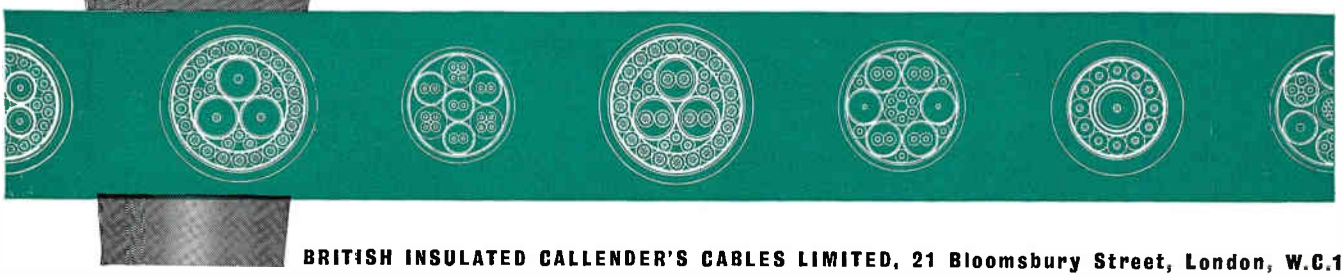


POLYPOLE COUPLER SYSTEMS

BICC design and manufacture to customers' specific requirements polypole coupler systems incorporating flexible multicore cables. These systems cover a wide variety of sizes and types of cable each terminated with robust moulded-on couplers.

BICC polypole coupler systems are particularly suitable for the reliable outdoor interconnection of mobile radio or electronic control equipment.

Further information is contained in Publication No. TD TPC 5—available on request.



BRITISH INSULATED CALLENDER'S CABLES LIMITED, 21 Bloomsbury Street, London, W.C.1

Now!

d.c. to V.H.F.

The TYPE A75

a range of WIDE BAND ATTENUATORS

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–to be sure!

FOUR MODELS

★ **TYPE A75** Range 99dB in 1dB steps. Input and Output Impedance 75 ohms. Price **£19** net in U.K.

★ **TYPE A76** Range 90dB in 10dB steps. Input and Output Impedance 75 ohms. Price **£8 10s.** net in U.K.

★ **TYPE A79** range 99dB in 1dB steps. Input and Output Impedance 75 ohms. Price **£22 10s.** net in U.K.

★ **TYPE A84** Range 99dB in 1dB steps. Input and Output Impedance 75 ohms. Price **£17 10s.** net in U.K.

TYPE A94 *Type A76 supplied less resistors for customers to fit their own network.* Price **£7 10s.** net in U.K.

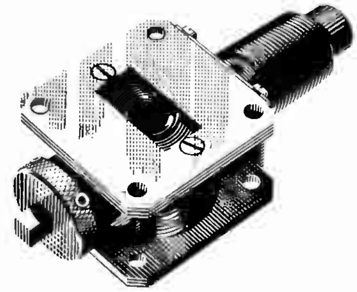
To our comprehensive range of attenuators, we announce the addition of four new models.
These attenuators employ resistive ladder network and provide accurate attenuation over a wide frequency band.
They are small compact units, ideal for building into customers' own equipment or for bench use in the laboratory.
Full details are given in leaflet No. R57 gladly sent on request.

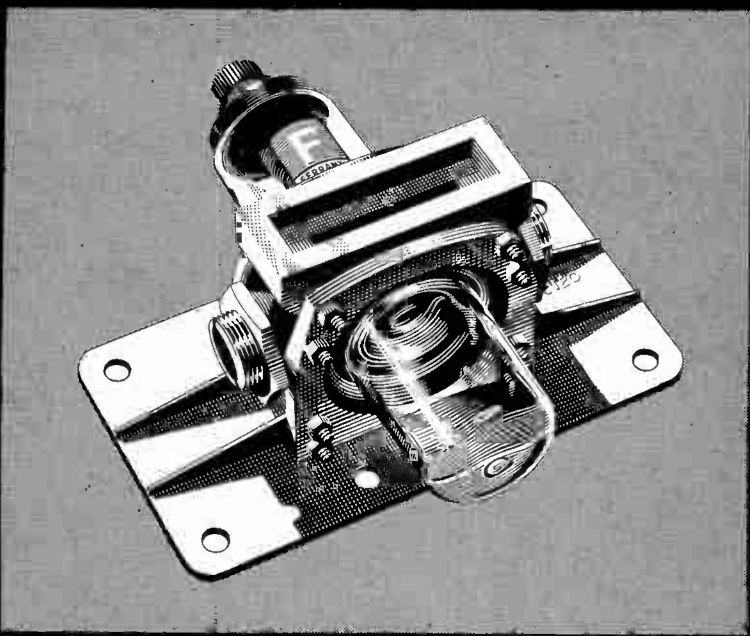
GD 68/A **Advance** COMPONENTS LIMITED · ROEBUCK ROAD · HAINAULT · ILFORD · ESSEX · TELEPHONE : HAINAULT 4444

FERRANTI

T. R. CELLS

for Radar Equipment





Ferranti Ltd. first made T. R. Cells in 1942 and by the end of the Second World War production was running at about 100,000 a year. The skill and experience attained during that period steadily increased until to-day the T. R. Cell research, development and production facilities of Ferranti Ltd. in Scotland are among the foremost in the world. The advice and co-operation of a highly skilled team of engineers is extended to all radar manufacturers.

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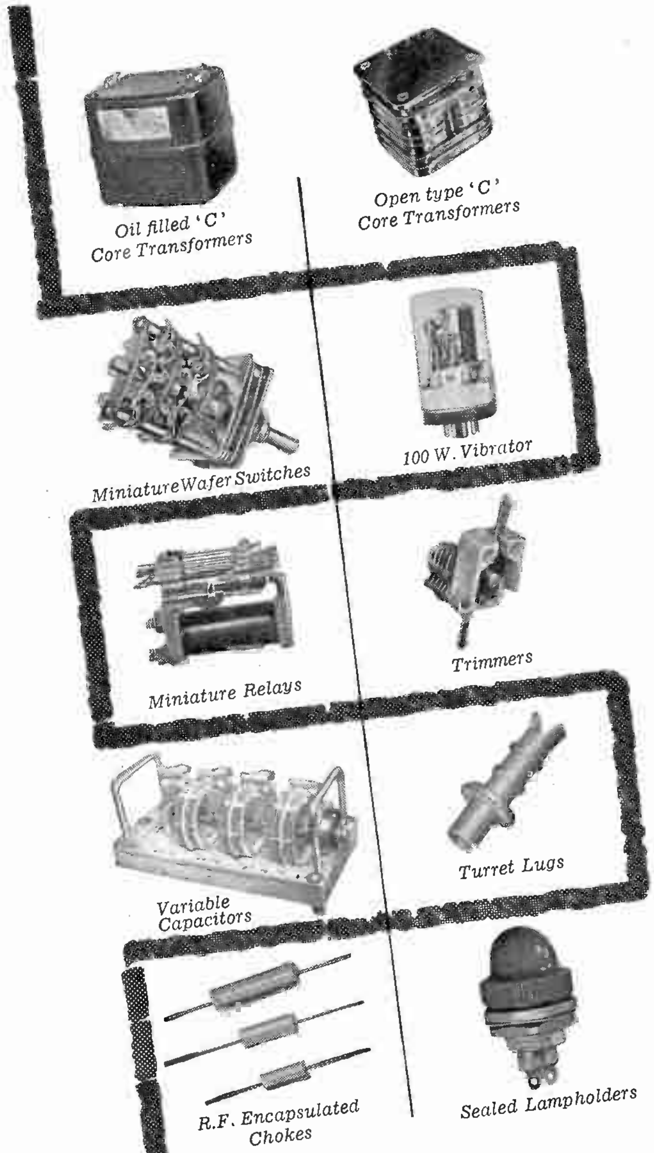
ES/T49



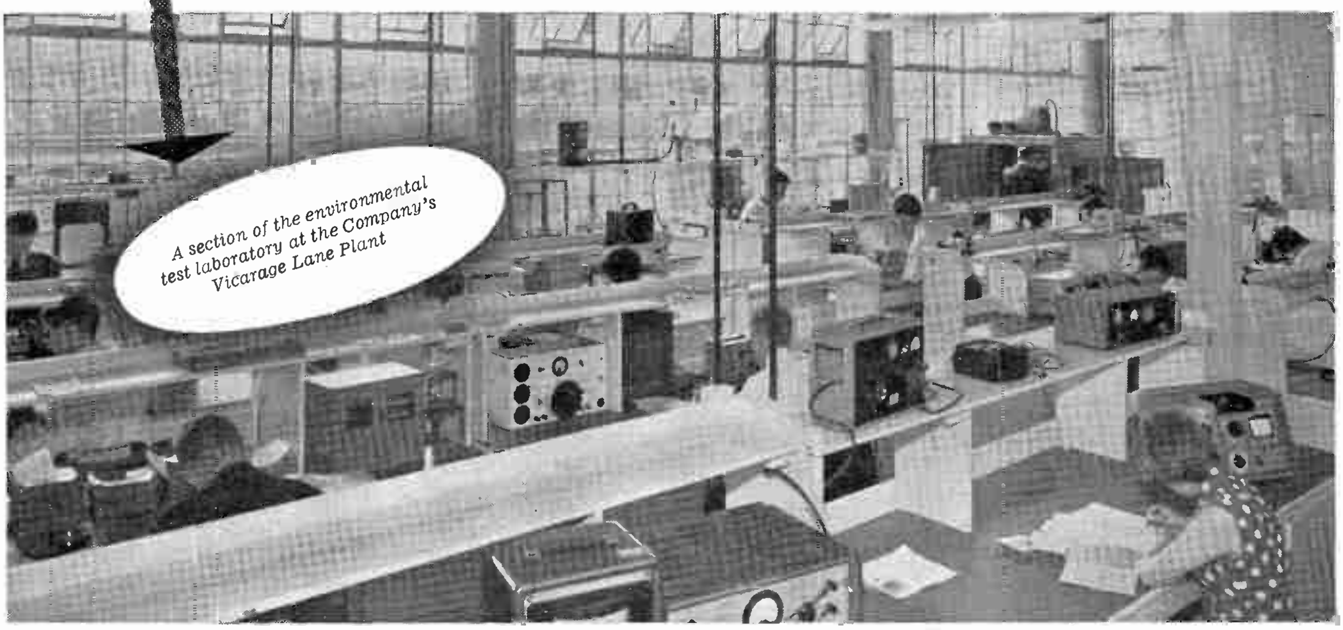
A lot depends on

Plessey

In its various factories throughout the Country, Plessey produces an immense range of Type Approved electronic components. The Company supplies much of the material for vital communications equipment and major research projects, which can range from human voice synthesis to rockets.

Modern electronics technique demands a degree of precision and accuracy rivalled in few other fields—with brains and manufacturing facilities to match. All this is aimed at utter reliability; no component must fail in service. Plessey are proud to be making such a contribution to an exacting science.

Plessey Type Approved components are available to Industry in many forms. Some are illustrated here. To give more detailed information a series of publications exists, and will gladly be sent on receipt of a letter stating your interests.



COMPONENTS GROUP | THE PLESSEY COMPANY LIMITED · ILFORD · ESSEX

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CP2

You can nearly always find a screw in the vast Unbrako range calculated to do just what you want better than any other screw.

But modern developments sometimes call for special screws not even standard to Unbrako.

When that happens our highly trained team of fastener-minded experts really get enthusiastic, responding to the challenge. They like to co-operate with you at the blueprint stage for preference, helping to design the perfect screw for the job, or they will simply make the screw to your specification, just about as well as a screw can be made.

So, standard or special, you can always safely specify Unbrako, the people who offer the most comprehensive specialised screw service in the world.

To be on the safe side, better get in touch with Unbrako over any fastener question. Remember what they say — Unbrako screws cost less than trouble.

There are two Unbrako lists you should have in your library, and a postcard or telephone call will bring them by return. They form a detailed and comprehensive guide to the whole Unbrako range, standard and non-standard, and no progressive firm should be without them.

Unbrako screws cost less than trouble

UNBRAKO

UNBRAKO SOCKET SCREW COMPANY LIMITED · COVENTRY

ALPHASIL—*the modern core material*

The inset curves illustrate the superior magnetic properties of Alphasil cold-reduced grain oriented silicon steel over those of a typical hot-rolled grade (Ferrosil 80). Alphasil has a maximum permeability four times that of the hot-rolled transformer sheet and its core losses are approximately one-third. Initial and incremental permeability, stacking factor and ductility are considerably better than those of hot-rolled sheet.

Alphasil ·013″ thick is produced in coil 30 inches wide, and can be supplied slit to narrower widths by arrangement.

TABLE OF WATTS LOSSES

	Frequency cycles/second	Guaranteed max. total losses at B. Max. 15 Kilogauss.
ALPHASIL 44	50	·62 watts/lb.
ALPHASIL 40	50	·56 watts/lb.
ALPHASIL 37	50	·51 watts/lb.
ALPHASIL 33	50	·46 watts/lb.

ABOVE
A 4,000-lb. coil of 30″ wide x ·013″ thick, ready for despatch.

RIGHT
Core-loss testing of Alphasil by the 'double-lap' Epstein method.

Thin Alphasil for high frequency work is also available in coil in ·004″ thick in widths up to $5\frac{1}{2}$ inches, and in ·002″ thick, in widths up to $4\frac{1}{2}$ inches.

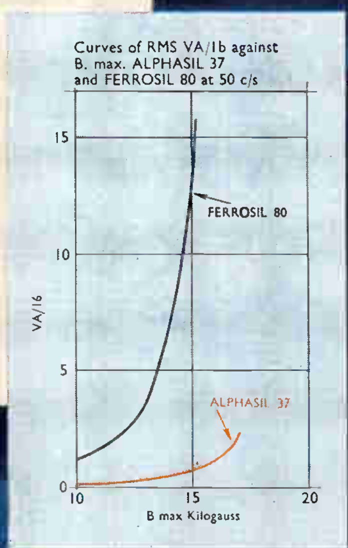
	Frequency cycles/second	Guaranteed max. total losses
ALPHASIL ·004HF	400	8·00 watts/lb. at B. Max 15 kilogauss
ALPHASIL ·002HF	8,000	9·50 watts/lb. at B. Max 2 kilogauss

Full technical data will be supplied on request

RICHARD THOMAS & BALDWINS LTD

Lamination Works: COOKLEY WORKS, BRIERLEY HILL, STAFFS.
Midland Section Office: WILDEN, STOURPORT-ON-SEVERN, WORCS.
Head Office: 47 PARK STREET, LONDON, W.1.

Our Cookley Works is one of the largest in Europe specializing in the manufacture of Stampings and Laminations for the electrical industry.

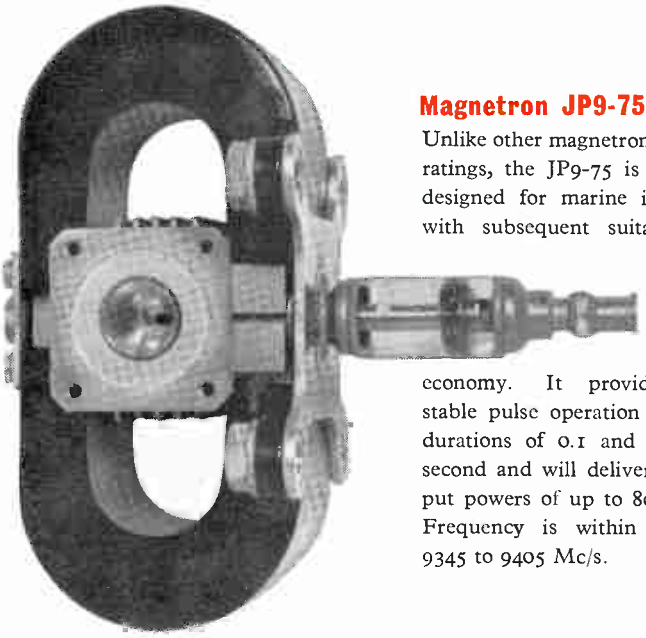


For large sets . . .

A line-up of large radar valves specially designed to provide high performance and reliability at a very competitive price.

Valves for Marine

80 kW



Magnetron JP9-75

Unlike other magnetrons of similar ratings, the JP9-75 is specifically designed for marine installations with subsequent suitability and economy. It provides highly stable pulse operation with pulse durations of 0.1 and 1.0 micro-second and will deliver peak output powers of up to 80 kilowatts. Frequency is within the range 9345 to 9405 Mc/s.

Pulse Modulator XH8-100 (CV1787)

Hydrogen-filled pulse modulator with a peak cathode current of 90 amps. Directly equivalent to the American 4C35.

Klystron KS9-20A (CV2792)

Mechanically tuned reflex klystron directly equivalent to the American 2K25. Frequency range 8500 to 9660 Mc/s.

Pulse Modulator QV20-P18 (CV2752)

Vacuum pulse modulator with a peak anode current of 18 amps. Directly equivalent to the American 4PR60A.

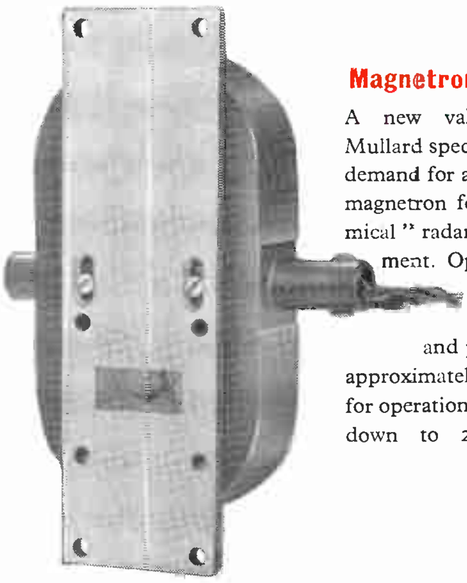
The valves in this advertisement are not shown to scale.

Radar

For small sets . . .

This Mullard line-up of radar valves is designed to achieve the maximum economies in power consumption and in the cost of associated components—both essential requirements in small ship installations.

3 kW



Magnetron JP9-2.5

A new valve designed by Mullard specifically to meet the demand for a low power pulsed magnetron for use in " economical " radar navigation equipment. Operating frequency within the range 9345 to 9475 Mc/s and peak output power approximately 3 kW. Tested for operation with pulse lengths down to 20 millimicrosecs.

Pulse Modulator EL360

Another Mullard development. This vacuum pulse modulator is recommended for operation with the JP9-2.5 magnetron. Maximum peak cathode current 4.0 amps.

Klystron KS9-20 (CV1795)

Mechanically tuned reflex klystron directly equivalent to the American 723A/B. Frequency range 8702 to 9548 Mc/s.

Abridged guide to Mullard Valves for Radar

Please write to the address below for further information.

	Mullard No.	Services No.	American No.
Magnetrons	JP8-02	—	—
	JP9-01	—	—
	JP9-2.5	—	7028
	JP9-7	CV3676	2J42
	JP9-7A	CV370	—
	JP9-7B	—	—
	JP9-7D	CV1866	—
	JP9-15	CV3997	2J42A
	JP9-75	—	6972
	JP9-80	CV3569	4J52
	JP9-80A	CV5018	4J52A
	JP9-250	CV2284	4J50
	JP9-250A	CV3953	4J78
	JP16-40	—	—
	JPT6-01	—	—
	JPT8-01	—	—
	JPT9-01	CV2420	—
	JPT9-02	CV2421	—
	JPT9-60	CV3560	2J51A
Hydrogen-filled Pulse Modulators	XH3-045	CV372	3C45
	XH8-100	CV1787	6268 4C35
	XH16-200	CV2520	6279 5C22
	XH25-500	CV3521	5949
Vacuum Pulse Modulators	EL360	—	—
	QQV5-P10	CV2295 CV3599	3E29
	QV10-P8	—	—
	QV20-P18	CV2752	4PR60A
	QYS50-P40	CV313	—
Klystrons	KS7-85	—	2K26
	KS7-85A	—	5976
	KS9-20	CV1795	723A/B
	KS9-20A	CV2792	2K25
	KT9-150W	—	—

Mullard

GOVERNMENT AND INDUSTRIAL VALVE DIVISION

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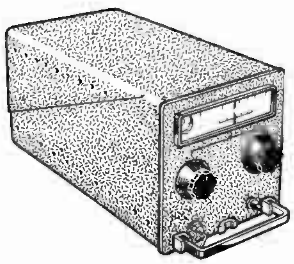
MVT 354


OPERATIONAL RELIABILITY
RADIO PRODUCTS
EDDYSTONE
TRADE MARK
MODEL 770R
EDDYSTONE
VHF & UHF
(A M & F M)
Communications
Receivers
Model 770R.
19-165 Mc/s.
Model 770U.
150-500 Mc/s.
Agents in all parts of the world
Please write for full Technical Specifications to the Manufacturers
STRATTON & CO. LTD., BIRMINGHAM, 31

A.W.A ELECTRONICS

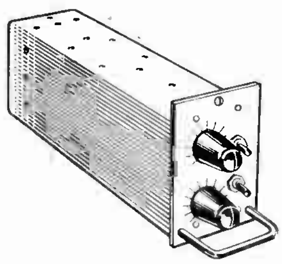
New concepts in electronics have been developed at AWA, as a result of experience with missile systems. Now they have a wider application. Here are some of the new AWA devices now available to industry.



U.H.F. WIDEBAND RECEIVER

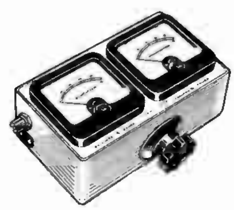
Basic arrangement consists of R.F. amplifier, mixer, local oscillator, I.F. amplifier (A.G.C. controlled), cathode follower output stage. Tuning indicator (EM 34) is also fitted to receiver. The standard forms: one for airborne racking with special separate power supply unit, the other on larger chassis including power supply unit (conventional 19″ front panel). *Standard specification: 420-470 M/cs frequency range; 4 M/cs overall bandwidth, approximately 10 db noise factor; approximately 70 ohms input impedance. 200-250 V and 50-60 c/s input supply. Input is unbalanced, output is via low impedance (cathode follower) stage.*

TRANSISTOR GALVANOMETER AMPLIFIER

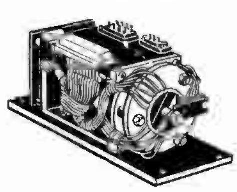


This Amplifier has been designed to drive viscous damped recording galvanometers which normally have a resistance of 50 ohms and a working range of DC to 2 Kc/s in frequency. The amplifier has a switched attenuator at its input and will accept single ended or push pull signals from ± 1 Millivolt to ± 500 volts and will feed a maximum of ± 50 Milliamps to the galvanometer. There is also a range of ancillary units available for use with this Amplifier as part of a comprehensive instrumentation system. *Standard specification: Dimensions: 4¼ in. x 3¾ in. x 10 in.; Frequency response: Flat from DC to 2 Kc/s, 5% down at 3 Kc/s, 3db down at 6 Kc/s; Noise Level: Less than 10 Microvolts; Input impedance: 40,000 ohms on range 5, 110,000 ohms all other ranges; Gain: Maximum 7.5 Milliamps/Millivolt, minimum 0.04 Milliamps/Volt; Power requirements: ± 6 Volts DC 220 Milliamps each line.*

DIRECTIONAL COUPLER



Of the 'Loop' type, suitable for measurements of RF power and Standing Wave Ratio in coaxial cables. Directional properties are largely unaffected by frequency changes, so coupler may be used to help obtain optimum termination of a 52 ohm coaxial system up to 600 M/cs. *Standard specification: Size 7″ x 4″ x 2½″; weighs 4 lbs. 3 ozs.; Power Measurement Range is Low range 1 w.cw.max. High range 5 w.cw.max.; less than 1% attenuation; better than 2% accuracy at frequency of calibration.*



ROTARY SWITCH FOR TELEMETRY

Based on a conception of British Ministry of Supply's Research and Development Establishment, gives facilities previously unobtainable from mechanical sampling devices. The Standard Model enables two 24 channel banks to be sampled at speeds up to 200 r.p.s.

All devices are adaptable to suit customers' own requirements. For further information consult:

COMMERCIAL ELECTRONICS DEPT.

SIR W. G. ARMSTRONG WHITWORTH AIRCRAFT LTD.,
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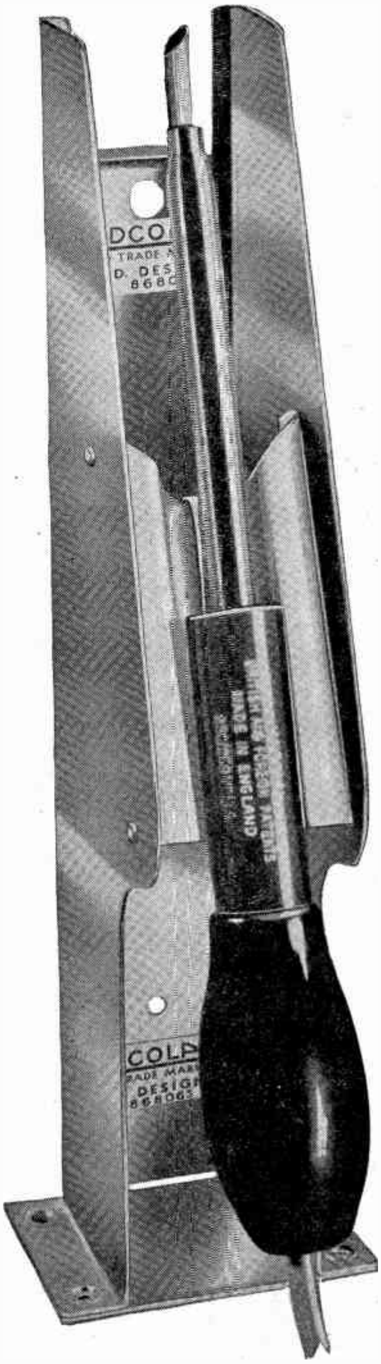
ADCOLA
(Regd. Trade Mark)

Soldering Instruments and Equipment



Comprehensive Range of Models

P.V.C. Cable Strippers

Solder Dipping Pots

Supplied in ALL VOLT RANGES

A PRODUCT FOR PRODUCTION

RADIO, TV
RADAR
ELECTRONICS
TELECOMMUNICATIONS
ETC.

(Illustrated)

Protective Shield
List No. 68

$\frac{3}{16}$" Detachable
Bit Model
List No. 64

Traditional British Quality and Workmanship

ADCOLA PRODUCTS LIMITED

Head Office, Sales and Service:
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LONDON, S.W.4 Telephones MACAULAY 3101 & 4272

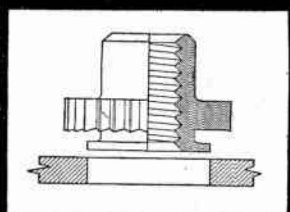
Improve your production and your product

Use the new, sheet metal fastening THE ROSAN PRESS NUT. This nut provides a quick and secure deep tapped hole in metal sheet or plate—and it is cheaper. Only a year after its introduction it is being widely used in industry.

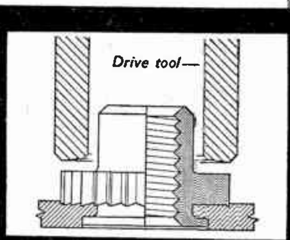
Famous concerns using the Press Nut include:

Keith Blackman Ltd.
Chubb & Sons Lock & Safe Co. Ltd.
Dynatron Radio Ltd.
Elliott Brothers (London) Ltd.
Igranic Electric Co. Ltd.
Claude Lyons Ltd.
Marconi's Wireless Telegraph Co. Ltd.
The Plessey Co. Ltd.
Pye Telecommunications Ltd.
Standard Telephones & Cables Ltd.
Vickers-Armstrongs (Engineers) Ltd.
Westinghouse Brake & Signal Co. Ltd.
and many others.

HOW IT WORKS



Drill or punch hole in sheet metal to specified diameter. Insert smaller flange into hole. Apply impact force to exposed edge of serrated flange of Press-Nut.



Note that impact flows sheet metal into space between flanges, locking the Press-Nut securely in place. Teeth of the serrated flange broach themselves into the parent metal preventing rotation.

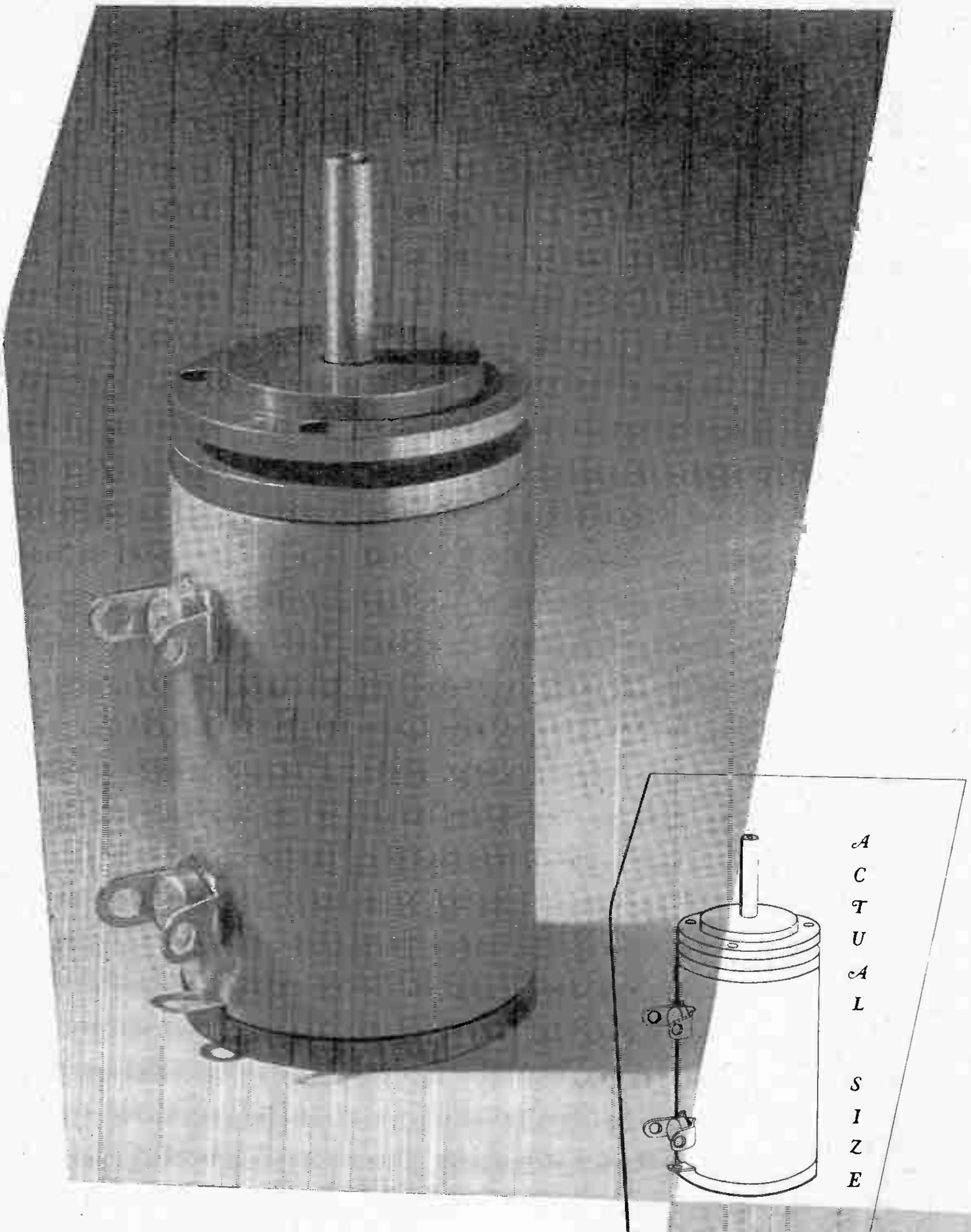
If you do not know about it write today for a descriptive leaflet and samples, without cost or obligation to:—

INSTRUMENT SCREW CO. LTD., Northolt Rd., South Harrow, Middx. Tel: BYRON 1141

Pat. App. No. 19826/57

ACTUAL SIZE OF A 4 BA. NUT

TAS/IS 6



BECKMAN Instruments Limited

Series 'AJ' HELIPOT

The Beckman Helipot Series AJ 10 turn helical potentiometer is manufactured by Beckman Instruments Limited at their new factory in Scotland. This is a miniaturised potentiometer and is available with bush mounting, servo mounting sleeve type bearing, and also servo mounting with miniaturised ball-bearings.

Full details available from :—

WINSTON ELECTRONICS LIMITED

GOVETT AVENUE
SHEPPERTON
MIDDLESEX

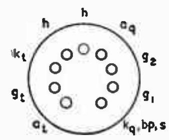
TELEPHONE: WALTON-ON-THAMES 6321

Introducing another outstanding Ediswan Mazda Valve type 30FL1

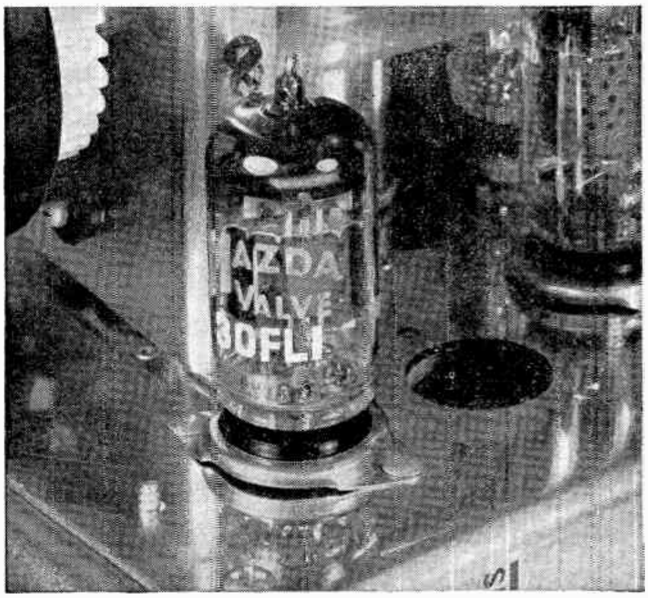
For the information of set designers we are publishing details of individual 0.3 amp heater valves in our "First Preference" Range for TV circuits. If you are a TV manufacturer we shall be pleased to supply full technical details of our "First Preference" Range, together with a set of valves for testing, on receipt of your enquiry. The valve dealt with here is the Type 30FL1, a general purpose triode-tetrode for ac/dc mains television receivers.

Maximum overall length (mm)		56.0
Maximum seated height (mm)		49.0
Maximum diameter (mm) ..		22.2
Heater current (amps) ..	I_h	0.3
Heater voltage (volts) ..	V_h	9.4

B9A BASE



View of free end



MAXIMUM DESIGN CENTRE RATINGS

		Tetrode	Triode
Anode voltage (volts)	$V_{a(max)}$	250	250
Screen voltage (volts)	$V_{g2(max)}$	250	—
Anode dissipation (watts)	$P_{a(max)}$	3	2
Screen dissipation (watts)	$P_{g2(max)}$	1	—
Heater to cathode voltage* (volts r.m.s.)	$V_{h-k(max)}$	150	—

*Measured with respect to the higher potential heater pin.

CHARACTERISTICS

		Tetrode	Triode
Anode voltage (volts)	V_a	170	200
Screen voltage (volts)	V_{g2}	170	—
Anode current (mA)	I_a	10	10
Mutual conductance (mA/V)	g_m	7.5	3.4
Amplification factor	μ	—	16

CAPACITANCES (pF) (without holder)

	Tetrode		Triode	
Grid/Earth	c_{g1-E}	7.9	c_{gt-E}	3.6
Anode/Earth	c_{aq-E}	3.2	c_{at-E}	2.6
Grid/Anode	c_{g1-aq}	0.03	c_{gt-a}	2.7

APPLICATION NOTES

The two sections of the Ediswan Mazda 30FL1 are completely independent. Both the internal shielding and the arrangement of the base connections result in a very low level of capacitive and electron coupling so that the sections may be used in diverse applications without their performance being significantly affected by interaction. The chart illustrates a variety of permissible combinations in a television receiver.

The triode sections of the Ediswan Mazda valves type 6/30L2, 30PL1 and 30FL1 have identical characteristics.

TRIODE SECTION	TETRODE SECTION Sync. Pulse Separator	*Video Amplifier	Line Scan Generator	Frame Scan Generator
Frame Sync. Clipper	A	A	B	A
Line Sync. Phase Splitter	A	A	A	B
Line Scan Phase Comparator	A	A	A	B
Line Scan Generator	A	A	A	A
Frame Scan Generator	A	A	A	A
Gated A.G.C.	A	A	A	B
*Video Cathode Follower	A	A	A	B

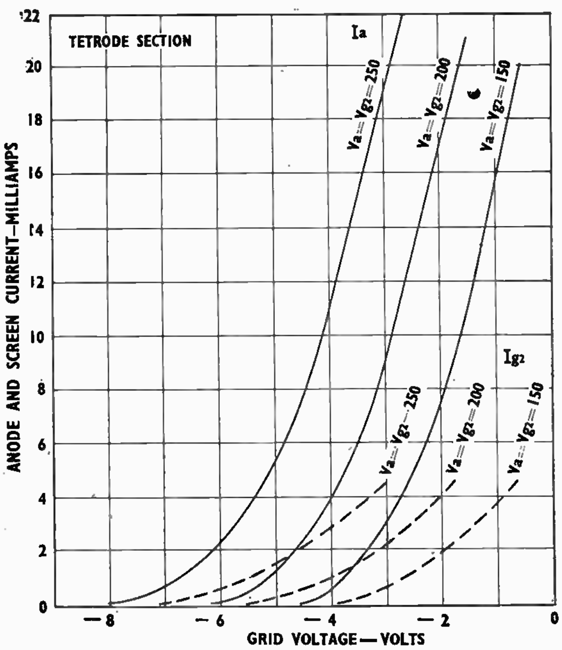
NOTES:— A—These combinations are satisfactory for the types of circuit and component values normally used.

B—In general, these combinations should be satisfactory but extra care may be needed to minimise stray coupling in wiring.

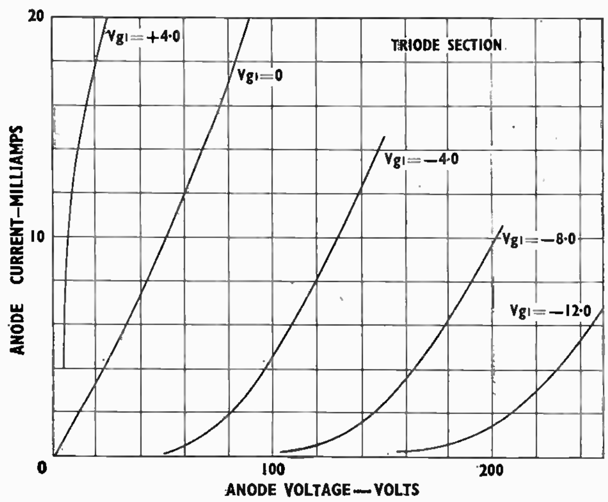
* With an H.T. line voltage of 190 V, the use of the tetrode to provide picture modulation to the cathode ray tube is only recommended with an anode load resistance not less than 10 kΩ. The triode section is not recommended as a cathode follower to provide picture modulation to the cathode ray tube unless the effective capacity loading is low enough to give adequate frequency response with a cathode resistance not less than 22 kΩ.

If the H.T. line voltage is higher than 190 V, adequate output under limit conditions can be obtained with lower resistance values.

Characteristic curves of average Ediswan Mazda Valve type 30FL1



Characteristic curves of average Ediswan Mazda Vale type 30FL1



SIEMENS EDISON SWAN LIMITED *An A.E.I. Company,*

Technical Service Department, 155 Charing Cross Rd., London, W.C.2

Telephone: GERrard 8660. *Telegrams:* Sieswan Westcent, London.

EDISWAN
MAZDA

CRC 15/34

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

HUGH S. POCOCK M.I.E.E. *Managing Editor*
W. T. COCKING M.I.E.E. *Editor*

CONTENTS

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Permanent Magnets | **Design Advisory Service**

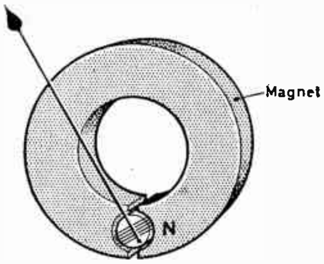
No. 11

Moving Coil Applications – 2

Advertisements in this series deal with general design considerations. If you require more specific information on the use of permanent magnets, please send your enquiry to the address below, mentioning the Design Advisory Service.

One of the most important applications for permanent magnets is in moving coil indicating and recording instruments.

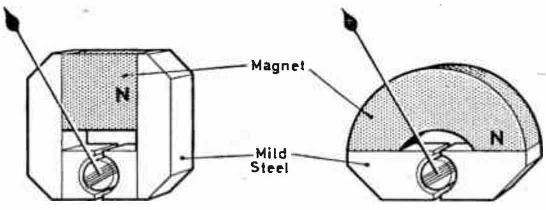
The majority of electrical indicating instruments use a permanent magnet/moving coil assembly, and as a result of the introduction of new permanent magnet materials having extremely reliable and stable magnetic performance, these instruments can now be made with a compactness previously unobtainable. This type of instrument consists of a current carrying coil pivoted axially allowing free rotation in narrow air gaps energised by a permanent magnet, and controlled by a suitable return spring. The uniformity of scale divisions depends on the uniformity of the magnetic field in the air gaps.



(a)

Old type 35% Cobalt magnet moving coil system.

Prior to the commercial availability of modern high performance anisotropic magnets, moving coil instruments followed closely design (a). Designs (b) and (c) using 'Ticonal' magnets are now in general use.



(b) (c)

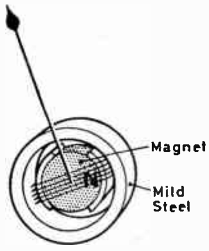
Typical arrangement of 'Ticonal' magnets used for moving coil indicating instruments.

Design (b) shows a sound, robust and simple mechanical construction with a magnetic efficiency of approximately 40% (i.e. 40% of the total flux produced by the magnet is usefully employed in the air gaps).

Design (c) is suitable for fast production methods but as curved magnets do not operate at (B_dH_d) max. throughout, they have to be slightly larger to obtain the same magnetic field as produced by design (b).

The extremely high coercivity of Mullard 'Ticonal' magnets have made possible meters designed as shown in (d) where the magnet takes the place of the usual mild steel central core and is surrounded by a mild steel yoke ring which acts as a return flux path and is also a most effective magnetic screen. Instruments of this construction can be made very much smaller than previous designs—an important factor in these times of miniaturisation and limited space. These instruments can be mounted close together and on steel panels without the danger of interference, interaction or loss of calibration accuracy.

From examination of design (d) it will be seen that small pole shoes are fitted to the magnet—these are essential if uniform scale divisions are required. Practically all the flux from the magnet becomes useful, *i.e. in the order of 80% efficiency.*



(d)

Design using a 'Ticonal' internal core magnet arrangement.

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MC272

VOLUME 36 NUMBER 2 **ELECTRONIC & RADIO ENGINEER**

FEBRUARY 1959 *incorporating WIRELESS ENGINEER*

Reliability and Redundancy

IT is well known that a certain amount of redundancy in a message greatly reduces the chances of errors in transmission causing misunderstanding. It is now being suggested that redundancy in electronic apparatus affords a better chance of improving its reliability than any practical improvements in the reliability of individual parts, and with much less engineering effort.

The idea of providing duplicate equipments as a safeguard against a breakdown in one is very old; e.g., the stand-by transmitter. The increase in reliability obtainable in this way is not enormous, however. The greatest increase of reliability is obtained when each individual part is duplicated in such a way that either part may fail without affecting the operation of the equipment as a whole. Simultaneous faults in each of such a pair of parts are then necessary to bring about an equipment failure.

Such duplication will usually be impracticable. It is not often possible simply to connect two parts in parallel; fault detectors actuating change-over mechanisms may be needed and there is an obvious practical limit to the number of such devices that can be employed.

A compromise between these two extremes, called group redundancy, is advocated by J. H. S. Chin ("Optimum Design for Reliability", 1958 I.R.E. Wescon Convention Record, Part 6, p. 23). In this, an equipment is divided into a number of tandem-connected groups and each group is provided with its duplicate. In one of his examples, the author considers an equipment of 500 elements, failure of any one of which will put the equipment out of action. Without redundancy, the reliability is 50% after a certain operating time. Dividing it into five tandem groups of 100 elements and duplicating each group increases the reliability to about 90% while, dividing it into 100 groups of 5 duplicated elements, brings it up to some 99%.

The increased amount of apparatus needed is, of course, a drawback but in these days of miniature techniques this is less important than it once would have been and the system appears an attractive one. The author does stress, however, the necessity for including redundancy in all stages of an equipment. To retain even one stage in a cascade without redundancy largely nullifies the redundancy of the other stages.

H.F. Exponential-Line Transformers

DESIGN AND CONSTRUCTION OF FOUR-WIRE TYPE AT RUGBY RADIO STATION

By S. G. Young, B.Sc.(Eng.), A.M.I.E.E.*

SUMMARY. *The mathematical theory of an exponential line is summarized and a practical method for designing such lines is described. The design and constructional details of the relatively simple exponential lines used at the Rugby 'B' Radio Station are given. These lines are of four-wire construction and match balanced impedances of* 195 *and* 565 *ohms over the frequency range* 4–27 *Mc/s.*

A large radio transmitting station, such as the new station recently completed at Rugby[1], which carries diverse traffic to most parts of the world, is necessarily equipped with a large number of transmitters and a far greater number of aerials. If aperiodic aerials are employed, the average number of aerials per transmitter can be kept down to a minimum of two or three, dependent upon the destinations and durations of the various services. Even so, a large degree of flexibility in transmitter to aerial connections is required in order to employ the transmitters economically. The achievement of this flexibility with safety and rapidity of operation presents a major problem in station design. The aerial feeders must be concentrated at one or more switching points within the station building without producing excessive crosstalk and with safety to operators. Coaxial feeders provide the most convenient solution to these problems.

On the other hand, the relatively long feeders from the building to the aerials are provided most economically in the form of open-wire lines. To secure the advantages of both types of feeder where they are most appropriate, aperiodic coupling devices capable of matching the impedances of coaxial and open-wire feeders are required. By adopting balanced twin coaxial feeders the coupling device is simplified, in that the need for a balance-to-unbalance transformation is obviated.

Aperiodic matching devices suitable for carrying high radio-frequency powers usually consist of tandem sections of line whose characteristic impedances conform to some designated progression in values. At the Rugby radio station extension, the exponential form of impedance variation was preferred for both mechanical and electrical reasons. The absence of abrupt impedance changes reduced the possibility of unwanted electrical discontinuities, and the required variation in impedance could be approximated by a series of linearly-tapered lines. The number of supports was therefore reduced, and the loads upon the wires and spreaders decreased. A more complicated form of construction than an open four-wire line was considered unwarranted, and the necessity to withstand icing restricted the lower limit of impedance to about 200 ohms.

This order of impedance is conveniently obtainable from twin coaxial feeders having a sheath-to-core diameter ratio of five. Such feeders were adopted for the internal distribution and switching, and have a measured characteristic impedance of 195 ohms. The external lines were constructed of twin 400-lb/mile (0·16 in. diameter) copper wires, spaced at nine inches, giving a characteristic impedance of 565 ohms, which conveniently matches the average input impedance of three-wire rhombic aerials.

The design methods described in the following section led to the development of a line, 135 feet long (with only one intermediate support) which matched these impedances over the full frequency range 4 to 27·5 Mc/s.

Summary of Mathematical Theory

The properties of a balanced transmission line having a characteristic impedance increasing or decreasing exponentially with distance along the line were mentioned in several patents in the 1920's[2,3,4]. Interest revived some ten years later in the United States[5,6] and, later, in Germany[7,8,9], but the exponential line has only recently been brought into general use, and for this reason its properties may not be so well appreciated. The most important properties, for practical design purposes, are therefore summarized in the next section. References 5, 7 and 8 supply a complete theoretical treatment.

The characteristic impedance Z_x of any small length of exponential line, distance x from the sending end is given by

$$Z_x = Ze^{qx} \qquad (1)$$

where q is a constant and Z is the characteristic impedance of the line at the sending end.

It follows that the characteristic impedance at the receiving end of the line of length l is

$$Z_l = Ze^{ql} \qquad (2)$$

The line behaves as a high-pass filter, the cut-off wavelength being given by

$$\lambda_c = 4\pi/q \qquad (3)$$

* *Post Office Engineering Department, Radio Planning and Provision Branch*

If at the receiving end, the line is terminated in an impedance Z_r such that

$$Z_r/Z_l = k \quad \ldots \quad (4)$$

then the impedance seen when looking into the line at any point distance x from the sending end is given by

$$Z_x' = Z_x \frac{\left[1+k\left(\frac{T+q/2}{p}\right)\right]-\left[1-k\left(\frac{T-q/2}{p}\right)\right]\cdot e^{-2T(l-x)}}{\left[k+\left(\frac{T-q/2}{p}\right)\right]-\left[k-\left(\frac{T+q/2}{p}\right)\right]\cdot e^{-2T(l-x)}} \quad \ldots \quad (5)$$

where $T = \sqrt{p^2+q^2/4} = p\sqrt{1-v^2}$, $v = j(q/2p)$, and $p = j\omega\sqrt{L_x C_x}$ is the propagation constant of any uniform line having the same distributed constants as the exponential line at the point x. For a non-dissipative air-spaced line $p = j2\pi/\lambda$ which is independent of spacing, and then

$$v = \lambda/\lambda_c = f_c/f \quad \ldots \quad (6)$$

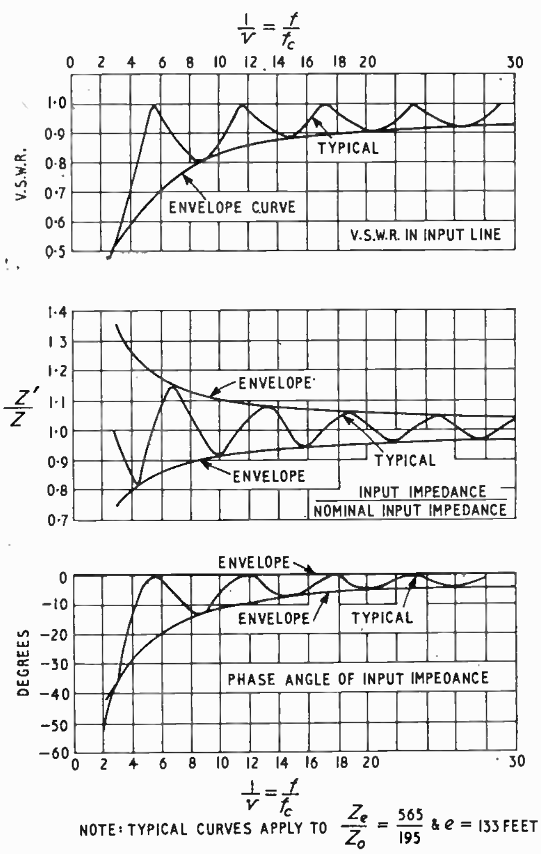
Equ. (5) may be re-written as

$$\frac{Z'x}{Z} = \frac{[1+k(\sqrt{1-v^2}-jv)]-[1-k(\sqrt{1-v^2}+jv)]\cdot e^{-2T(l-x)}}{[k+\sqrt{1-v^2}+jv]-[k-\sqrt{1-v^2}+jv]\cdot e^{-2T(l-x)}} \quad \ldots \quad (7)$$

Let $k = 1$; i.e., the line is terminated in $R_L = Z_l$ (its characteristic impedance at that point). The input impedance at $x = 0$ is then given by

$$Z' = Z\cdot e^{-2jE}\cdot\left\{\frac{1+(\tan E)\cdot e^{j(2E+\frac{\pi}{2}-n)}}{1+(\tan E)\cdot e^{-j(2E+\frac{\pi}{2}-n)}}\right\} \quad \ldots \quad (8)$$

Fig. 1. *Exponential line; v.s.w.r., input impedance and phase angle when terminated in R_L*



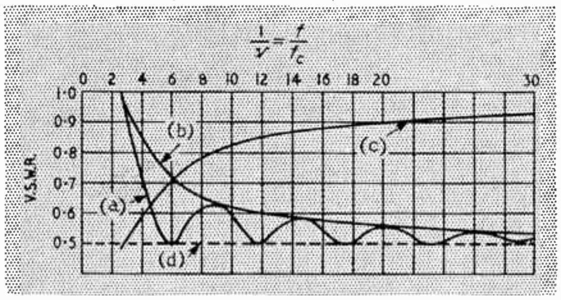


Fig. 2. *Exponential line; effect of incorrect termination on input impedance. (a) Permissible v.s.w.r. in output line for v.s.w.r. of* 0·5 *in input line, for* 195–565-Ω *transforming line* 133 *ft long; (b) envelope curve of (a); (c) envelope input v.s.w.r. when terminated in R_L; (d) curve (d) = (b) × (c). Note that v.s.w.r. in input line = v.s.w.r. in output line × v.s.w.r. when perfectly terminated*

$$\text{where } n = -2Tl \quad \ldots \quad (9)$$

$$\text{and } \sin 2E = v = f_c/f \quad \ldots \quad (10)$$

This shows that the value of input impedance oscillates about the nominal characteristic impedance at the input end between the limits given by

$$Z\,(1+f_c/f) \text{ and } Z/(1+f_c/f) \quad \ldots \quad (11)$$

The greatest deviations in impedances occur when the line is approximately an odd number of eighth wavelengths long and for these values the phase angle of the input impedance is equal to $(-2E)$. The line acts as a perfect transformer when it is an even number of quarter wavelengths long, and it then has a resistive input impedance equal to Z. When it is an odd number of quarter wavelengths long the input impedance is again equal to Z, but the phase angle than has its maximum deviation and is equal to $(-4E)$. The phase angle is always negative, and varies between the limits of 0 and $(-4E)$. The worst standing-wave ratio in a uniform line connected to the input end of the exponential line will be experienced when the repeated line is approximately an odd number of quarter wavelengths long, and its value can be calculated from the fact that the input impedance of the exponential line is then given by

$$Z'' = Z\,.\,e^{-j4E} \quad \ldots \quad (12)$$

This equation may be employed to draw an 'envelope' curve of the worst possible voltage-standing-wave ratios for a line, the variable being v; as indicated in Equ. (10). The particular frequencies at which peaks of impedance and of v.s.w.r. occur will depend upon the length of the exponential line and its transformation ratio. The input impedance, phase angle, and v.s.w.r. 'envelope' curves for any exponential line are given in Fig. 1, together with values for a typical line (of the type used at Rugby).

In practice, the terminating impedance of the exponential line will vary within certain limits. This variation in impedance can be conveniently expressed in the terms of the v.s.w.r. that would cause such a variation to exist along a uniform line of characteristic impedance Z_l. It is necessary, therefore, to investigate the performance of an exponential line under such conditions, using Equ. (7). In Fig. 2 a typical curve is

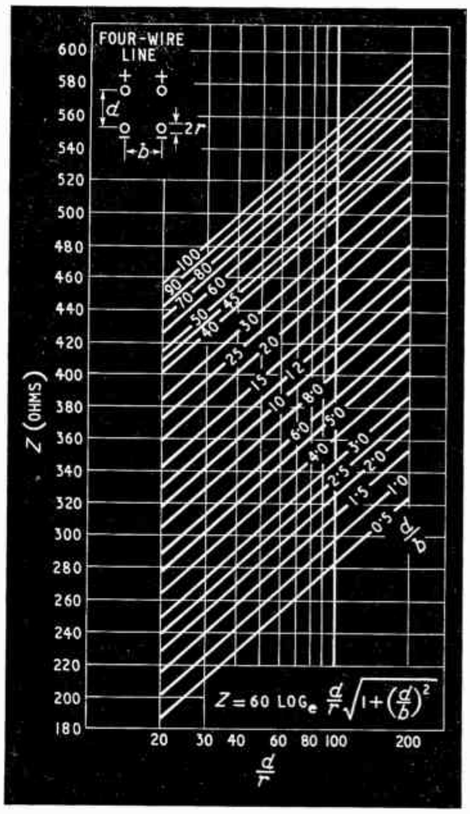


Fig. 3. *Characteristic impedance of four-wire line with adjacent wires commoned*

given of v.s.w.r. in the output line, for a limiting value of v.s.w.r. of 0·5 in the input line.

An 'envelope' curve may again be drawn. It can be seen that this curve, when multiplied by the envelope curve of v.s.w.r. for the case of a 'perfectly' terminated line (Fig. 1), produces a constant value of v.s.w.r. sensibly equal to the limiting value. This holds for other limiting values of input v.s.w.r. and, therefore, we may state that an exponential-line transformer increases the v.s.w.r. already existing in the output line by a factor, equal to its own v.s.w.r. when 'perfectly' terminated.

A Practical Method of Design

The achievement, in practice, of a line in which the characteristic impedance varies exponentially along its length, is a matter of some difficulty. The insertion of spacers at calculated points is not necessarily satisfactory because, if the number of spacers is small, the impedance variation between them may differ greatly from the exponential law and, if a large number are used, they throw additional lumped capacitance across the line, and so adversely affect its performance. Solutions have taken the form of plate lines, or lines of multiwire construction, both of which are expensive and difficult to build and maintain.

Christiansen[10] has shown that a four-wire line, with the two wires on each side connected in parallel, can produce a sensibly exponential taper of characteristic impedance, provided that the wire spacings in plan and elevation taper in opposite directions. (It does not follow that any line of this form is exponential). This is, obviously, a practical proposition, provided that the number of changes in (mechanical) taper, and hence the number of spacers, can be kept reasonably low, and that the lowest impedance is about 200 ohms.

The characteristic impedance of such a uniform four-wire line is given by

$$Z = 60.\log_e \left[\frac{d}{r}\sqrt{1+\left(\frac{d}{b}\right)^2}\right] \quad \ldots \ldots \ldots (13)$$

where r = wire radius, b = spacing between the parallel wires, and d = spacing between the pairs.

A chart for the determination of such impedances is given in Fig. 3.

For an exponential line, from Equ. (1),

$$\log_e Z_x = q\,x + \log_e Z \quad \ldots \ldots \ldots \ldots (14)$$

Thus $\log_e Z_x$ varies linearly with x. If the wire spacings taper linearly, then (d/r) and (b/r) also vary linearly with x. Hence $\log_e Z_x$ varies linearly with (d/r) and (b/r).

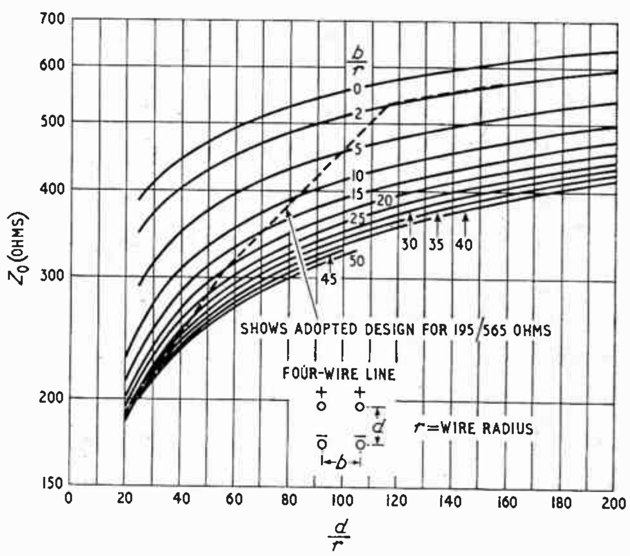
A convenient design chart is given in Fig. 4, calculated from Equ. (13) in which Z is plotted on a logarithmic scale against (d/r) for constant values of (b/r). An exponential four-wire line with uniform linear tapers of (d/r) and (b/r) will be represented by a straight line on this chart, with the (b/r) lines making equal intercepts upon it. Christiansen[10] employed this type of chart for designs between 350–600 ohms. As extended in Fig. 4 it is useful for preliminary design work, but it has not been found sufficiently accurate, and an extension of this method has been employed as follows.

First, convenient values of (d/r), (d/b) at the low impedance end of the line are selected. If the tapers are to be linear, then it follows that

$$\frac{db}{dx} = B, \text{ and } \frac{dd}{dx} = D \text{ where } D, B \text{ are constants.}$$

and hence $B/D = C$, another constant.

Fig. 4. *Exponential-line design chart (Christiansen's method). Notes:—straight taper in depth d is a straight line on graph, equal intercepts by b/r constant lines means a straight taper in b breadth. If line tapers uniformly in d (i.e., $d = k_1 + k_2 l$) then straight line represents a line $\log_e Z_l = ql$ or $Z_l = e^{ql}$*



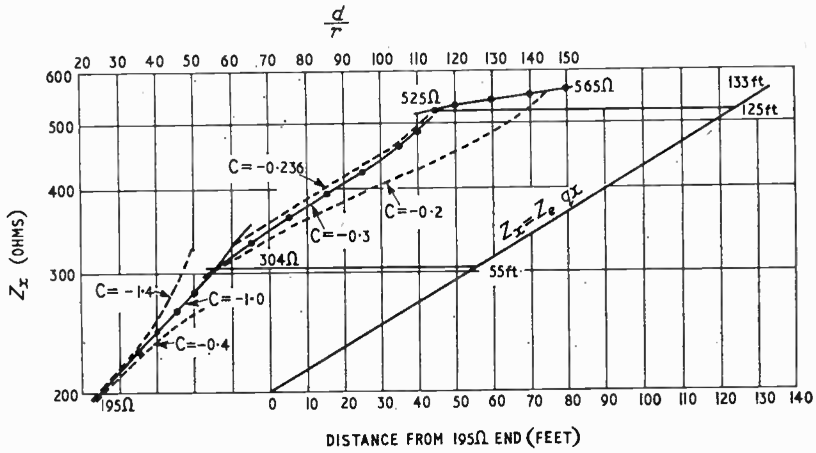


Fig. 5. *Design of 4-wire exponential line for* 195–565 Ω

For increasing values of (d/r), values of (b/r) and, hence (d/b) are calculated for a given value of C. The logarithms of the corresponding values of Z, found from Fig. 3, are then plotted against (d/r). The resultant graphs are not necessarily straight lines, and the value of C is adjusted until as wide a range of variation in $\log_e Z$ as is possible is achieved on a sensibly straight line. A new value of C is then chosen, commencing from the terminating values of (d/r) and (d/b), for the previous section, and a further satisfactory section is found. The distances down the line at which these changes of taper are to take place are found directly from a plot of Z versus distance down the line. The procedure is illustrated in the following section by a practical example.

The power-handling capacity of the line is limited by the possibility of corona and flash-over, the current-handling capacity of the wires and joints being, generally speaking, more than adequate. The maximum voltage gradient for a line of this four-wire form is given approximately by

$$\frac{dV}{dr} = \frac{V}{2r.\log_e\left[\frac{d}{r}\sqrt{1+\left(\frac{d}{b}\right)^2}\right]} \quad \ldots \quad \ldots (15)$$

were V = voltage across line.

This formula holds provided d and b are considerably greater than r. Substituting from Equ. (13) we find that

$$\frac{dV}{dr}\,(\text{peak}) = \frac{30}{r}.\sqrt{\frac{W}{Z}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots (16)$$

where W = power transmitted.

For operating under all the weather conditions experienced in the United Kingdom the voltage gradient should not exceed 10 kV/in. on bare wires, or about 1 kV/in. across bridging insulators.

Design of the Four-Wire Exponential Lines for Rugby

The line was required to transmit a continuous power

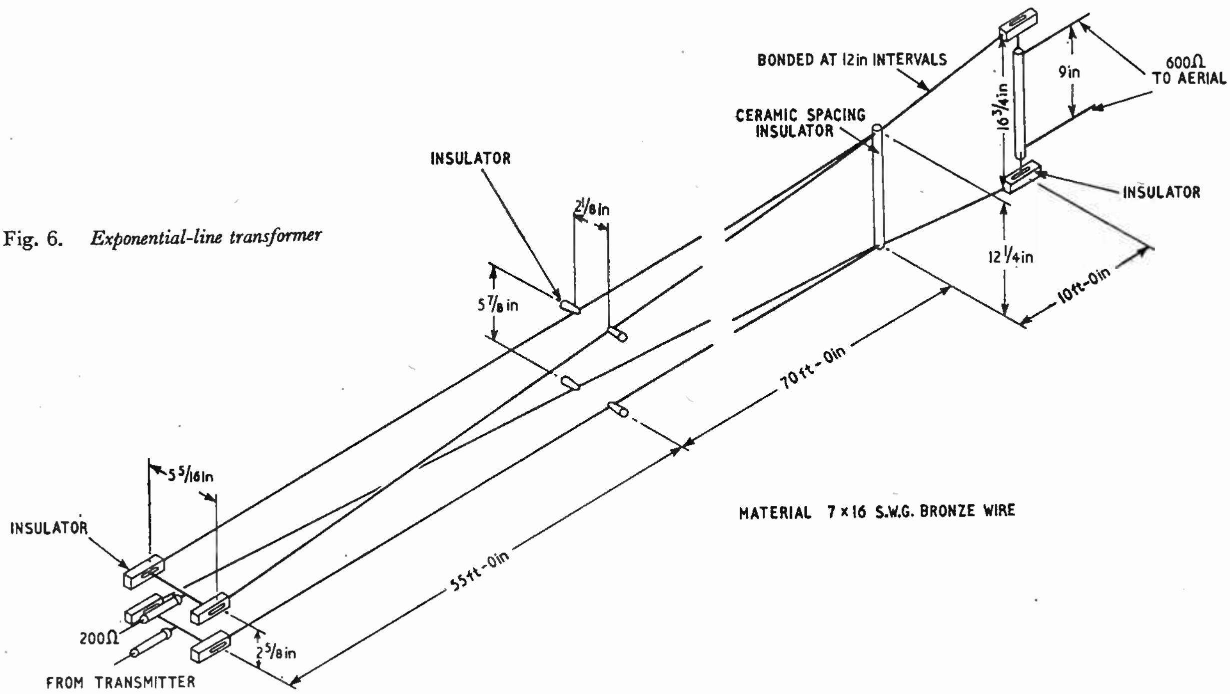


Fig. 6. *Exponential-line transformer*

of 20 kW, or a peak envelope power of 30 kW under single-sideband multichannel conditions, between 195 ohms and 565 ohms balanced impedances at all frequencies from 4–27·5 Mc/s. The v.s.w.r. in the 195-ohm line was to be better than 0·5 for an assumed constant v.s.w.r. of 0·67 in the 565-ohm line.

Referring to the previous section (to satisfy the last condition) the v.s.w.r. of the line when perfectly terminated must be better than 0·5/0·67 = 0·74. To achieve this, it is seen from Fig. 1 that a minimum value of $1/v = 6{\cdot}25$ is necessary.

Therefore $f_c = 4/6{\cdot}25 = 0{\cdot}64$ Mc/s and $\lambda_c = 1{,}560$ ft.

From Equ. (3) $q = 0{\cdot}008\ \text{ft}^{-1}$ and $l = \frac{1}{q} \cdot \log_e\left(\frac{Z_e}{Z}\right) = 133$ ft.

Values of $(b/r) = 50$, $(d/r) = 25$ were mechanically convenient for the input end of the line.

The maximum voltage gradient (under the conditions stated) using 400-lb copper wire will, from Equ. (16), be 9·3 kV/in.

Referring to Fig. 5, the procedure of the previous section showed that an impedance change from 195 to 304 ohms was found to be possible before a spacer was required. The procedure was then repeated with a new rate of taper which carried the line on to 525 ohms. From this point onwards a plain taper with the commoned wires touching was satisfactory.

A line as constructed is illustrated in Fig. 6. Stranded conductors, 7/16 s.w.g., have been employed in place of the solid 400-lb/mile conductors which were prone to kinking, the spacings being adjusted experimentally to allow for the slight increase in effective diameter. The lines were found to be relatively simple to construct and maintain, regulation being facilitated by four strainers at the low-impedance end.

The measured input impedance and the v.s.w.r. in the 195-ohm line when the exponential line is 'perfectly' terminated in a resistance of 565 ohms are shown in Fig. 7. The overall performance of a 195-ohm twin-coaxial line, an exponential transformer, a 565-ohm open-wire line and a rhombic aerial, all in tandem, is also given.

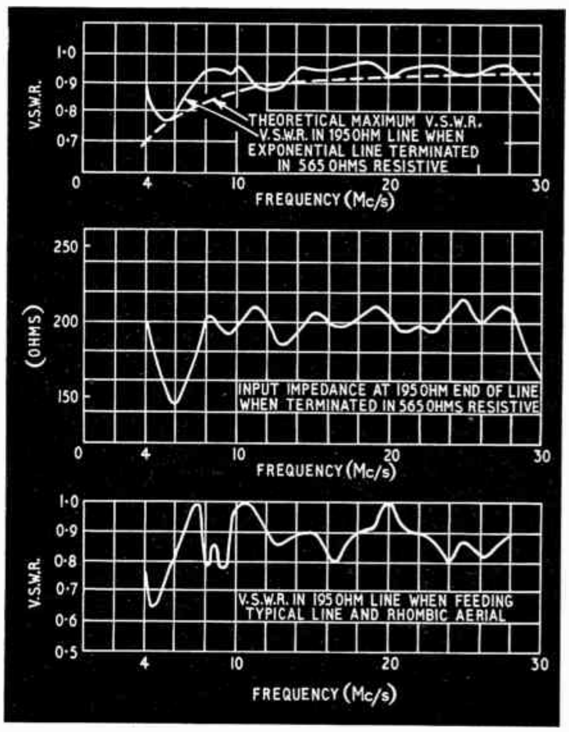


Fig. 7. *Performance of 4-wire exponential line. Note:* $Z = 195\ \Omega$, $Z_e = 565\ \Omega$, $l = 133$ *ft.*

Conclusions

A four-wire exponential line with linear tapers of wire spacing has been designed and tested, and has been shown to have excellent properties. It has only two changes of taper, and hence only two intermediate spacers are required. Lines of this type are simple to construct, easy to maintain, robust and appear to be insensitive to weather effects. They are now in large-scale use in conjunction with twin-coaxial distribution and switching at the Rugby 'B' radio station.

Acknowledgements

The author wishes to express his gratitude for the valuable assistance given by many members of the staff of the Engineer-in-Chief's Office of the General Post Office.

Acknowledgement is made to the Engineer-in-Chief of the General Post Office for permission to publish this paper.

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List of Symbols

Z = characteristic impedance of exponential line at sending end
l = length of line
Z_x, Z_l = characteristic impedance of exponential line at distances x, l, from sending end
q = impedance taper coefficient given by $Z_x = Z \cdot e^{qx}$
Z'_x = impedance presented by line at distance x from sending end
Z_r = terminating impedance
k = $\dfrac{Z_r}{Z_l}$
λ = wavelength in line
λ_c = cut-off wavelength; i.e., wavelength at which line presents infinite input impedance
f = frequency
f_c = cut-off frequency
v = f_c/f
p = propagation coefficient of line
T = transfer coefficient of line
n = $2Tl$
E = is given by $\sin 2E = v$

Diode Phase Detectors

CHARACTERISTICS OF SIMPLE AND BALANCED PUSH-PULL CIRCUITS

By S. Krishnan, M.Sc.*

SUMMARY. *A theory common to diode phase-meters and phase-sensitive detectors is worked out for two types generally known as the simple push-pull and the balanced push-pull detectors. The general case of unequal sinusoidal voltages being compared for phase is discussed and it is shown that equality of the two comparison voltages, assumed by previous workers to be the most suitable for phase-meter use, is inconvenient in certain cases. An alternative operating condition is suggested in which the signal is considerably larger than the reference voltage. Discussion of the effect of component variations establishes the superiority, in certain cases, of the balanced push-pull arrangement. A study of the effect of the output time constant indicates that, if the output capacitor is dispensed with, the output of the balanced push-pull arrangement becomes strictly proportional to the signal. Curves suitable for the design of phase-detectors and for the evaluation of their accuracy are included.*

The function of a phase-detector is the measurement of the phase difference between two sinusoidal voltages (or currents) of the same frequency. One of these is called the reference voltage and the other the signal voltage. This labelling of the two voltages as 'reference' and 'signal' may, in some cases, be purely relative whereas in others it may have a definite significance. For example, the circuit may be sensitive to amplitude

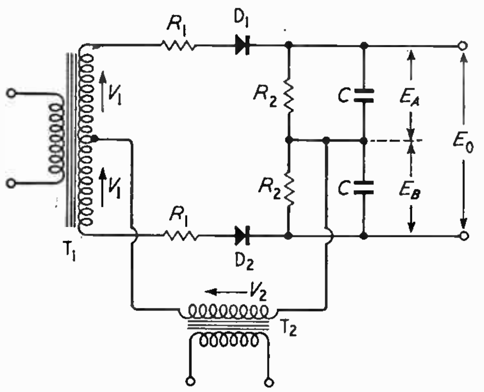

Fig. 1. *Circuit diagram of the simple push-pull detector*

variations of the 'reference' voltage but not so to the 'signal'. The diode phase detectors described in this paper are based on the principle of forming d.c. voltages dependent on the (vector) sum and difference of the two voltages whose phase difference is required. These detectors possess the advantages of simplicity and adequate accuracy for most purposes. The theory of two circuits, the simple push-pull detector and balanced push-pull phase detector, is worked out in detail.

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The theory is also applicable to another class of detectors known as phase-sensitive detectors. A short introduction to these is given before deriving the general theory.

Phase-Sensitive Detector[1]

A phase-sensitive detector is one whose output voltage reverses polarity as the phase of the input signal changes through 180°. These detectors also operate with a reference voltage which has the same frequency as the signal and whose phase remains constant. Such detectors are well known and some of their uses are as follows:

(a) For bridge balance indication. This seems to have been the first use of phase-sensitive detectors[2,3].

(b) Error detectors in servomechanisms.

(c) For the detection of a modulated carrier, the modulation being at extremely low frequencies. Modulation enables transmission of information at very low frequencies through transformers, amplifiers, etc., which may not have a response at these frequencies. One use is for producing very low-frequency sine waves.

(d) For coherent signal detection with large time constants when a physical phenomenon is suitably modulated; e.g., in the detection of nuclear magnetic resonance signals.

In the last case, the real advantages arise from the fact that the phase-sensitive detector is highly frequency selective and hence useful in picking up the signal from the noise accompanying it[4,5].

The present investigation arose out of the need to produce very slow sine waves for use in an analogue computer. Search for a suitable detector circuit led to a realization of the fact that the theoretical treatment of simple circuits has only been made under restricted conditions, such as the equality of comparison voltages in the case of the phase-meter. A theory was, therefore, worked out with no restrictions on the comparison voltages. Dependence of output voltage on component variations in the circuit was also worked out with a view to evaluating the accuracy to be expected of an instrument constructed on these principles.

The theory pertains to two commonly used detector circuits. In the simple push-pull detector two d.c. voltages are formed, one proportional to the amplitude of the (vector) sum of the two sinusoids to be compared, and the other to the difference between the two sinusoids. The difference between these two d.c. voltages turns out to be a comparatively simple function of the amplitudes of the a.c. voltages and their phase-difference. In the case of the balanced push-pull detector also, the (vector) sum and difference of the a.c. voltages are formed, but the derived d.c. voltages are more complicated functions of these.

A point, which does not seem to have been clearly appreciated, is that the difference between phase-meters and phase-sensitive detectors is often one of the actual operating conditions. Either the amplitude or the phase of the signal may be looked upon as the variable. When the sum and difference principle is employed, identical circuits can be used for phase measurement and for phase-sensitive detection.

In the existing theory it is common to assume that the 'reference' and 'signal' are of the same amplitude for the phase detector on the one hand, and the reference is assumed to be much larger than the signal for the phase-sensitive detector on the other. In the following theory no such assumptions are made and phase and phase-sensitive detectors emerge as special cases of the theory. It also follows that certain improved operating conditions are possible.

General Theory of Simple Push-Pull Detectors

The circuit for the simple push-pull detector is shown in Fig. 1.

Let $V_1 = E_1 \sin \omega t$ (reference), and $V_2 = E_2 \sin (\omega t + \phi)$ (signal) across the secondaries of transformers T_1 and T_2 represent the two voltages whose phase difference ϕ, is required. Let E_0 be the d.c. output voltage of the detector. The generators themselves are assumed to have negligible internal impedances, and R_1 represents all source and transformer impedances. R_2/R_1 is assumed to be much larger than unity. Further,

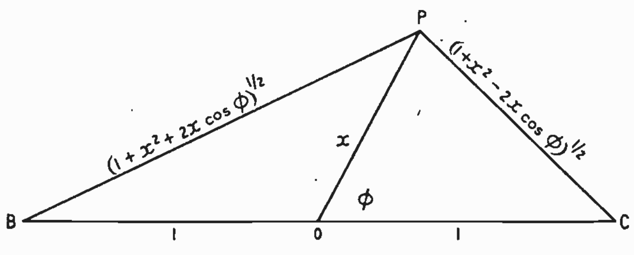

Fig. 2. *Graphical evaluation of the output voltage of the simple push-pull detector. (PB—PC) gives the ratio of output voltage (d.c.) to the reference amplitude; x is the ratio of signal to reference voltages*

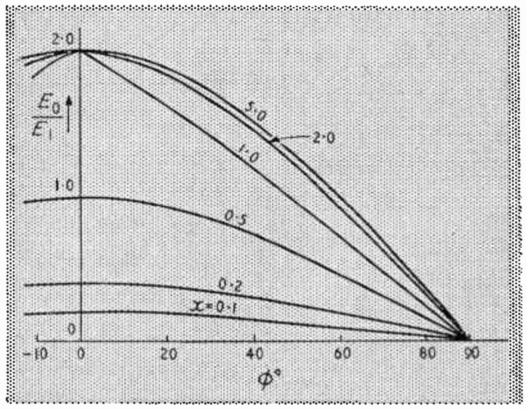

Fig. 3. *Curve of output voltage against phase-angle for the simple push-pull detector. Ordinate is the ratio of output (d.c.) to reference amplitude; x is the ratio of signal to reference voltages*

CR_2 is sufficiently large to make either half of the detector work as a peak-reading voltmeter. Under such conditions, it is easy to see that the voltage $E_0 = E_A - E_B$ is given by[6]

$$E_0 = (E_1^2 + E_2^2 + 2\,E_1\,E_2 \cos \phi)^{\frac{1}{2}} - (E_1^2 + E_2^2 - 2\,E_1\,E_2 \cos \phi)^{\frac{1}{2}} \quad \ldots \quad (1a)$$

$$\text{or } E_0/E_1 = (1 + x^2 + 2\,x \cos \phi)^{\frac{1}{2}} - (1 + x^2 - 2\,x \cos \phi)^{\frac{1}{2}} \quad \ldots \quad (1b)$$

$$\text{or } E_0/E_1 = \alpha - \beta \quad \ldots \quad (1c)$$

where $x = E_2/E_1$, $\alpha = (1 + x^2 + 2\,x \cos \phi)^{\frac{1}{2}}$
and $\beta = (1 + x^2 - 2\,x \cos \phi)^{\frac{1}{2}}$. The positive value of the square root has to be taken in each case. From the symmetry of Equ. (1a), it is readily seen that

$$(E_0/E_1)_{x = x_1} = (E_0/E_2)_{x = 1/x}$$

from which

$$(E_0)_{1/x} = (E_0)_x \cdot \frac{1}{x}, \quad \ldots \quad (2)$$

provided E_1 remains constant. The importance of Equ. (2) is that we need evaluate Equ. (1b) only for $x < 1$; and for $x > 1$, it is easy to evaluate E_0 using Equ. (2). Fig. 2 shows a geometric method of evaluating $(\alpha - \beta)$.
The figure is self-explanatory (PB — PC) being the required quantity.

Fig. 3 gives the curves of E_0/E_1 against ϕ for various values of x. The following special cases are important.

Case I. $x = 1$

When $x = 1$, Equ. (1b) reduces to

$$E_0/E_1 = 2\,\{|\cos \phi/2| - |\sin \phi/2|\} \quad \ldots \quad (3)$$

which, in the range $\phi = 0$ to $\pi/2$, yields

$$E_0/E_1 = 2\sqrt{2} \sin (\pi/4 - \phi/2) \quad \ldots \quad (4)$$

So far, $x = 1$, has been suggested as the operating condition for phase detection[7,8,9]. The curve corresponding to this in Fig. 3 is seen to possess an advantage in that the relationship between E_0 and the phase angle is nearly linear in the range $\phi = 0$ to $\pi/2$. [Actually, of course, it is part of the sinusoid of Equ. (4)]. But for use in this condition it is essential to make E_2 accurately equal to E_1. Any difference in the voltages will be wrongly measured as a phase change.

Case II. $x \gg 1$.

In this case taking the limit of Equ. (1b) as $x \to \infty$ we get $E_0 = 2\,E_1 \cos\phi$.

If we work under conditions of $x > 5$, maintaining the reference E_1 alone steady, then even a considerable variation in the voltage E_2 introduces no appreciable change in E_0. In fact, as E_2 increases from 5 to ∞, the change of ordinate at any point in the curve is less than 2%.

A word of caution however is necessary. The output voltage is the difference between the voltages developed by the upper and lower rectifiers, and any unbalance voltage will affect the accuracy. This is discussed in detail in a later section. In spite of this limitation, there is no doubt that the ratio $E_2/E_1 \gg 1$ will be very convenient in many applications, since an accurate adjustment of the amplitude E_2 will not then be necessary.

Case III. $x < 1$, $\phi = 0$ or π.

At $\phi = 0$ (or π), $E_0 = +\,2\,E_2$ or $(-\,2\,E_2)$ when $x < 1$.

Thus, making E_1 (reference) greater than the largest E_2 to be measured, the circuit can be used as a 'perfectly linear' phase-sensitive detector. To adjust for $\phi = 0$ or π, one should incorporate a phase-shifting network in one of the circuits which should be adjusted for maximum output.

Case IV. $x \ll 1$; ϕ, general.

In this case, $E_0 = 2\,E_2 \cos\phi$.

Kitai[6] has described a "phase-sensitive valve voltmeter" based on operation in this range.

General Theory of Balanced Push-Pull Detector

Farren[1] seems to have been the first to give a theory for the balanced phase detector. Dishington[8] has generalized the theory to take into account the non-linear characteristics of diode conduction. But both authors have discussed the case only for $x = 1$. The theory for general x is here described.

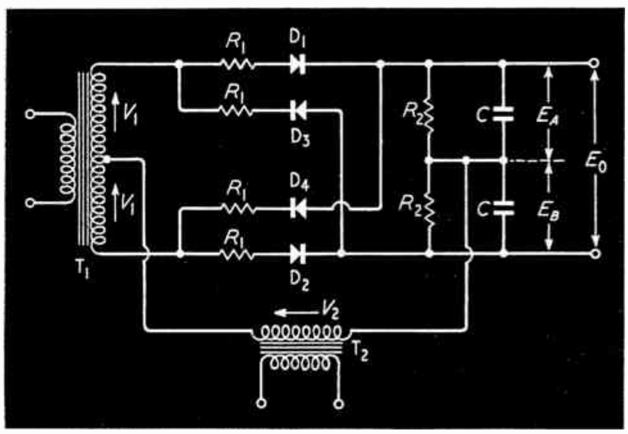

Fig. 4. *Circuit diagram of the balanced push-pull detector*

Fig. 4 is the circuit diagram of a balanced push-pull phase-detector. It is clear from the circuit that this is obtained from the simple push-pull detector by the addition of two more rectifiers D_3, D_4.

If we assume in this case also, that it is possible to lump all the resistances in series with each diode, the circuit is separable into two identical parts, except for the polarities of the diodes. One of these is shown in Fig. 5.

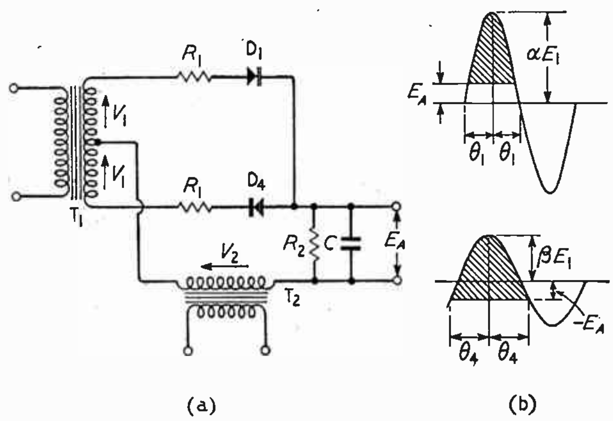

Fig. 5. *One of the separable halves of the balanced arrangement. The other half will be identical except for the polarity of connections of the diodes*

If, on the average, the current through D_1 is larger than that through D_4, a net positive voltage will appear at the junction of R_2C and D_4; i.e., E_A is positive. But this voltage on the capacitor C can leak through R_1 D_4 and, therefore, to realize a reasonably constant voltage across R_2 over a complete cycle, both time constants CR_1 and CR_2 must be large. We shall assume this to be the case. We shall further assume that $R_2/R_1 \gg 1$. The physics of the problem is as follows:

When E_A (positive) exists at the cathode of diode D_1, it conducts for a period of less than 180°. A similar argument shows that the diode D_4 conducts for more than 180°. Remembering the direction of the currents, the current through D_1 minus the current through D_4 (averaged over a whole cycle) equals the current through R_2, viz., E_A/R_2. Since, however, we have assumed $R_2/R_1 \gg 1$ we may neglect this current, and equate the current through D_1 over a cycle minus the current through D_4 over one cycle to zero. This is the basis of the following equation.

Amplitude of sinusoid applied to D_1 is αE_1 and to D_4 is βE_1, where α, β have the same significance as in Equ. (1).

If $2\theta_1$, $2\theta_4$ are the angles of conduction of diodes D_1, D_4, then it is clear from Fig. 5(b) that

$$\cos\theta_1 = E_A/\alpha E_1 \quad \ldots \quad (5)$$

and $$\cos\theta_4 = -\,E_A/\beta E_1, \quad \ldots \quad (6)$$

subject to $0 < \theta_1 < \pi/2$, and $\pi/2 < \theta_4 < \pi$.

Equating the charge through D_1 over one cycle to that through D_4 we get

$$\frac{1}{R_1}\int_{-\theta_1}^{+\theta_1}(\alpha E_1 \cos \omega t - E_A)\, d(\omega t)$$

$$-\frac{1}{R_1}\int_{-\theta_4}^{+\theta_4}(\beta E_1 \cos \omega t + E_A)\, d(\omega t) = 0 \quad .. \quad (7)$$

$$\text{or } \alpha E_1 \sin \theta_1 - \beta E_1 \sin \theta_4 = E_A(\theta_1+\theta_4) \quad .. \quad (8)$$

Putting $E_A/E_1 = y$ in Equ. (8) we get

$$y(\theta_1 + \theta_4) = \alpha \sin \theta_1 - \beta \sin \theta_4 \quad .. \quad (9)$$

It is necessary to appreciate that Equ. (9) does not lend itself to a straightforward solution. This is because determination of y requires a knowledge of θ_1 and θ_4 which themselves are related to y through Equs. (5) and (6). Further, both the angles themselves and their sine functions occur in Equ. (9). The following method is adopted for solving it.

TABLE 1

Relation between conduction angles of diodes D_1 and D_4 ($2\theta_1$, $2\theta_4$ are the angles of conduction).

θ_1	θ_4
77° 27′	180° 00′
78° 00′	135° 46′
79° 00′	121° 12′
80° 00′	113° 36′
81° 00′	108° 23′
82° 00′	104° 32′
83° 00′	101° 28′
84° 00′	99° 00′
85° 00′	96° 54′
86° 00′	95° 06′
87° 00′	93° 35′
88° 00′	92° 15′
89° 00′	91° 03′
90° 00′	90° 00′

Substituting for α, β from Equs. (5) and (6), Equ. (9) becomes

$$\tan \theta_1 - \theta_1 = \theta_4 - \tan \theta_4 \quad .. \quad .. \quad .. (10)$$

This shows that the relation between θ_1 and θ_4 does not depend on the particular values of x and ϕ responsible for them. Equ. (10) can be solved empirically for chosen values of θ shown in Table 1.

If we define $\theta_5 = \theta_4 - \pi/2$, we can combine Equs. (5) and (6) so that

$$\frac{\cos \theta_1}{\sin \theta_5} = \frac{\beta}{\alpha} = \left[\frac{1 + x^2 - 2x \cos \phi}{1 + x^2 + 2x \cos \phi}\right]^{\frac{1}{2}} \quad .. \quad .. (11)$$

Solving Equ. (11) for $\cos \phi$, we get

$$\cos \phi = \frac{1 + x^2}{2x}\left[\frac{\sin^2 \theta_5 - \cos^2 \theta_1}{\sin^2 \theta_5 + \cos^2 \theta_1}\right] \quad .. (12)$$

Using Equ. (12) to find ϕ for various values of θ_1 (and corresponding θ_5) for any value of x, we can then use Equ. (5) to calculate $y = E_A/E_1$. Thus we can tabulate y against ϕ for different values of x. Fig 6. gives curves of y against ϕ for different values of x.

In solving Equ. (10), we find that values of θ_4 can only be found for θ_1 greater than 77° 27′. For this value of θ_1, we find $\theta_4 = 180°$ ($\theta_5 = 90°$) which is the limit of θ_4. (This is readily appreciated physically by inspection of Fig. 5; $\theta_4 = 180°$ corresponds to conduction over the entire cycle.) As θ_1 increases from 77° 27′, θ_5 decreases, becoming 0° at $\theta_1 = 90°$.

When $x=1$, $\theta_1 = 77°$ 27′ corresponds to $\phi = 24°$ 30′. For $\phi < 24°$ 30′, therefore, we have θ_5 remaining constant at 90° ($\theta_4 = 180°$ and diode D_4 conducts over the entire cycle) and consequently θ_1 at 77° 27′. Thus from Equ. (5) we find

$$y = 2 \cos \phi/2 \ . \ \cos 77° 27' \quad .. \quad .. \quad .. (13)$$

[It is instructive to note that Equ. (6) breaks down, while Equ. (5) still holds for lower values of ϕ].

Equ. (13) for y gives a nearly flat response for y against ϕ since the cosine changes very slowly at low values of $\phi/2$ (<12° 15′ in the present case). Thus, the detector is very insensitive to phase angle changes at phase angles less than 24° 30′.

We shall now determine how the transition angle changes with x.

If there is to be no flat portion of the curve of y against ϕ (due to D_4 conducting over an entire cycle) the critical value, x_{crit}, can be found by putting $\theta_1 = 77°$ 27′, $\theta_5 = 90°$ and $\phi = 0$ in Equ. (11).

Substitution in Equ. (11) of the above values of θ, θ_1 and θ_5 gives $x = 0{\cdot}6433$. For values of x below this, we can find θ_1, θ_5 satisfying Equ. (10) and also Equ. (5) and (6). This is clearly seen in the graphs where the dotted line marks the boundary outside which Equs. (5), (6) and (10) are simultaneously satisfied at all values of ϕ.

The following special cases may now be discussed.

Case A. $x \geqslant 1$; ϕ, arbitrary

For $x=1$, the curve is nearly flat in the region $\phi = 0$ to 24° 30′. Dishington[8] has suggested a method of eliminating the flat portion, which depends on realizing the non-linear characteristics of the diodes. This, however, may often be a disadvantage because, apart from matching difficulties, one is constrained to work at low impedance levels (since one cannot include large series resistances with the diodes and yet realize non-linear characteristics). The curves of Fig. 6 clearly suggest an alternative solution. Instead of working at $x=1$, one may work at, say, $x=4$, where there is no

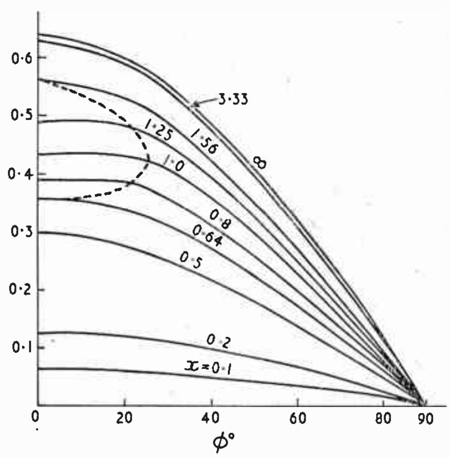

Fig. 6. *Curves of output voltage against phase-angle for the balanced push-pull detector. Ordinate is the ratio of output (d.c.) to reference amplitude; x is the ratio of signal to reference voltages*

really flat portion of the curve and one realizes good sensitivity over the entire range of phase angles without having recourse to non-linear operation. For matching purposes, we can now include sufficiently large resistances in series with the diodes so that the operation of the diode-resistance combination can be considered as linear.

Case B. $x < 1$; $\phi = 0$

In this case a linear relationship does not exist between y and x at $\phi = 0$. The circuit can be used as a phase-sensitive detector where a linear relationship is not necessary; e.g., in bridge balance indication or error-detection. Also, if $x \ll 1$, the output can be shown to be $2\,E_2/\pi$. In other words, if $x \ll 1$, a linear relationship exists between y and x, and this can be made use of.

Case C. $x \gg 1$

The expressions are symmetric with reference to E_1, E_2 in this case also and Equ. (2) can be used to find y for $x > 1$ once the values for $x < 1$ have been determined. Useful results are obtained when $x \gg 1$ and $E_A = 2/\pi\,[E_1 \cos\phi]$, giving $E_A = 2E_1/\pi$, at $\phi = 0$, which is the maximum E_A that can be obtained for the E_1. (Compare with the case of the simple push-pull, remembering that E_0 in the present case is $2E_A$.)

Effect of Unbalance in Circuit Parameters

The transformer T_1 can be made to give equal voltage on either side of the centre tap. In a circuit (once balanced) unbalance can be caused by a change in the forward conductance of one of the diodes or in the value of a resistance. These can be represented by a change in the ratio R_2/R_1.

Simple Push-Pull Detector

In this case, if the ratios R_2/R_1 for the two halves are not equal, the ratio of the output d.c. to the amplitude E_i (E_i stands for αE_1 or βE_1) of the input sinusoid will be different for the two halves, resulting in a net unbalance voltage to be added to E_0 [Equ. (1)]. It is easy to show that if 2ψ denotes the angle of conduction of the diode, then[7]

$$\frac{R_2}{R_1} = \frac{\pi\cos\psi}{\sin\psi - \psi\cos\psi} \quad \ldots \quad (14)$$

Also

$$\cos\psi = E_0/E_i \quad \ldots \quad (15)$$

What we require is the percentage change in E_0/E_i for a given percentage change in R_2/R_1; i.e., $\dfrac{d\,(\cos\psi)}{\cos\psi}$ in terms of $\dfrac{d\,(R_2/R_1)}{R_2/R_1}$. Differentiation of Equ. (14) and evaluation of the quantity required gives

$$\delta \equiv \frac{d\,(\cos\psi)}{\cos\psi} = \frac{1-(\cos 2\psi + \psi\sin 2\psi)}{2} \cdot \frac{d\,(R_2/R_1)}{R_2/R_1} \quad \ldots \quad (16)$$

This is plotted in Fig. 7 where δ (%) is the ordinate for 1% change in R_2/R_1.

We shall now determine the change in output voltage that will result from a 1° change of ϕ, so that we can express δ (%) by the equivalent error in degrees of phase; i.e., the error in measurement of phase-angle. Differentiating Equ. 1(b) and evaluating (ΔE_0) for $\Delta\phi = 1°$, we get

$$(\Delta E_0)_{\Delta\phi = 1°} \approx \frac{x\sin\phi}{57} \cdot \frac{\alpha+\beta}{\alpha\beta} \quad \ldots \quad (17)$$

For a 1% change (δ) in one of the branches, assuming the error to be in E_A, we obtain

$$(\Delta E_0)_{\delta = 1\%} = (1 + x^2 + 2\,x\cos\phi)^{\frac{1}{2}}/100 \quad \ldots \quad (18)$$

Combining Equs. (17) and (18), the error $\Delta\phi$ for a 1% δ is given as

$$(\Delta\phi)_{SPP} \approx \frac{0{\cdot}57}{x\sin\phi} \cdot \frac{\alpha^2\beta}{\alpha+\beta} \quad \ldots \quad (19)$$

(The subscript SPP stands for 'simple push-pull' detector.)

This is plotted against ϕ for various values of x in Fig. 8. To find the error in ϕ for a given change in R_2/R_1, the curves of Fig. 7 and 8 will have to be used. The error ($\Delta\phi$) read in Fig. 7 has to be multiplied by

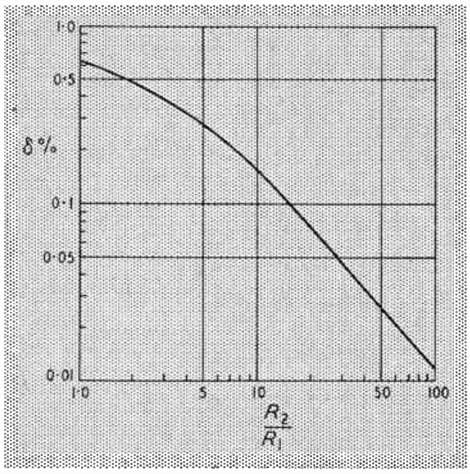

Fig. 7. *Curve of percentage error in output voltage of a diode-detector resulting from 1% change in the ratio of the shunt to series resistance in the diode circuit, drawn against the ratio of the resistances as abscissa*

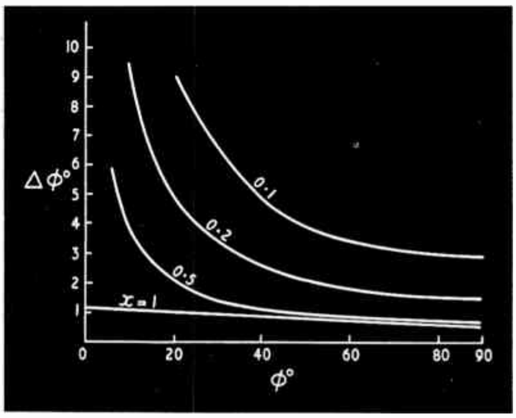

Fig. 8. *Error curve for the simple push-pull detector. This curve has to be used with Fig. 7. (See text)*

the δ (for the particular R_2/R_1 used) from Fig. 8 to get $\Delta\phi$ for a 1% change in R_2/R_1.

Balanced Push-Pull Detector

The analysis in this case is similar to the above but slightly more involved. We assume θ_1 and $\theta_4 \approx \pi/2$ (an assumption justified for most values of x and ϕ since we are concerned only with orders of magnitude). The result expressed in terms of $(\Delta\,\phi)_{SPP}$ is

$$(\Delta\,\phi)_{BPP} \approx 0{\cdot}8\,(\Delta\,\phi)_{SPP}\,(1-\beta/\alpha) \qquad \ldots \ldots (20)$$

(The subscript BPP stands for "balanced push-pull".) It may be remarked that $(\Delta\,\phi)_{SPP}$ is calculated for 1% δ while $(\Delta\,\phi)_{BPP}$ is for a 1% change in R_2/R_1.

The advantage over the simple push-pull detector is particularly evident for small values of x when $\beta/\alpha \approx 1$ and the error $(\Delta\,\phi)_{BPP}$ becomes negligibly small compared to $(\Delta\,\phi)_{SPP}$. The approximation made in deriving Equ. (20) (viz., θ_1, $\theta_4 \approx \pi/2$) is also particularly valid for small (and large) values of x.

Effect of Output Time Constant

It is interesting to see what happens if the output time constant is not large as has been assumed above. Let us take the extreme case where no capacitor is present.

In the simple push-pull detector, this merely results in a reduction in the output voltage (d.c. component) because the voltages developed across the upper and the lower halves both get reduced by the same factor.

In the balanced detector, however, the case is very different. Suppose D_1, D_4 are conducting (Fig. 4). If the current through R_2 is negligibly small compared with the currents through the R_1 resistance, we see that the output point should be at the potential of the centre tap of T_1; i.e., at the potential of the signal (relative to ground). This is also true when D_2 and D_3 are conducting with the only difference that the other output terminal has to be considered. The polarity of the reference signal determines which pair of diodes conducts. We find that for 180°, one pair of diodes conducts, and for the remaining 180° of a cycle, the other pair conducts. Thus, the reference signal switches the signal to the upper and lower rectifiers. The output (d.c. component) for one half is

$$\frac{1}{2\pi}\int_0^{\pi} E_2 \sin\,(\omega t+\phi)\; d(\omega t) = \frac{E_2}{\pi}\cos\phi$$

Therefore

$$E_0 = (2\,E_2/\pi)\cos\phi \qquad \ldots \quad \ldots \quad \ldots \quad \ldots (21)$$

The reference need only be large enough to switch the circuits efficiently and the output is proportional to both signal and cos ϕ. We have considered only the d.c. component and, if it is necessary to dispose of the ripple, a filter, presenting an essentially resistive load beyond the output points, may be included. A 'full-wave-type balance detector', similar to the one discussed, has been described by Morton.[10]

Effect of Noise and Harmonics

It is well known that phase-sensitive detectors are preferred where S/N ratios of the signal are very low. A particularly detailed account of the effect of noise in coherent detectors has been given by Tucker[4] and more recently by Kitai[5]. These discussions are applicable to the type of detectors here described. The most important result is that when one can afford a considerable 'waiting time' for getting the output, the coherent detector effects a considerable improvement over the incoherent.

To the extent that the circuits are balanced, there is no response to even harmonics. The response to odd harmonics depends upon the angle of conduction. For example, if a diode conducts over a 120° angle, the third and multiples of the third harmonics will not be detected[11]. In the case of the simple detector here described, the angle of conduction depends only on the ratio R_2/R_1. Thus, it is possible in this case to eliminate 3rd, 6th, etc., harmonics by choosing $R_2/R_1 \approx 4{\cdot}5$ (which corresponds to an angle of conduction of 120°). In the balanced push-pull detector, the angle of conduction (with a large output capacitor) depends upon x and ϕ, and is not within our control. The only thing that can be said is that the response to the 3rd, 5th, etc., harmonics is at least as low as 1/3, 1/5, etc., to that of the fundamental.

Acknowledgement

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LOW-FREQUENCY SINE-WAVE GENERATORS

In this article in the December issue, it was said that the lowest-frequency instrument produced one cycle in 33 minutes or 0·0005 c/s. We are informed that the Solartron JO744 will go down to 0·0001 c/s, or 1 cycle in 165 minutes. Two cascade integrators are used with an inverter in the feedback loop. A decade control of frequency is obtained by switched capacitors. A feature of the instrument is the way in which a large build-up time is avoided. The capacitors are given initial charges equal to their proper steady-state charges by pressing a button; oscillation can then start off from initial conditions equal to the steady-state ones.

In referring to the Venner Electronics quartz-crystal oscillators, it was stated in error that they included point-contact transistors. In fact, they use junction transistors.

HELIUM REFRIGERATORS

Although liquid helium, in one form or another, is the only genuine and authentic superfluid, it is rarely mentioned publicly beside those loud pretenders to the title whose excellences (as discerned by their purveyors) are commended to us daily. It may be because it can go no better than super, and the bidding nowadays starts at fabulous. Unfermented and unpotable, non-aromatic and with zero octane rating, never to be made-in-a-jiffy-the-easy-way, and not even a drug on the market, it has negative social prestige and is ignored in chilly silence. Indeed, it fares but little better than some of the truly fabulous products that our benefactors have hitherto omitted to invent—such as the tergents, which Shrink Smalls Smaller while converting them into perfectly black bodies. But justice will in the end prevail. After all, it took some time for liquid oxygen to establish itself as a standard industrial product; so there is no reason to doubt that even helium may one day achieve a place in the sun. And what has bent (I resent the term 'warped') my fairly inflexible mind in this direction is the realization that helium liquefiers have been marketed commercially for upwards of a decade, while helium-operated refrigerators have been going for some time. I have had a chance to look into this recently, for the processes are discussed, with a good deal of historical background, in "Expansion Engines for Low Temperature Processes", by S. C. Collins and R. L. Cannaday (Oxford University Press, 1958). The Collins liquefier, and its development at the Massachusetts Institute of Technology, is described with more detail in the paper by S. C. Collins on "A Helium Cryostat" (*Review of Scientific Instruments*, 1947, Vol. 18, pp. 157–167.)

There are three aspects to refrigeration. First, the use of the first law of thermodynamics to reduce the internal energy of a body; next, the second-law problem of doing this as efficiently as possible; and thirdly, the use of the body cooled in this way to extract heat from other bodies, and to do this continuously. The body, or working substance, can in general be a solid, a liquid in equilibrium with its saturated vapour, or a gas. This article will treat gas refrigeration from a general point of view, though with special reference to helium; and the 'body' will be taken, for most of the calculations, to be one gram of gas.

Internal Energy Reduction

For unit mass of substance, the internal energy is denoted by U, the enthalpy or total heat by H, and the entropy by S. The first two are expressed in calories per gram, while S is in calories per gram per degree K; all three are functions of the pressure p, volume V, and absolute temperature T °K.

We may get into trouble over units before long. I should prefer to work in mechanical energy units, but all the Collins' calculations seem to be in calories. If, in what follows, you take p to be in dynes per sq. cm. and V in c.c., then the expression pV is in ergs, which you must mentally divide by $4{\cdot}2 \times 10^7$, as J has not been written in; it would be better to suppose that all this has been done for you, and to accept pV as expressed in calories. The same applies to external work. But we had better start off straight, as I dare not mess about with k and R.

The internal energy U (for 1 gram) of an ideal gas is the total kinetic energy of translation and rotation of the molecules; for a monatomic gas its value is $\frac{3}{2}NkT$ ergs, where N is the number of molecules per gram, and k Boltzmann's constant; this may also be written as $\frac{3}{2}RT/M$ ergs, where M is the molecular weight and R the universal gas constant; U is a function of T alone. To get U in calories per gram, divide by J.

But for a real gas, there is a potential energy term due to molecular attraction, and U depends on p, V, and T together. There are, of course, what may be called hyperfine-structure contributions to U that matter in other contexts; but you don't count the cost of the ink when you are worrying about writing out a cheque. In any case, helium can safely be treated as an ideal gas except at the very last stage of its descent to the liquid state.

The enthalpy H is defined by the equation $H = U + pV$. We might get along very well without H for an ideal gas, possibly; but it is useful in dealing with heat changes in a real gas, particularly where there is prospect of a change of state.

Entropy S is usually defined incrementally. If a body receives a quantity of heat ΔQ at temperature T, then $\Delta S = \Delta Q/T$, so that the heat-gain itself can be written $\Delta Q = T\Delta S$.

Values of H and S are available in tables; U is readily calculated for an ideal gas. The only other quantities to be mentioned are the specific heat at constant pressure, C_p, and the ratio γ of the specific heats at constant pressure and at constant volume. The value of γ is 1·67 for helium, a monatomic gas. We shall really be thinking all the time of one gram of gas, but the word 'body' will be used for this where a statement is more generally true.

From the first law of thermodynamics, if heat ΔQ *supplied to* a body results in an *increase* ΔU in the internal energy and the performance *by the body* of external work ΔW (all measured in calories), then

$$\Delta Q = \Delta U + \Delta W.$$

If the body is thermally insulated from the surroundings, and then made to do external work without supply or loss of heat, $\Delta Q = 0$; the change is called adiabatic, and since ΔS is also zero it is isentropic. In this case $\Delta U + \Delta W = 0$, and there is a loss of internal energy equal to the external work done by the body.

An important feature of such a change is that it is reversible, in the sense that if ΔW is done on the body by an outside agent, then an equal increase ΔU in the internal energy results. Strict reversibility demands that the process, either way, can proceed by infinitesimally small steps through a succession of equilibrium stages.

An ideal gas, expanded in this way from p_1 to p_2, falls in temperature from T_1 to T_2, where $T_2/T_1 = (p_2/p_1)^{\gamma-1/\gamma}$ The heat energy extracted from unit mass is then $C_p\,(T_1 - T_2) = C_pT_1\,[1 - (p_2/p_1)^{\gamma-1/\gamma}]$. Putting in the value of γ for helium, 1·67, it is seen that a pressure ratio of 1/10 gives a temperature ratio of 0·4, a pressure ratio 1/100 a temperature ratio 0·16, a pressure ratio 1/10,000 a temperature ratio 0·0256, giving temperatures of about 120 °K, 50 °K, and 8 °K respectively, starting at 300 °K.

Starting at air temperature, then, you could expand from a pressure of 10,000 atmospheres to one atmosphere, and still fail to reach the boiling-point of helium at atmospheric pressure, which is 4·2 °K. On the other hand, starting at 40 °K with a 10-atmosphere expansion would get down to 16 °K, and starting again from there with a second 10-atmosphere expansion would come down to 6·4 °K, after which a third would do the trick. Collins and Cannaday suggest that four stages from air temperature might be sufficient, though this is not the actual method used. The Collins liquefier has two relatively low-pressure expansions, and the incoming gas is cooled on its way to the expansion chamber by a counter-current heat exchanger in which it flows in thermal contact with a returning stream of expanded and cooled gas. A single adiabatic expansion was indeed used by Simon, starting at a low enough temperature; and in the Claude and Heylandt liquid-air processes there is only one adiabatic expansion stage, while the older gas-liquefaction processes for air and

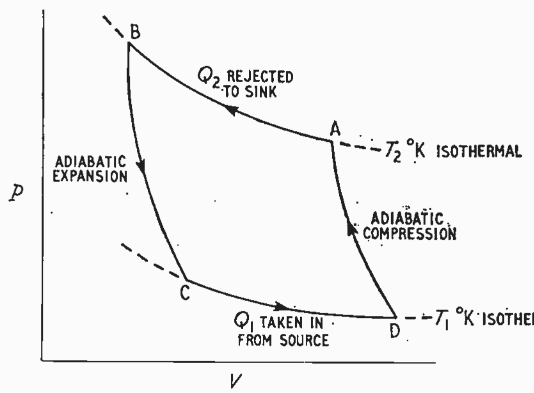


Fig. 1. *The Carnot cycle. Although the working substance need not be an ideal gas, this is what the curves (as shown) apply to. Lines CD and AB are the Boyle's-law hyperbolas for T_1 and T_2; lines DA and BC are adiabatic 'pV^γ = constant' curves*

hydrogen did not use adiabatic expansion at all, relying on the Joule-Kelvin effect. This is a real-gas effect, and is irreversible; it is the final process in the Collins apparatus.

The Joule-Kelvin effect is called a throttling process, and also an isenthalpic process. If unit mass of gas at p_1, V_1, is allowed to expand freely through a valve, or by seeping through a porous plug (the pressure on the input side being maintained steadily at p_1 by a compressor which does the necessary *external* work) and issues at pressure p_2 and with volume V_2, then there is no change in enthalpy, and $U + pV$ is the same on both sides. Thus $U_1 + p_1V_1 = U_2 + p_2V_2$, and $U_2 - U_1 = p_1V_1 - p_2V_2$. Anything may happen here. For an ideal gas, U depends on T alone anyhow, and does not change. If $p_1V_1 < p_2V_2$, then $U_2 < U_1$, and cooling results; if $p_1V_1 > p_2V_2$, then $U_2 > U_1$, and the temperature rises. The curves of pV plotted against p at different temperatures for various gases appear in all the books; for all gases, except hydrogen and helium, $p_1V_1 < p_2V_2$ at air temperature provided the initial pressure is not extremely high; so they are cooled by throttling expansion. Hydrogen and helium behave in this way only if they are cooled to a low enough initial temperature before expansion; at moderate initial temperatures they heat up. This 'low enough temperature' in the case of helium is somewhere between 25 °K and 60 °K at round about one atmosphere pressure. The effect is not quite as simple as it appears in the more elementary books, where you may find a value for 'the' inversion temperature calculated on the assumption that the gas follows van der Waals' equation. The point is that with helium the Joule-Kelvin effect can only be invoked if the temperature has already been very considerably reduced; when it does come into its own, the drop in temperature is nearly proportional to the pressure difference (not ratio), and is considerable. In the Collins liquefier, helium arrives at the expansion valve at 15 atmospheres and 8 °K, and is cooled to 4·2 °K at approximately one atmosphere.

We might break off for a moment to consider air or nitrogen, where the isenthalpic cooling is about a quarter of a degree per atmosphere difference at 300 °K, and about 2½ degrees per atmosphere at 100 °K. Using a counterflow exchanger between the incoming and outgoing streams, no very great pressure is needed to liquefy air, though some processes did go up to 200 atmospheres. The whole thing seems so simple, that one asks why expansion engines are ever employed for oxygen and nitrogen at all. The reason is a matter of efficiency, and thus of cost. A completely irreversible effect such as throttling is simply more expensive to run than the theoretically reversible expansion engine.

Supplying Energy in order to Extract It

Few of my non-scientific friends seem to be at all surprised that they have to light a gas-burner underneath the fridge in order to cool the contents. It may be that they think that gas, like water, comes out of two sorts of tap, respectively h. and c. But I suspect that they are well enough grounded in the first law to realize that you can't make anything work without supplying energy that has to be paid for. Maxwell's enunciation of the second law of thermodynamics was to the effect that you cannot transfer heat from a cooler body to a hotter body without the performance of work.

It is fair to set up the Carnot cycle as a basis for comparison because, although it could not be made into a satisfactory refrigerator, the Stirling cycle which is very much like it (and has the same efficiency between the same two temperatures) is used practically in some machines. In the Carnot cycle we imagine a fixed mass of working substance (which may as well be an ideal gas, but needn't necessarily be so) enclosed in a cylinder by a piston which can always be loaded just to balance the pressure from within. The cylinder has non-conducting walls and a perfectly conducting base (and so on) and we suppose that it can be put into perfect thermal communication with a low-temperature source of heat at T_1 and a high-temperature sink at T_2, and that at the appropriate stages in the cycle it can be perfectly isolated for adiabatic purposes. The steps shown in Fig. 1 can be described in the cyclic order AB, BC, CD, DA. A quantity of heat Q_1 is taken in at T_1; a larger quantity Q_2 is rejected at T_2; there is no heat change during either of the adiabatic operations; so the difference $(Q_2 - Q_1)$ represents, by the first law, the external work that has been done by the working substance. As it ends up where it began, its internal energy has not altered, so all must have come from outside. Thus, to extract a quantity of heat Q_1 a quantity $(Q_2 - Q_1)$ has had to be put in first. Using the ideal gas isothermal work expression $RT \log_e (p_2/p_1)$ for a pressure ratio of 10 and temperatures approximately 300 °K and 4 °K, Q_1 works out to be about 5 calories and $(Q_2 - Q_1)$ about 340 calories for one gram of helium. This kind of arrangement is not practicable, but it sets a sort of limit; at least, as to cooling the gas without liquefaction.

Next, suppose that, without throttling, liquid helium could be obtained by a reversible process at 4·2 °K from gaseous helium at 300 °K, both at one atmosphere pressure. The entropies per gram of gaseous and liquid helium under these conditions are 7·53 and 0·80 calories per degree. The entropy change ΔS is thus 7·53 − 0·80 = 6·73 units, and the heat rejected at 300 °K, $T\Delta S$, is 300 × 6·73 = 2,019 calories.

The enthalpies of the gas and the liquid are 375 and 2·34 calories per gram; so the total heat extracted is 375 − 2·34 = 372·7 calories. Thus, Q_1 = 372·7 cal., and Q_2 = 2019 cal., so $(Q_2 - Q_1)$ = 2,109 − 372·7 = 1,646·3 cal. This is an energy balance, with no argument about ways and means, and without invoking the second law at all; merely from the first law, it can be seen that no process for producing *liquid* helium could operate on less than this energy supply per gram. The Carnot cycle, it will be noted, was simply cooling the gas anyhow, not liquefying. The second law shows that irreversible processes must necessarily lead to reduced efficiency, needing a greater energy supply. Some of the irreversibility in any apparatus comes from imperfect insulation and friction, but the throttling process is deliberately irreversible. Collins calculates that, for two stages of reversible adiabatic expansion followed by an isenthalpic expansion, the minimum energy supply should really be 4,675 cal. per gram. In actual practice,

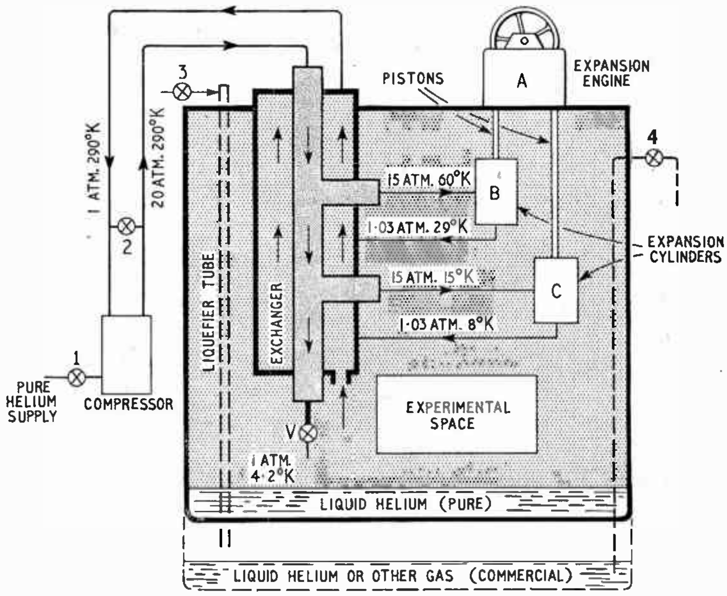


Fig. 2. General scheme of the Collins liquefier, set out in the conventional way. The thick outline represents the cryostat proper, while all the broken lines refer to the liquefaction of gas from an outside supply; all this is within the insulated outer case which is not shown. Details of purifiers, etc., are not included; it is assumed that the charge of very pure helium introduced at 1 circulates indefinitely without loss. The Chad-like object is the flywheel of the expansion engine

with the liquefier of his 1947 paper it came out to about 13,000 cal. per gram.

Taking the figure given by Kaye and Laby for the density of liquid helium at 4 °K, 0·15 gm. per c.c., and doing the necessary unit conversion for comparison with the published figures for liquid air plants, this works out to about 10·4 kilowatt-hours per gallon. The figures for the early Hampson process and the Heylandt process for liquid air are 10·3 and 3·2 kWh per gallon, while the entirely reversible Keesom cascade process required 2·04 kWh per gallon. From the point of view of running cost a modern helium liquefier thus seems about on a par with a primitive air liquefier, and three times as costly as a modern air liquefier using the same principles.

The Collins Liquefier

Fig. 2 shows the helium circulation in the conventional way, but does not give a fair picture of the apparatus itself. The whole is a slightly tapered double-walled cone about 4 feet deep and 8½ inches wide at the top, with the expansion engine (A) mounted on the lid, and the two actual expansion cylinders (B) and (C) suspended inside. The heat exchanger, drawn as two coaxial cylinders, was not at all like this. It is hard to give a true impression without a diagram, which would then take a whole article to explain. The inflowing gas passed through a 300-foot helix of finned copper tubing (later models used getting on for a mile) wound on the inner wall, while the outflowing gas circulated up the annular space between the walls. This outfit was then surrounded by a metal dewar vessel, at the bottom of which helium could accumulate. The apparatus was designed as a cryostat bath for low-temperature experiments as well as a liquefier, and there was a space above this liquid which served as an experimental chamber.

The vessel was in turn mounted in a leak-proof insulating case. The compressor, outside the apparatus, received pure helium gas as required through the tap 1; this was compressed to 20 atmospheres, cooled to air temperature, and delivered to the exchanger, down which, when things were steady, there was a steady temperature gradient. Part of the gas, at the 60 °K level, was taken in by the expansion cylinder (B), made to expand adiabatically, and restored to the low-pressure stream at 29 °K; lower down, at the 15 °K level, a further portion was expanded in the cylinder (C), to rejoin the upward stream at 8 °K. The final isenthalpic expansion from 15 atmospheres to about 1 atmosphere through the valve (V) reduced the temperature to 4·2 °K and liquefied the gas. The expansion engine was, of course, driven by the energy extracted in (B) and (C) from the gas, but a small electric motor was used to regulate its speed. In order to make sense of the temperatures written in Fig. 2, you must remember that the horizontal flow-tubes to the cylinders are not really placed like that, and that the whole space in the cryostat is filled with gas (or saturated vapour if you're a physicist) at 4·2 °K.

To obtain liquid helium, a separate liquefier tube, shown by broken lines in the figure, wound round the conical vessel like the high-pressure side of the exchanger, is used. This goes through to a chamber below the cryostat proper; commercially pure helium is supplied through tap 3, and siphoned off through tap 4—just like that! Tap 2, joining the high- and low-pressure sides of the compressor leads, can be used to regulate the pressure, and hence the temperature, within the cryostat. To liquefy nitrogen or hydrogen—which both solidify above 4·2 °K anyhow—the full severity of the process is not invoked. Valve (V) is kept closed, only the two adiabatic expansions of helium are used, and the gas comes in at 3 and is drawn off as liquid at 4. This particular machine would produce about 1 litre of liquid helium, or about 2 litres of liquid nitrogen or hydrogen per hour.

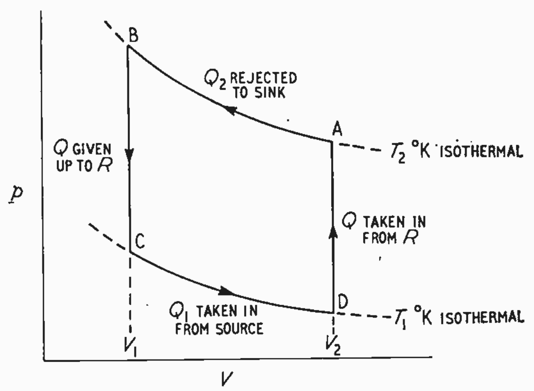


Fig. 3. *The Stirling cycle. Stages AB and CD are just as in the Carnot cycle; but DA is a constant-volume heat intake, and BC a constant-volume heat rejection*

The chief features contributing to the high performance are (i) efficient heat exchange; (ii) the use of flexible piston rods which enabled the pistons to align themselves in the cylinders instead of fighting friction; (iii) suspension of the whole apparatus within the cryostat, and (iv) the reduction of gas leakage outwards, and heat leakage inwards from the surroundings.

Why not a Turbine?

In a reciprocating engine, the piston grating along the cylinder wall generates heat in just the very worst place possible; a pistonless engine would be better. This was pointed out in 1898 by Rayleigh, and several air-liquefaction turbines seem to have been designed. The first to work was that of Kapitza, in 1939; this was followed in 1942 by the Elliott-Sharples turbine liquefier. Collins and Cannaday devote a whole chapter to problems of turbine design. All that need be said here is that, although the working fluid comes out of a nozzle, the process is an isentropic adiabatic expansion, and not an isenthalpic throttling. Turbines have not got down to liquefying helium yet; but it would seem that the turbine may be the machine of the future.

The Stirling Cycle

The cycle of Fig. 3, which was devised in 1816 by Robert Stirling, and used for refrigeration in 1873 by A. G. Kirk, is a completely reversible cycle resembling the Carnot type operating on a fixed mass of working substance. Heat Q_1 is extracted at the low temperature T_1, and heat Q_2 rejected at the higher temperature T_2, just as in Fig. 1; but the intervening strokes between the two isothermal heat-changes take place at constant *volume*. On these two strokes, no external work is involved, but the enthalpy changes; and heat is absorbed on the stroke DA at constant volume V_2, and rejected on the stroke BC at constant volume V_1; these two quantities of heat have the same value, say Q.

To make the cycle work, this heat Q has to be banked in a reservoir during the stroke CD. This is done by passing the heated gas at constant volume through a pad of finely-compressed copper gauze on the stroke BC; Q is taken in by the gauze, stored, and then returned when the cooled gas is circulated through it at constant volume V_2 on the stroke DA. The process, as explained in *Nature* (1st October, 1955) by Dr. J. W. L. Köhler of the Philips Research Laboratories, Eindhoven, is illustrated in Fig. 4. The main piston P_1 does the isothermal compression AB and isothermal expansion CD. The displacing piston P_2 travelling out of phase with P_1, moves in a narrower cylinder surrounded by the hot heat sink at T_2, the cold heat source at T_1 (both represented in the figure by coils through which fluid can circulate), and by the reservoir or regenerator R. The lettering A, B, C, D, A corresponds to the points on the cycle of Fig. 3. Between B and C, the displacer P_2 drops, forcing the heated gas through R; between D and A, the displacer P_2 rises, circulating the cooled gas through R the other way. Since the cycle is reversible, and the only heat changes are isothermal as in the Carnot cycle, the efficiency should be the same as that

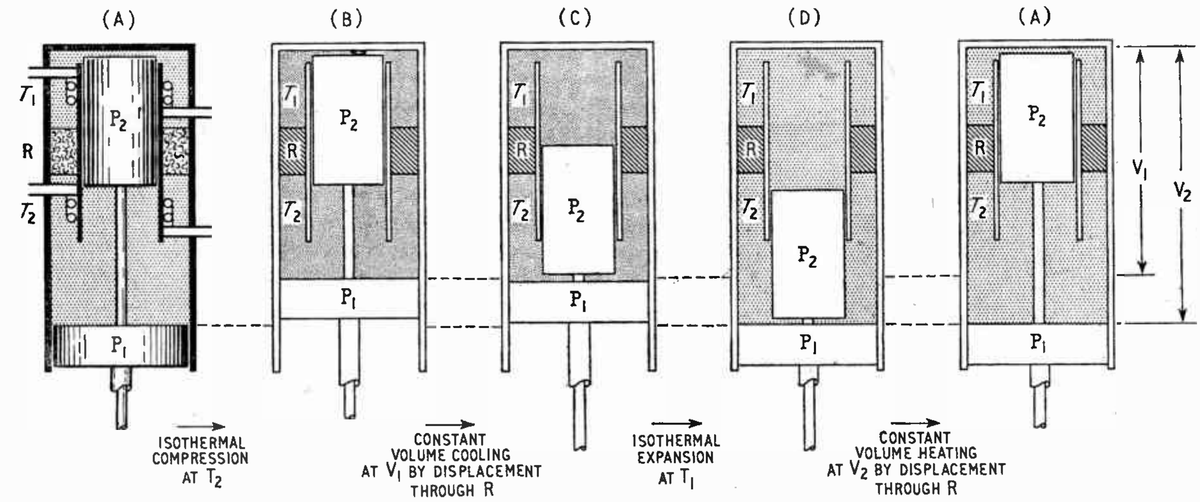


Fig. 4. *The Stirling cycle as used in the Philips helium gas refrigerator. The letters A, B, C, D refer to Fig. 3. The best way of following the diagram is to take one piston at a time. Piston P_1 makes two strokes—up between A and B (isothermal compression), and down between C and D (isothermal expansion); piston P_2 also makes two strokes—down between B and C, sending the gas from the bottom to the top of the cylinder through R (constant-volume cooling), and up between D and A, sending the gas from the top to the bottom through R (constant-volume heating)*

of the ideal Carnot cycle operating between the two temperatures.

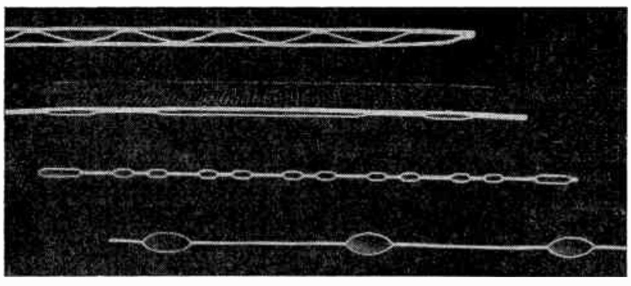
In the Philips gas refrigerator described in Köhler's article, helium is used as the working substance. Air is liquefied by passing it through the equivalent of the T_1 coil. As with the Collins machine, the actual refrigerator does not really look much like the schematic diagram given here. Collins and Cannaday describe the mechanism fully, and point out that the relatively high capacity (5·5 litres of liquid air an hour) results from the high crankshaft speed. Its energy consumption is about 1 kWh per kilogram, which, expressed as 3·7 kWh per gallon, compares favourably with other liquid-air systems.

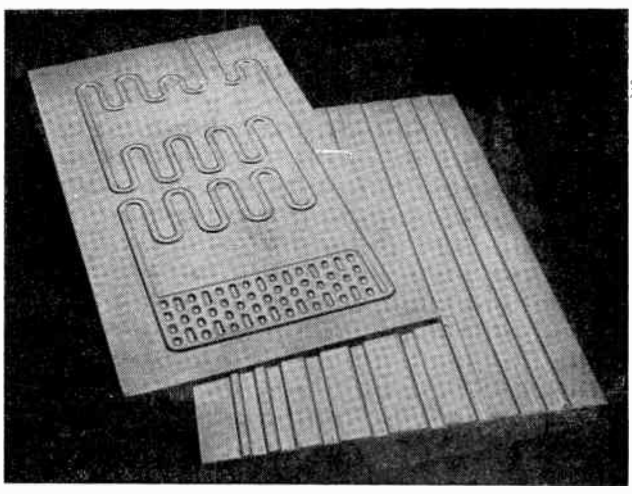
Conclusion

The scope of this article has been limited to one point of principle—the use of the almost-ideal gas helium as the medium for temperature reduction. Its liquefaction is to some extent merely incidental. Later I hope to deal with the same principle applied to solid-state media. This interest in low-temperatures is not a sort of reaction from thermonucleonics; the subject is well in the fringe of the field—to mention only one example, the three-level maser work has so far been done in a helium cryostat. There is indeed matter for discussion here; but keep calm—it will not be heated discussion.

DUCTED ALUMINIUM SHEET

The photographs show channelled sheets of aluminium alloy which are being produced by the Northern Aluminium Company Ltd. (Bush House, Aldwych, London, W.C.2) for use in heat-transfer equipment. In the electronics field it can be used for cooling radio chassis by blowing air or water through the internal system of passages or ducts. The material looks neat and the absence of joints minimizes the risk of coolant leakage.





Surface Impedance

MEASUREMENT AT V.H.F.

By J. C. Anderson, M.Sc., A.M.I.E.E., A.M.Brit.I.R.E.*

SUMMARY. *The high-frequency surface impedance of a conducting material may be measured by use of a coaxial transmission line. The method described does not require the use of a calibrated r.f. meter, and employs a specimen in the form of a disc. Measured values of impedance may be used to yield the v.h.f. permeability of a ferromagnetic specimen, and results are given for measurements on pure nickel.*

To determine the suitability of a conducting material for use in the v.h.f. range, it is necessary to measure its surface impedance as defined by Maxwell's equations; this is particularly the case for coated conductors, where the skin depth for current at the frequency employed may be comparable with the thickness of the coating.

In the method here described all measurements required are of length, and these measurements are related to the transmission-line equations so as to yield a value for the surface impedance of a disc terminating the coaxial line.

The quantity actually observed is the current in the disc (acting as terminating impedance to the line) as it varies with position of an adjustable short-circuiting piston. The position of the piston is read for the first and second peaks of current in the termination, and also for the points at which the first peak has a height equal to the maximum height of the second peak.

Theory of the Method

It has been shown[1] that the propagation constant of a transmission line may be written as

$$P = \alpha + j\beta = 2\pi/\lambda . (\tan\delta + j) \qquad \ldots \quad \ldots (1)$$

where $\tan\delta_1 = R/\omega L$, $\tan\delta_2 = G/\omega C$ and $\delta = \frac{1}{2}(\delta_1 + \delta_2)$; R, L, C, G, λ and ω having the usual significance.

If the line, of length l, is terminated in an impedance $Z_R = (R + jX)$ and carries a current I_R, then we have, from the normal equations

$$I_R = \frac{V_s}{Z_R \cosh Pl + Z_0 \sinh Pl}$$

where Z_0 is the characteristic impedance (assumed to be purely resistive) and V_s is the signal voltage applied at the sending end of the line.

Substituting for P and Z_R and taking a modulus we get

$$\left| I_R \right| = \frac{V_s}{[RZ_0 \sinh 2\alpha l - XZ_0 \sin 2\beta l + \frac{1}{2}(R^2 + X^2)(\cosh 2\alpha l + \cos 2\beta l) + (Z_0^2/2)(\cosh 2\alpha l - \cos 2\beta l)]^{\frac{1}{2}}}$$

For the condition of maximum current in the termination we differentiate and equate to zero. Substituting for α and β from Equ. (1), remembering that β is a lagging phase angle and must therefore carry a negative sign, we introduce $u = l/\lambda$, when we get

$$\tan\delta\,[R Z_0 \cosh(4\pi u \tan\delta) + \tfrac{1}{2}(R^2 + X^2 + Z_0^2)\sinh(4\pi u\tan\delta)] - XZ_0\cos 4\pi u - \tfrac{1}{2}(R^2 + X^2 - Z_0^2)\sin 4\pi u = 0$$

This equation may be solved for u numerically or by use of Newton's method. Letting $y = 4\pi u$, the latter states that if $y = y_1$ is an approximate solution of $f(y) = 0$, then a better solution is $y_2 = y_1 - f(y_1)/f'(y_1)$. For a line terminated in an impedance less than Z_0, which will normally be the case, a peak will occur at $l = \lambda/2$ approximately. Thus an approximate solution is $y_1 = 2\pi$, and applying Newton's method

$$y_2 = 2\pi - \frac{(R Z_0 \tan\delta - X Z_0 + \pi Z_0^2 \tan^2\delta)}{Z_0^2/2 . (1 + \tan^2\delta)}$$

$$\text{for } Z_0^2 \gg (R^2 + X^2) .$$

There will obviously be a maximum value for current in the disc when $l = 0$ (i.e., when $u_1 = 0$), so that we may write the distance between successive peaks in terms of u as

$$u_2 - u_1 = \tfrac{1}{2}(1 - \tan^2\delta) - (R\tan\delta - X)/2\pi Z_0 \quad (2)$$

whence the physical distance between peaks, d, is

$$d/\lambda \approx \tfrac{1}{2} - (R\tan\delta - X)/2\pi Z_0 \qquad \ldots \quad \ldots (3)$$

where $\tan^2\delta$ is assumed negligible compared with unity. We now have the impedance of the disc in terms of the distance between successive peaks but, in order to separate R and X, we require another equation. This is obtained by the measurement of the width of the first peak at a height equal to that of the second peak.

The experimental condition is represented by the equation $|I|(u_{3,4}) = |I|(u_2)$ where $u_4 - u_3$ is the width of the first peak as shown in Fig. 1. Since the

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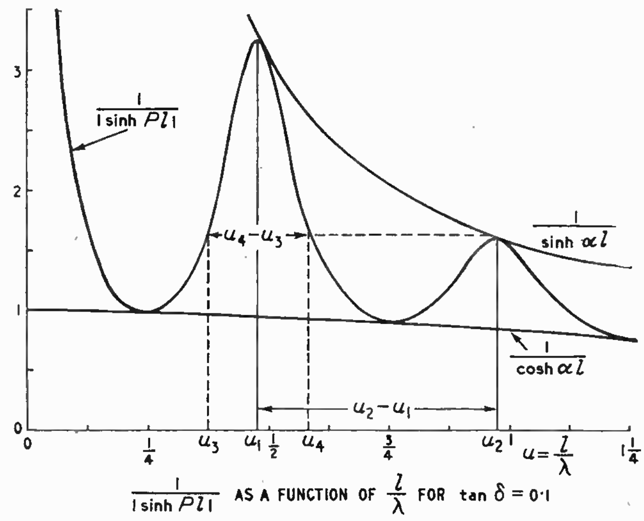


Fig. 1. *Variation of current in terminating disc for case of $Z_R = 0$, $\tan\delta = 0\cdot 1$ as a function of length in wavelengths*

first peak is invariably sharp, little error is introduced by the assumption that the width of the first peak is a linear function of tan δ. This may be represented by the statement

$$\left.\begin{aligned} u_3 &= \tfrac{1}{2} - E\tan\delta \\ u_4 &= \tfrac{1}{2} + E\tan\delta \end{aligned}\right\} \text{where } E \text{ is a constant}$$

whence $u_4 - u_3$ = width of peak $= 2E\tan\delta = b/\lambda$ where b is the width in cm. We now substitute the value $(\frac{1}{2} + E\tan\delta)$ into the expression for $|I|$ and equate this to the same expression with u_2 substituted for u. Taking $u_1 = \frac{1}{2}$, and using the expression for $u_2 - u_1$, from Equ. (2), we obtain the value for u_2 as $u_2 = 1 - z$, where $z = (R\tan\delta - X)/2\pi Z_0$. The resulting algebra is tedious; using the first two terms of the respective expansions for sinh, cosh, sin, and cos, and neglecting quantities small to the third order (for which purpose it should be noted that z itself is a small quantity), we finally obtain

$$\begin{aligned} &2E\tan\delta\,[2\pi Z_0(R\tan\delta - X)] \\ &\qquad - 4E^2\tan^2\delta\,.\,2\pi^2(R^2 + X^2 - Z_0^2)/2 \\ &= 2\pi RZ_0\tan\delta + \tfrac{1}{2}(R^2 + X^2 + Z_0^2)\,6\pi^2\tan^2\delta \\ &\quad + 4\pi z Z_0(X - R\tan\delta) - \tfrac{1}{2}(R^2 + X^2 - Z_0^2)\,8\pi^2 z^2 \end{aligned}$$

If we let $u_4 - u_3 = b/\lambda = 2E\tan\delta$ we have

$$\frac{4b}{\lambda}\left(\frac{1}{2} - \frac{d}{\lambda}\right) + \frac{b^2}{\lambda^2} - 3\tan^2\delta + 4\left(\frac{1}{2} - \frac{d}{\lambda}\right)^2 = \frac{2R\tan\delta}{\pi Z_0} \quad \ldots \ldots \ldots \ldots \ldots (4)$$

where we substitute for z from Equ. (3), and b is the width, in cm, measured on the first peak. This equation enables evaluation of R, provided that λ and tan δ are known, from the measured values of the width b and the distance d.

To obtain the value of tan δ let us assume that $R = X = 0$. Then Equ. (2) becomes

$$d_1/\lambda_0 = \tfrac{1}{2}(1 - \tan^2\delta) \quad \ldots \ldots \ldots \ldots (5)$$

Similarly in Equ. (4), to the second order of small quantities, we have

$$b_1/\lambda_0 = (3)^{1/2}\tan\delta \quad \ldots \ldots \ldots \ldots (6)$$

Combining Equs. (5) and (6) an expression for tan δ in terms of b_1/d_1 may be obtained; i.e., $b_1/d_1 = 2(3)^{1/2}\tan\delta$. Values of tan δ for a range of values of b_1/d_1 are given in Table 1. Experimentally the specimen disc is replaced by a good conductor, such as silvered copper, and the distance d_1 and the width b_1 are measured for the range of frequencies being used. If the conductors are of sufficient uniformity it is found that tan δ is a regular function of frequency, and need be determined once only for a given apparatus over a range of frequencies.

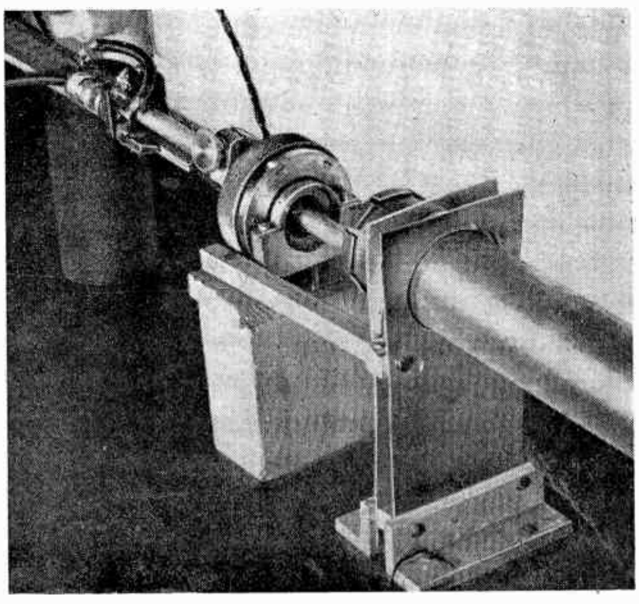
TABLE 1

$\frac{b_1}{d_1}$ $\times 10^{-4}$	tan δ $\times 10^{-4}$	$\frac{b_1}{d_1}$ $\times 10^{-4}$	tan δ $\times 10^{-4}$
1	0·2887	11	3·1757
2	0·5774	12	3·4644
3	0·8661	13	3·7531
4	1·1548	14	4·0218
5	1·4435	15	4·3305
6	1·7322	16	4·6392
7	2·0109	17	4·9079
8	2·3196	18	4·8206
9	2·4103	19	5·4853
10	2·8870	20	5·7740

The wavelength λ may be arrived at by considering the case of a loss-free line, terminated in a lossy disc. The contributions to the width b from the line losses and the disc losses can be shown to be additive. Thus for the case $\tan\delta = 0$ the width b_2, due to the disc alone, will be given by Equ. (4) with tan δ set to zero. This will readily be seen to lead to the condition

$$\lambda = 2d - b_2 \quad \ldots \ldots \ldots \ldots \ldots (7)$$

Fig. 2. *Input end of test apparatus showing slow-motion drive and vernier*



where $b_2 = b - b_1$. This equation yields the value of λ.

If it is desired to know the actual inductance, as opposed to the reactance of the disc, it is necessary to know the exact operating frequency. This is readily obtainable from the measurements from which $\tan \delta$ is determined. Combining Equs (5) and (6) we have

$$\lambda_0 = 2d_1 (1 + \tan^2\delta) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

for $\tan^2\delta \ll 1$, where λ_0 is the free-space wavelength of the frequency used.

It is pointed out that the inductive reactance X of the disc comprises two parts; its surface inductance, due to skin effect, and its 'external' inductance due to the circuit of which it forms part. If the specimen is non-magnetic, with a permeability of unity, Maxwell's equations show that the surface inductive reactance is equal to the surface resistance. Thus it is unnecessary to determine X, the total inductive reactance, explicitly. In the case of a ferromagnetic specimen, however, the permeability must be treated as complex, so that the relative permeability is given by

$$\mu_r = \mu_1 - j\mu_2$$

Substitution of this in the relevant skin-effect equation leads to

$$R_s + j\omega L_s = 1/2\pi \, [j\omega\rho\mu_0 (\mu_1 - j\mu_2)]^{\frac{1}{2}} \log_e (b/a)$$

where ρ = resistivity

b = inner radius of outer conductor

a = outer radius of inner conductor

The explicit expressions for permeability are

$$\mu_2 = (R_s^2 - X_s^2)/K^2 \text{ and } \mu_1 = 2R_sX_s/K^2$$

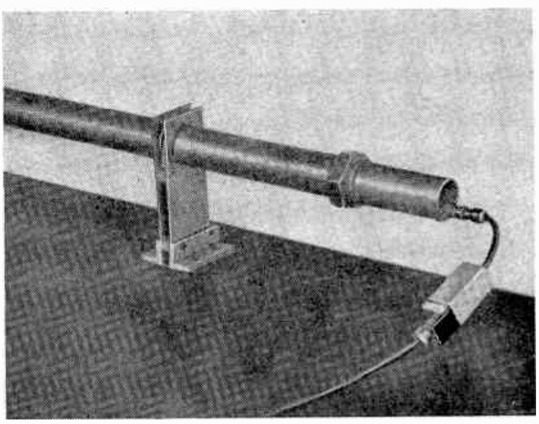
where $K = 1/2\pi \, (\omega\mu_0\rho)^{\frac{1}{2}} \log_e (b/a)$

To determine X_s it is necessary to evaluate the external inductive reactance X_e, which may be done by a series of measurements on a non-ferromagnetic disc and using $X_e = X - X_s$. The value of the external inductance L_e is, of course, a constant for a given apparatus.

Experimental Apparatus

From the point of view of screening and prevention of losses by radiation, it is preferable to use a coaxial system for the experimental apparatus. The experiments on nickel were over a frequency range of 200 to 450 Mc/s, for which purpose the apparatus shown in Figs. 2 and 3 was built.

Fig. 3. *Output end of test apparatus showing detector and r.f. filter*



It is necessary to consider whether the transmission-line equations, solved in the previous section, are applicable to the coaxial arrangement, in view of the fact that it may also be regarded as a coaxial resonant chamber. This has been fully treated by Harris[2], who considers a 'general-principle wave' in the system, conforming with the guiding conductors. Orthogonal curvilinear co-ordinates are used and the electromagnetic-field equations are solved in terms of these co-ordinates. He then shows that for a wave-front which bends slightly (i.e., deviates from planarity by only a small amount), the transmission-line equations are a good approximation. This would be the case for lossy conductors. For loss-free conductors, the wave-front is plane and the standard transmission-line equations follow rigorously from the form of Maxwell's equations in which the electric field is radial in cylindrical co-ordinates.

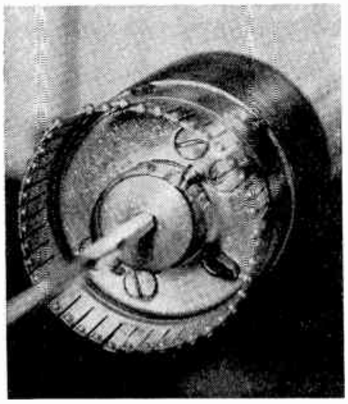


Fig. 4. *Detail of piston*

The experimental apparatus was made to be virtually loss-free by using a silvered-copper tube as the outer conductor and a solid silver rod, 0·13 in. diameter, as the inner conductor. It was found preferable to use a solid silver rod, as the constant abrasion of the piston contacts was found to wear off the silver plating when using a silvered-copper conductor. The design of the contacts between the short-circuiting piston and the inner conductor caused some trouble, the most satisfactory form found comprising silver brushes mounted after the style of commutator brushes in an electric motor. The detail of the piston is shown in Fig. 4, in which is also shown the input loop for the signal. The piston is actuated by a hollow tube through which is run the coaxial cable carrying the input signal. The end of the tube (remote from the piston) carries a vernier scale and moves over a rigidly-mounted metre rule. It was found essential to have a slow-motion drive for the piston, which can be seen in Fig. 2.

A pick-up loop similar to the input loop collects a signal proportional to the current on the terminating disc on which it is mounted. As a detector it was found that an ex-government type of silicon crystal diode could conveniently be mounted inside a Belling-Lee coaxial plug. This is connected directly to an r.f. filter, whence the d.c. goes to a galvanometer.

If the coupling between input and line or line and output is too tight, the wavelength and positions of the peaks on the line are affected. For a detector with a linear characteristic, this coupling should be adjusted until the second peak is not more than half the height of the first. Further reduction of the coupling will not improve accuracy, but it should be noted that if the detector had a pure square-law characteristic the ratio between first and second peak heights would not be less than 4 : 1.

The terminating disc is maintained in place by the tension of the inner conductor, and must make good contact with the rim of the outer conductor. In the apparatus shown it was desired to heat the specimen disc, so the whole of the last six inches of the line was made removable to avoid having to handle the hot disc.

A Marconi Instruments signal generator type TF801B was used as a signal source, and drove a lecher-line amplifier employing a Mullard QQVO-3/20A double tetrode with internal neutralizing. The valve was run at a standing current of 100 mA and anode voltage of 350 V. Pick-up from the amplifier was by means of a coupling loop from which up to 8 mA of r.f. current could be obtained. The coaxial feeder carrying the signal to the line was stub-matched at the amplifier end.

The presence of harmonics in the signal supplied to the line seriously affects the accuracy of the results. It was found, with the above arrangement of a push-pull amplifier, that the harmonic content was negligibly small.

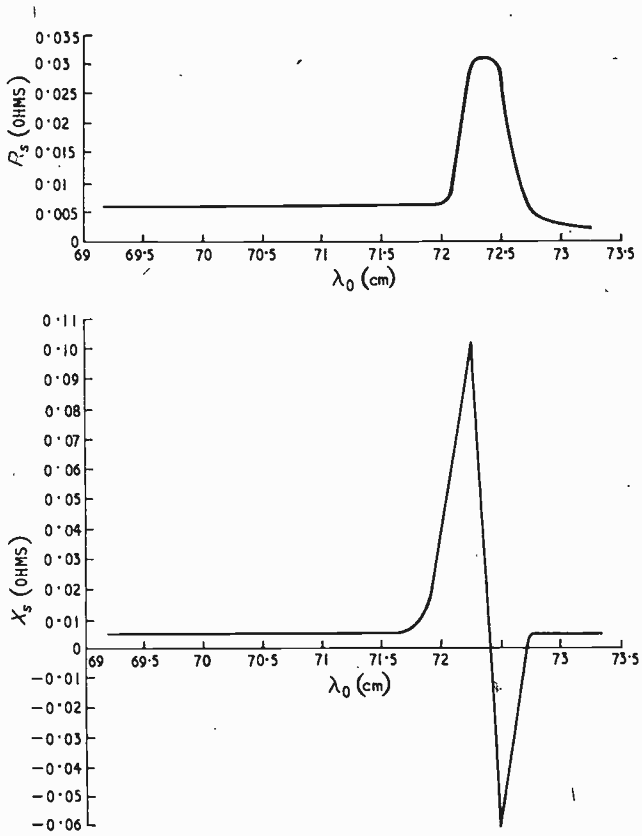


Fig. 5. *Variation of surface resistance and reactance, as a function of free-space wavelength, for pure nickel at 20 °C*

Results

The variation of surface inductance and resistance of pure nickel with free-space wavelength over the range 410 to 435 Mc/s is shown in Fig. 5, while Fig. 6 shows the values for real and imaginary parts of permeability derived from these results. It should be emphasized that the permeability calculated in this way has a purely formal significance. The value and phase of B (the induction) with respect to H (the applied alternating field) will vary from point to point in the specimen. The permeability plotted in Fig. 6 is, in effect, defined after the manner suggested by Kittel[3], being a parameter such that

$$Z\,(\mu,\,\omega)_{\text{calc.}} = Z\,(\omega)_{\text{expt.}}$$

The deviations in both Figs. 5 and 6 in the region of 417 Mc/s were, in fact, the primary object of the measurements described here. Similar deviations in nickel-iron alloys have already been reported[4], found with a different method of measurement. The disc-type specimen used in the present apparatus was chosen to facilitate varying its temperature, which was done by blowing hot air over it. The frequency at which the deviations in the permeability–wavelength curves are obtained varies rapidly with temperature. The physical mechanism accounting for the effect is to be discussed elsewhere, but sufficient evidence has been obtained to ascribe it to electron spin-resonance in the magneto-crystalline anisotropy field.

For purposes of comparison, the d.c. resistance of the

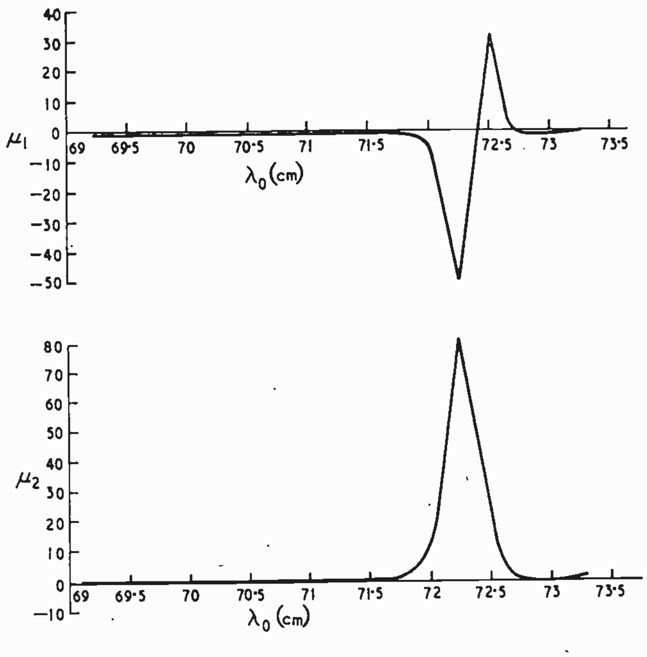


Fig. 6. *Variation of real and imaginary parts of relative permeability of pure nickel at 20 °C, as a function of free-space wavelength*

disc is calculated as $1{\cdot}97 \times 10^{-5}$ ohm, which implies a skin depth of $8{\cdot}79 \times 10^{-6}$ metre.

Conclusion

While the present experiments have been oriented in the direction of research in solid-state physics, it is felt that the re-arrangement of the transmission-line theory, and the experimental technique based upon it, should be of value in a number of problems in the v.h.f. field. Microwave techniques for similar measurements are well-developed and documented, but there has been no such interest hitherto in the v.h.f. range, which is becoming increasingly important in the communications field.

Acknowledgements

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Currents on Strip Aerials

By T. B. A. Senior*

SUMMARY. *An exact expression is obtained for the longitudinal distribution of current excited on a perfectly-conducting strip by a normally-incident plane wave, and computations are carried out for quarter and half-wave aerials. By using the known transverse variation of the current on strips of small width, the complete surface distribution is determined, leading to an expression for the total current carried by the aerial. This is compared with the current distribution for a thin wire, but little agreement is found. Some reasons for the differences are given.*

It is frequently assumed that the currents excited on a strip aerial have a longitudinal distribution which is similar to that for a thin wire. As a result, the variation of the current as a function of position can be closely represented by a cosine term. This appears a reasonable assumption for narrow strips when the incident field is 'edge-on', but at normal incidence the analogy with the wire is not quite so obvious, suggesting that further consideration be given to this case.

In the following, attention is confined to an idealized aerial consisting of a perfectly-conducting infinitely-thin strip of arbitrary length and width. As such, the strip can be likened to some types of indoor TV aerials. The excitation is by means of a normally-incident plane wave and no account is taken of aerial connections or impedance losses.

It is then shown that the longitudinal distribution of the surface current on a strip of length $2a$ can be deduced from the transverse distribution on a strip of width $2a$ and infinite length. The latter case can be expressed as a series of Mathieu functions, and computations are carried out for $a = \lambda/8$ and $a = \lambda/4$, corresponding to quarter and half-wave aerials.

For strips of relatively small width the total current distribution can be obtained using the known variation across the width of the strip, and this is done in the next section; the resulting distribution is compared with that for a thin wire. If the length of the strip is not large compared with the wavelength, the two distributions do not agree, and cannot be brought into agreement whatever the radius of the wire. Although the differences are not necessarily significant, it is clear that for *practical* strip aerials of small length, the new distribution will represent a better initial approximation in any iterative scheme for finding the true current.

The Method

An expression for the current is most easily obtained by considering the current which would be induced in an infinite strip by an incident plane wave.

In terms of Cartesian co-ordinates x, y, z the infinite strip will be assumed to occupy the region $-a<x<a$, $-\infty<z<\infty$ of the plane $y = 0$. A plane wave is incident in the direction of the negative y axis and, if its magnetic vector is entirely in the z direction, then

$$\mathbf{H}^i = (0, 0, e^{-iky}), \qquad \qquad (1)$$

where the affix 'i' is used to denote the incident field and $k = 2\pi/\lambda$ is the propagation constant. M.k.s. units

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are employed and a time factor $e^{-i\omega t}$ is suppressed throughout.

From symmetry considerations it is clear that the magnetic vector in the scattered field must also be confined to the z direction and, since the induced current is determined by the component of the total magnetic field parallel to the surface of the strip, the current vector $\mathbf{I}$ must be entirely in the x direction. This implies that the only current flow is across the strip. Moreover, the fact that the whole problem has two-dimensional symmetry shows that the current is the same at all points along the length of the strip and, hence, $\mathbf{I}$ can only be a function of x. Since $\mathbf{I} = \mathbf{n} \wedge \mathbf{H}$, where $\mathbf{n}$ is a unit vector normal to the surface, we now have

$$\mathbf{I} = (I, 0, 0)$$

and on the upper surface ($y = +0$), $I = I_+(x)$ with

$$I_+(x) = [H_z]_{y=+0} = 1 + [H_z^s]_{y=+0} \quad \ldots \quad (2)$$

On the lower surface of the strip the sign of I is reversed.

A strip aerial of the type under consideration can be obtained by chopping up an infinite strip with cuts parallel to the x axis, thereby producing a rectangular surface of length $2a$ and width $2d$ (say). In general, d will be small compared with a. The new edges which are formed in this way cannot generate any transverse current, nor can they affect the x dependence of the longitudinal current which was present on the larger surface. This fact enables us to identify the current on the aerial with that on the infinite strip. The only difference is that the actual aerial current is a function of z as well as x, and this other dependence has been discussed by, for example, Moullin & Phillips[1]. The variation with z is required in any calculation of the total (integrated) current carried by the aerial and will be referred to again later.

The Analysis

It is now a simple matter to determine the aerial current using the exact expression for the current excited on an infinite strip by a normally-incident plane wave.

From McLachlan[2] we have

$$[H_z^s]_{y=+0} = 2ib \sum_{n=0}^{\infty} B_1^{(2n+1)} \frac{Ne_{2n+1}^{(1)}(0)}{Ne_{2n+1}^{(1)\prime}(0)} se_{2n+1}(\eta)$$

where $b = \dfrac{ka}{2}$ and $\eta = \cos^{-1}\dfrac{x}{a}$. The Mathieu function coefficients are denoted by $B_{2m+1}^{(2n+1)}$ and in terms of these

$$se_{2n+1}(\eta) = \sum_{m=0}^{\infty} B_{2m+1}^{(2n+1)} \sin(2m+1)\eta,$$

$$\frac{Ne_{2n+1}^{(1)}(0)}{Ne_{2n+1}^{(1)\prime}(0)} = -\frac{\sum_{m=0}^{\infty}(-1)^m B_{2m+1}^{(2n+1)}}{\sum_{m=0}^{\infty}(-1)^m(2m+1)B_{2m+1}^{(2n+1)} + \pi b^2 \sum_{m=0}^{\infty}(-1)^m B_{2m+1}^{(2n+1)}(P_m - iQ_m)}$$

where P_m and Q_m are given by the Bessel function formulae

$$P_m = J_m(b)Y_m(b) + J_{m+1}(b)Y_{m+1}(b) - \frac{2m+1}{b} J_{m+1}(b)Y_m(b), \text{ and}$$

$$Q_m = \{J_m(b)\}^2 + \{J_{m+1}(b)\}^2 - \frac{2m+1}{b} J_{m+1}(b)\,J_m(b).$$

Hence,

$$I_+(x) = 1 + 2ib \sum_{n=0}^{\infty} B_1^{(2n+1)} \frac{Ne_{2n+1}^{(1)}(0)}{Ne_{2n+1}^{(1)\prime}(0)} se_{2n+1}(\eta). \quad (3)$$

For the lower surface ($y = -0$), the sign of I has to be reversed, but since η must also be replaced by $2\pi - \eta$, then

$$I_-(x) = I_+(x) - 2. \quad \ldots \quad (4)$$

The coefficients $B_{2m+1}^{(2n+1)}$ are, of course, functions of b and their values have been tabulated by the Computation Laboratory, National Bureau of Standards[3]. For large b, the series for H_z^s is only slowly convergent but, when b is less than unity, the convergence is sufficiently rapid for the series to be cut off after the first few terms.

Two examples will be considered, corresponding to quarter and half-wave aerials ($a = \lambda/8$ and $\lambda/4$ respectively). Using tabulated values of $B_{2m+1}^{(2n+1)}$ and of the Bessel functions (the latter being supplemented, where necessary, by direct calculation of $J_n(b)$ and $Y_n(b)$ from their series expansions), the following results are obtained:

$a = \lambda/8$:

$$I_+(x) = 1 - (0{\cdot}32212 - i\,1{\cdot}01278)\sin\eta + (0{\cdot}00636 - i\,0{\cdot}01485)\sin 3\eta - (0{\cdot}00005 - i\,0{\cdot}00021)\sin 5\eta \quad (5)$$

$a = \lambda/4$:

$$I_+(x) = 1 + (1{\cdot}80385 + i\,0{\cdot}88454)\sin\eta - (0{\cdot}14997 + i\,0{\cdot}03026)\sin 3\eta + (0{\cdot}00396 + i\,0{\cdot}00061)\sin 5\eta - (0{\cdot}00005 + i\,0{\cdot}00002)\sin 7\eta \quad \ldots \quad (6)$$

The coefficients of the higher trigonometrical functions are zero to the first 5 places of decimals.

The amplitudes and phases of $I_+(x)$ are plotted at functions of x in Figs. 2 and 3 respectively. Taking first the quarter-wave aerial, it is seen that the phase behaves

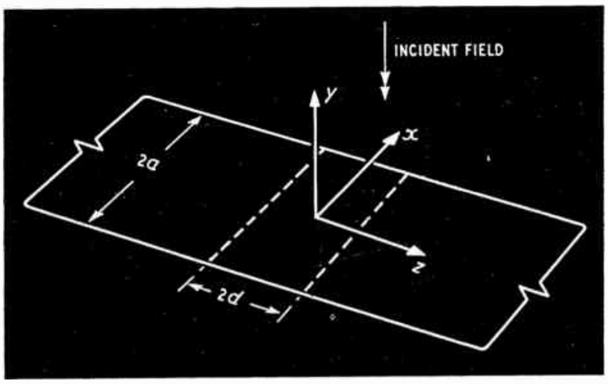

Fig. 1. *The co-ordinate system*

in a perfectly regular manner, increasing monotonically as x increases from 0 to a, but the amplitude curve shows a minimum at about $x = 0{\cdot}95a$. This would be explained if there were an appreciable build up of charge at the ends, giving rise to a capacitive effect. The aerial would then resonate at a wavelength which is greater than that predicted by its physical dimensions.

No such minimum is found in the amplitude curve for $a = \lambda/4$ and this can be attributed to the fact that the first harmonic (which is zero at the ends) now dominates the current distribution. Since the constant current is mainly responsible for any charge appearing at the ends, the capacitive effect is no longer important.

Comparison with Thin Wire Currents

The discussion so far has been devoted to the current $I_+(x)$ which would be measured if a probe were placed on the upper surface of the aerial. Such measurements are entirely feasible in spite of the experimental difficulties involved, and some results obtained with rectangular and triangular surfaces have been described by Hey & Senior[4].

For the lower surface of the aerial, the expression for the current differs from that of Equ. (3) in the sign of the constant term. This discontinuity between the currents on the illuminated and non-illuminated sides of the surface is to be expected and is the same as that occurring in the case of a half-plane (see, for example, Clemmow[5]).

To determine the total current carried by the aerial it is necessary to multiply each of the surface currents by a factor which takes into account the variation across the width of the strip. This dependence has been considered at length by Moullin & Phillips[1] and they have shown that for narrow strips ($kd \ll 1$) a close approximation to the transverse distribution is provided by the function

$$(1 - z^2/d^2)^{-1/2}.$$

This is precisely the z dependence which would be arrived at by a study of the current near to the edge of a half-plane, and the singularities at $z = \pm d$ are those which are required in order to satisfy the edge conditions in diffraction theory[6].

The distribution of current over the upper surface of the strip can now be represented as a function of x and z by

$$\frac{I_+(x)}{(1 - z^2/d^2)^{1/2}}$$

and to obtain the total current flowing in the x direction it is only necessary to integrate with respect to z from $z = -d$ to $z = d$. Since

$$\int_{-d}^{d} (1 - z^2/d^2)^{-1/2}\, dz = \pi d,$$

the current carried by both the upper and lower surfaces is

$$\begin{aligned} I_{total}(x) &= \pi d\,\{I_+(x) + I_-(x)\} \\ &= 2\pi d\,\{I_+(x) - 1\}, \qquad \text{(8)} \end{aligned}$$

and the magnitude of the current is plotted in Fig. 4 for $a = \lambda/8$ and $a = \lambda/4$. It will be observed that the curve for $a = \lambda/8$ does not show the minimum which occurred with $I_+(x)$.

The above expression for $I_{total}(x)$ is certainly valid when d is small and this fact suggests that the distribution should be comparable with that for a thin wire of suitably chosen radius. The longitudinal distribution of current on a thin wire has been considered by King & Harrison[7] and, if the source of excitation is a normally-incident plane wave, a general formula for the total current is

$$I_{total}(x) = A\left\{\cos kx - \cos ka + \frac{1}{\Omega} f(x)\right\} \qquad \text{(9)}$$

where $f(x)$ is a complicated function involving sine and cosine integrals and A is a normalizing constant. The parameter Ω is defined by the equation

$$\Omega = 2 \log \frac{2a}{r},$$

where r is the radius of the wire, and in the derivation of Equ. (9) it is assumed that $\Omega \gg 1$.

It is immediately apparent that the current distributions for a strip and a wire are entirely different in

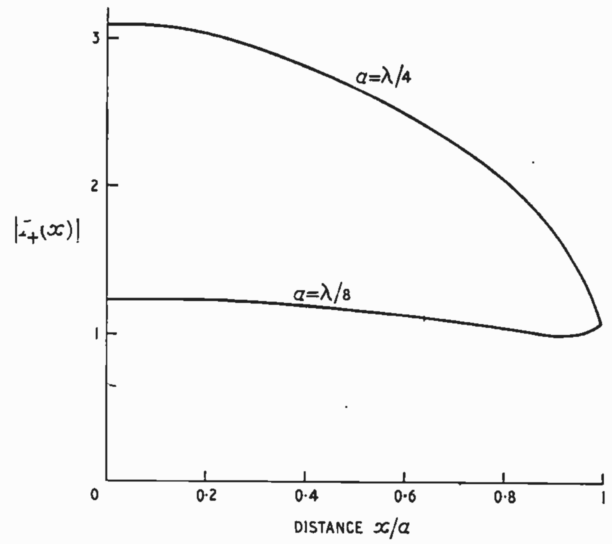

Fig. 2. *Amplitude of surface current $I_+(x)$ on quarter-wave ($a = \lambda/8$) and half-wave ($a = \lambda/4$) strips.*

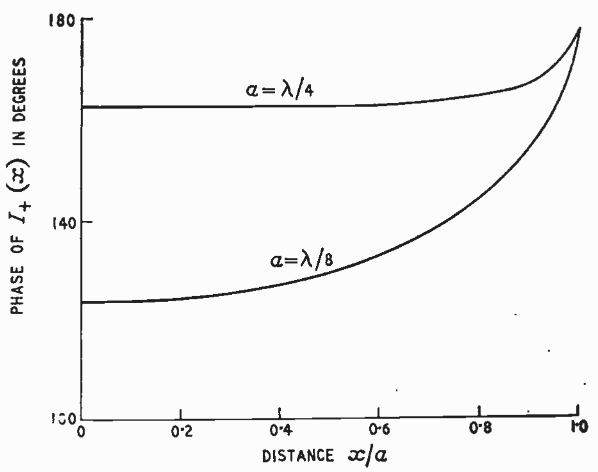

Fig. 3. *Phase of surface current $I_+(x)$ on quarter-wave ($a = \lambda/8$) and half-wave ($a = \lambda/4$) strips*

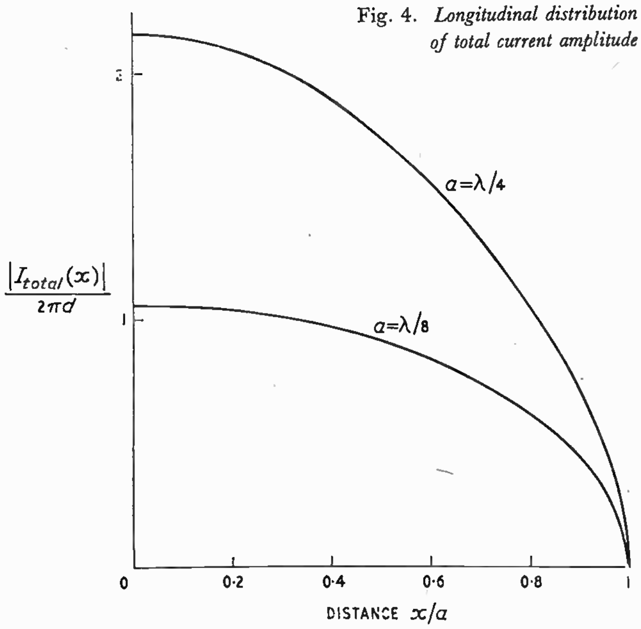

Fig. 4. *Longitudinal distribution of total current amplitude*

character, but this does not rule out the possibility of finding an alternative expression for the strip current which will bring the two into agreement. The natural expansion for $I_+(x)$, and hence $I_{total}(x)$, is based upon the functions $\sin(2n+1)\eta$, $n = 0, 1, 2, \ldots$, where $\eta = \cos^{-1}\frac{x}{a}$, and since

$$\sin\eta = \left(1 - \frac{x^2}{a^2}\right)^{1/2},$$

$$\sin 3\eta = 3\left(1 - \frac{x^2}{a^2}\right)^{1/2} - 4\left(1 - \frac{x^2}{a^2}\right)^{3/2}, \text{ etc.,}$$

the series can also be written in terms of the functions $\left(1 - \frac{x^2}{a^2}\right)^{n+1/2}$, $n = 0, 1, 2, \ldots$ Both of these expansions are rapidly convergent for values of ka of order unity, and a reasonable approximation to the current distribution can be obtained by neglecting all harmonics above the first. Thus, for $a = \lambda/8$,

$$I_{total}(x) \approx -2\pi d\,(0{\cdot}32212 - i\,1{\cdot}01278)\left(1 - \frac{x^2}{a^2}\right)^{1/2} \quad \ldots \quad \ldots (10)$$

and for $a = \lambda/4$,

$$I_{total}(x) \approx 2\pi d\,(1{\cdot}80385 + i\,0{\cdot}88454)\left(1 - \frac{x^2}{a^2}\right)^{1/2}. \quad \ldots \quad \ldots (11)$$

The function $\sin\eta$ can also be expressed in a Fourier series of the form

$$\sin\eta = \sum_{m=0}^{\infty} \frac{2}{2m+1}\, J_2\left\{\left(m + \frac{1}{2}\right)\pi\right\} \cos(2m+1)kx,$$

where J_2 is the second order Bessel function, and similarly

$$\sin 3\eta = \sum_{m=0}^{\infty} \frac{6}{2m+1}\left\{1 - \frac{8}{(2m+1)\pi}\right\} \times J_2\left\{\left(m + \frac{1}{2}\right)\pi\right\} \cos(2m+1)kx.$$

Such expansions, however, converge only slowly and if they are inserted into the equation for the total current, a large number of terms are required in order to reproduce the accuracy represented by the single terms in, for example, Equs. (10) and (11). The first term involving $\cos kx$ in no sense dominates the series when ka is small and hence, if an attempt is made to write the current distribution for a strip in a form analogous to that of Equ. (9), the correction term corresponding to $\frac{1}{\Omega} f(x)$ will be at least as important as $\cos kx$. As a result, the differences between the current distributions for a strip and a wire must be regarded as fundamental and, indeed, if a numerical comparison is made, no practicable value of Ω exists which will bring them into even an approximate agreement.

Conclusions

In the preceding section is has been shown that the longitudinal distribution of current on a strip of finite length a can be derived from the exact Mathieu function expansion for the current on an infinite strip. The resulting expression is valid for all values of a.

If the width of the strip is small, the known variation of the current in the transverse direction may be used to predict the entire surface distribution, leading to an expression for the total current $I_{total}(x)$ carried by the strip. For small ka it is found that

$$I_{total}(x) \approx 2\pi d\,\alpha\left(1 - \frac{x^2}{a^2}\right)^{1/2}, \quad \ldots \quad \ldots \quad \ldots (12)$$

where α is a complex numerical constant depending on the length a and given by a rapidly convergent Mathieu function series. In reality, the term on the right of Equ. (12) is the first in a series of ascending powers of $\left(1 - \frac{x^2}{a^2}\right)$, but for $ka < 1$ the coefficients of the higher powers are negligible.

Although it is freely admitted that the discussion has been confined to the most idealized type of strip aerial, Equ. (12) does suggest that any iterative scheme for finding the current on a practical aerial should be based on an initial approximation of the form $\left(1 - \frac{x^2}{a^2}\right)^{1/2}$. Providing ka is not large compared with unity, this will certainly be better than the usual assumption of a cosine variation, since the end effects are a major factor in determining the nature of the distribution. When $ka \gg 1$, however, the end effects are less important and in this case it may well be that a cosine dependence is a more accurate approximation.

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Divide and Rule

When it is necessary to compute the value of an algebraic expression, the obvious method of substituting a numerical value for the independent variable, say x, and for all its powers is by no means the only method available, and in many cases it is not the best procedure. Suppose, for example, that we require the value of

$$f_1(x) = x^6 - 3x^5 + 9x^4 - 27x^3 + 40x^2 - 60x + 100 \quad (1)$$

for $x = 2$ and for a number of values in the immediate neighbourhood of 2, such as 2·1 or 2·0345. Direct substitution is quite easy when x is exactly equal to a single-figure integer like 2, or when x is -2, 20, -300, 0·04, -5×10^{-4}, etc. It is still possible when x has a value like 2·1 which has two significant figures, since books of tables of logarithms, circular functions, etc., often include powers of two-figure integers up to the sixth, but there may be some loss of accuracy if these powers are only given to say five significant figures. If however x has a value like 2·0345 with several significant figures, or if x is complex, direct substitution becomes intolerably and unnecessarily tedious. It may be necessary to retain a large number of significant figures in the working because the resulting value of $f_1(x)$ is small in comparison with the value of several of the constituent terms.

In these cases it is much easier to use the Remainder Theorem. This states that the value of a function like $f_1(x)$ in Equ. (1) above when $x = x_0$ is equal to the remainder obtained when $f_1(x)$ is divided by $(x - x_0)$. Applying this to the $f_1(x)$ defined by Equ. (1) with x_0 equal to 2, we obtain

$$\begin{array}{rl}
 & x^5 - x^4 + 7x^3 - 13x^2 + 14x - 32 \\
x - 2\,) & x^6 - 3x^5 + 9x^4 - 27x^3 + 40x^3 - 60x + 100\,(\\
 & x^6 - 2x^5 \\
 & \quad - x^5 + 9x^4 \\
 & \quad - x^5 + 2x^4 \\
 & \qquad 7x^4 - 27x^3 \\
 & \qquad 7x^4 - 14x^3 \\
 & \qquad\quad - 13x^3 + 40x^2 \\
 & \qquad\quad - 13x^3 + 26x^2 \\
 & \qquad\qquad 14x^2 - 60x \\
 & \qquad\qquad 14x^2 - 28x \\
 & \qquad\qquad\quad - 32x + 100 \\
 & \qquad\qquad\quad - 32x + 64 \\
 & \qquad\qquad \text{Remainder} + 36
\end{array}$$

and it is easily verified that if x is replaced by 2 on the right-hand side of Equ. (1), the right-hand side, usually denoted by $f_1(2)$, does reduce to 36. Now the Remainder Theorem becomes obvious if we summarize what we have found out by performing the division and write

$$f_1(x) \equiv (x - 2)\,(x^5 - x^4 + 7x^3 - 13x^2 + 14x - 32) + 36 \quad \ldots \quad (2)$$

The three bars after $f_1(x)$ in Equ. (2) are deliberately put there to emphasize that $f_1(x)$ is always equal to the expression on the right-hand side, no matter what the value of x may be, positive, negative, integral, fractional, real or complex—Equ. (2) should, in fact, be called Identity (2), and any value whatever may be substituted for x on the two sides of it. Equ. (1) should likewise be called Identity 1. In particular, if 2 is substituted for x, the factor $(x - 2)$ on the right-hand side becomes zero, and therefore $f_1(x)$, or $f_1(2)$, reduces to the single term 36 as already mentioned. Similarly, if we require $f_1(2{\cdot}1)$, we can find the remainder when $f_1(x)$, given by Identity (1), is divided by $x - 2{\cdot}1$; for f_1 (2·0345) we divide by $x - 2{\cdot}0345$, and for $f_1(-1{\cdot}7)$ we divide by $x + 1{\cdot}7$. Alternatively, if all the values in which we are interested are in the immediate neighbourhood of 2, we can express $f_1(x)$ as another function $f_2(y)$ of $y = x - 2$. From the identity (2) above we have the first step in this transformation; we can write

$$f_1(x) \equiv f_2(y) \equiv y(x^5 - x^4 + 7x^3 - 13x^2 + 14x - 32) + 36 \quad \ldots \quad (3)$$

but unfortunately Identity (3) is a hybrid which contains both y and x. The next step towards eliminating x completely is to take the expression

$$f_2(x) = x^5 - x^4 + 7x^3 - 13x^2 + 14x - 32 \quad \ldots \quad (4)$$

which was the quotient in the original division, and divide it also by $y = x - 2$. This gives

$$\begin{array}{rl}
 & x^4 + x^3 + 9x^2 + 5x + 24 \\
x - 2\,) & x^5 - x^4 + 7x^3 - 13x^2 + 14x - 32\,(\\
 & x^5 - 2x^4 \\
 & \quad x^4 + 7x^3 \\
 & \quad x^4 - 2x^3 \\
 & \qquad 9x^3 - 13x^2 \\
 & \qquad 9x^3 - 18x^2 \\
 & \qquad\quad 5x^2 + 14x \\
 & \qquad\quad 5x^2 - 10x \\
 & \qquad\qquad 24x - 32 \\
 & \qquad\qquad 24x - 48 \\
 & \qquad \text{Remainder} + 16
\end{array}$$

Summarizing this division, we obtain the identity

$$f_2(x) \equiv y\,(x^4 + x^3 + 9x^2 + 5x + 24) + 16 \quad \ldots \quad (5)$$

If we now substitute this back into Identity (3), it takes the form

$$f_1(x) \equiv f_2(y) \equiv y^2(x^4 + x^3 + 9x^2 + 5x + 24) + 16y + 36 \quad \ldots \quad (6)$$

Continuing this process of division by $(x - 2)$, but recording only the summarized form of the result, analogous to Identities (2) and (5), we have

$$\left.\begin{array}{r}
x^4+x^3+9x^2+5x+24 \equiv y(x^3+3x^2+15x+35)+94 \\
x^3+3x^2+15x+35 \equiv y(x^2+5x+25)+85 \\
x^2+5x+25 \equiv y(x+7)+39 \\
x+7 \equiv y+9
\end{array}\right\} \quad (7)$$

Substituting from Identities (7) into Identity (6), we have

$$f_1(x) \equiv f_2(y) \equiv y^6 + 9y^5 + 39y^4 + 85y^3 + 94y^2 + 16y + 36 \quad \ldots \quad (8)$$

and we have thus obtained an expression entirely in terms of $y = x - 2$. It is worthy of note that the coefficients in Identity (8), reading from right to left, are the successive remainders in the divisions leading to Identities (2), (5) and (7).

If now we require the value of $f_1(x)$ when $x = 2{\cdot}1$ or $x = 2{\cdot}0345$ we may, as already mentioned, divide $f_1(x)$ by $(x - 2{\cdot}1)$ or by $(x - 2{\cdot}0345)$, or we can instead evaluate $f_2(y)$ from Identity (8) when y is 0·1 or 0·0345. Since 0·1 and 0·0345 are both small numbers, we may be able to neglect some of the higher powers of y in Identity (8), whereas we cannot neglect the corresponding powers of x in the original Identity (1) defining $f_1(x)$. Thus, if x is 2·1 and y is 0·1,

$$f_1(2{\cdot}1) = f_2(0{\cdot}1) = 0{\cdot}000001 + 9 \times 0{\cdot}00001 + 39 \times 0{\cdot}0001 + 85 \times 0{\cdot}001 + 94 \times 0{\cdot}01 + 16 \times 0{\cdot}1 + 36 = 38{\cdot}628991 \quad \ldots \quad (9)$$

and if the first three terms are omitted, we obtain the approximate value 38·625, which is only in error in the third decimal place. To obtain $f_1(2{\cdot}0345)$ or $f_2(0{\cdot}0345)$, it is immaterial whether we divide the right-hand side of Identity (1) by $x - 2{\cdot}0345$ or that of Identity (8) by $y - 0{\cdot}0345$ except that, in the latter case, we are likely to be able to omit the first three, if not four, terms. If therefore we require one isolated value like 2·0345 for x, it is probably best to divide by $x - 2{\cdot}0345$ using Identity (1), but if we require several awkward values all in the neighbourhood of 2, it will pay to form Identity (8) and omit as many terms as possible.

We now have to consider how to extend this process so that we can evaluate say $f_1(1 + j)$. Now the Remainder Theorem is still true even if complex numbers are involved; that is to say, if we could conveniently divide $f(x)$ given by Identity (1) by $(x - 1 - j)$, the remainder would still be $f_1(1 + j)$. It is, however, much easier to perform the equivalent of division by $(x - 1 - j)$ indirectly, by means of a simple trick, so that only division by a quadratic expression in x with real coefficients is actually carried out. To obtain this quadratic expression in x, we simply multiply $(x - 1 - j)$ by the factor $(x - 1 + j)$ in which j is replaced by $-j$, but x is left alone. We then have

$$(x - 1 - j)(x - 1 + j) = (x - 1)^2 + 1 = x^2 - 2x + 2 \quad \ldots \quad (10)$$

Now divide $f_1(x)$ by $x^2 - 2x + 2$, as follows:

$$\begin{array}{r|l}
 & x^4 - x^3 + 5x^2 - 15x \\
\hline
x^2 - 2x + 2 & x^6 - 3x^5 + 9x^4 - 27x^3 + 40x^2 - 60x + 100 \\
 & x^6 - 2x^5 + 2x^4 \\
 & \quad - x^5 + 7x^4 - 27x^3 \\
 & \quad - x^5 + 2x^4 - 2x^3 \\
 & \qquad 5x^4 - 25x^3 + 40x^2 \\
 & \qquad 5x^4 - 10x^3 + 10x^2 \\
 & \qquad\quad -15x^3 + 30x^2 - 60x \\
 & \qquad\quad -15x^3 + 30x^2 - 30x \\
 & \qquad\qquad \text{Remainder } -30x + 100
\end{array}$$

The summary of what we have thus found out, analogous to Identity (2), is

$$f_1(x) \equiv (x - 1 - j)(x - 1 + j)[x^4 - x^3 + 5x^2 - 15x] - 30x + 100 \quad \ldots \quad (11)$$

and again, (11) is an identity and not a mere equation; that is to say, it is true for all values of x including $(1 + j)$. If we put $(1 + j)$ instead of x, the first factor of the first term on the right-hand side of Identity (11) is zero; this sweeps away everything except the last two terms on the right-hand side, which came as the remainder after the division by $(x^2 - 2x + 2)$. We thus have that the value $f_1(1 + j)$ of $f_1(x)$ when x is put equal to $(1 + j)$ is given by

$$f_1(1 + j) = -30(1 + j) + 100 = 70 - 30j \quad \ldots \quad (12)$$

so that the complex number $(1 - j)$ has only had to be substituted into the remainder after the division by $(x^2 - 2x + 2)$; this remainder is at worst only linear in x whatever the degree of the original expression $f_1(x)$.

In Identity (8), we obtained an alternative expression $f_2(y)$ for $f_1(x)$, where y was an abbreviation for $x - 2$. Now, in this case it follows that

$$x = y + 2 \quad \ldots \quad (13)$$

We could therefore have obtained Identity (8) by substituting from Equ. (13) into Identity (1) defining $f_1(x)$, so as to eliminate all the x terms appearing in Identity (1) simultaneously. We would thus obtain

$$f_1(x) \equiv f_2(y) \equiv (y + 2)^6 - 3(y + 2)^5 + 9(y + 2)^5 - 27(y + 2)^3 + 40(y + 2)^2 - 60(y + 2) + 100 \quad \ldots \quad (14)$$

Now, in fact, Identity (14) must be the same as Identity (8), and it is often very useful, as a check on arithmetical accuracy, to be able to have two alternative ways of deriving an expression, which involve different numerical and algebraic operations. Identity (14) is only in a useful form provided that we can evaluate an expression like $(y + 2)^n$. This can be done by means of the 'Binomial Theorem' which will be the subject of our next article.

ELECTRONIC VALVE EXHIBITION

A private exhibition of electronic valves manufactured by the English Electric Valve Co. Ltd., will be held at the Kensington Palace Hotel, De Vere Gardens, London, W.8, from 18th March to 21st March 1959, inclusive.

An invitation is extended to all readers of *Electronic & Radio Engineer* to attend this exhibition which will be open from 3 to 8 p.m. on the first day and 10 a.m. to 8 p.m. on remaining days.

CONVENTION ON STEREOPHONY

The Radio and Telecommunication Section of the Institution of Electrical Engineers has arranged a convention on stereophonic sound recording, reproduction and broadcasting to be held at the Institution's building in Savoy Place, London, on 19th and 20th March 1959. The sessions of the convention will cover basic principles, stereophonic recording on film, tape and disc and stereophonic broadcasting techniques. Where appropriate there will be associated demonstrations of equipment.

The convention will be open to members and non-members of the I.E.E. and all wishing to attend will be required to register.

Further particulars and registration forms may be obtained from the Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

Network Synthesis

BALANCED ASYMMETRICAL RC TYPES

By J. T. Allanson*

SUMMARY. *A method is outlined for the synthesis of RC networks which are balanced with respect to earth but are asymmetrical, and it is shown that such networks may have a lower pad loss than the corresponding symmetrical lattice.*

The majority of existing methods of synthesizing RC networks lead to the development of three-terminal networks or to a balanced and symmetrical lattice. While it is desirable, for practical purposes, that a network should either be balanced or have one terminal common to both input and output circuits, the condition of symmetry may impose unnecessary restrictions on the performance of the network.

The method discussed leads to the realization of a required voltage ratio function by means of two asymmetrical balanced ladder networks, connected in parallel at one end and 'crossed' at the other, as shown in Fig. 10. The ladder structures are to be synthesized originally as three-terminal networks, and made balanced in the final stage of synthesis only. By this method, a network is produced by a comparatively simple technique and, although the number of components required may be greater than that needed if a symmetrical lattice is used, the pad loss will be less than that associated with such a lattice.

Properties of Symmetrical and Asymmetrical RC Networks

The properties of transfer functions associated with RC networks have been investigated by a number of authors[1–6], and the following is a brief summary of the main results. The voltage transfer function of

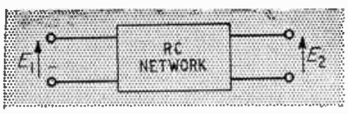

Fig. 1. *General RC network. The voltage transfer function is given by E_2/E_1*

a network is assumed to be given by Equ. (1), the voltages having the significance shown in Fig. 1.

$$\frac{E_2}{E_1} = \frac{N(p)}{D(p)} = \frac{a_m p^m + a_{m-1} p^{m-1} \ldots . + a_0}{b_n p^n + b_{n-1} p^{n-1} \ldots . + b_0} \qquad (1)$$

If the transfer function is to be realized by a RC network, the coefficients of both numerator and denominator must be real, and the roots of $D(p)$ must be simple and occur at negative real values of the complex frequency p. There may also be restrictions on the roots of $N(p)$ (i.e., on the zero of the transfer function), depending on the structure of the network to be used. In Table 1 a list is given of such restrictions and of the relationships permitted between m and n, for various types of network.

In addition to the properties listed above, there are restrictions[4,5] on the maximum value of the multiplying constant h obtainable with the various types of

TABLE 1

Type of network	Restrictions on roots of $N(p)$	Restrictions on m and n
General four-terminal	None	$m \leqslant n$
General three-terminal	Not positive real	$m \leqslant n$
Ladder	Only negative real	$m \leqslant n \leqslant m+s$ †
L-network	Only negative real	Either $m = n$ or $m = n-1$

† s is the number of stages of the ladder network.

network, and there are further restrictions on the location of the zeros of the L-network.

It will be noted that the general four-terminal network is the only one for which there are no restrictions on the zeros of the transfer function. Although Fialkow and Gerst[5] have developed a method for the synthesis of a general four-terminal network which normally results in a structure which is asymmetrical and unbalanced, all other methods of synthesis of four-terminal networks are limited to the production of a symmetrical lattice. For practical reasons it is desirable that four-terminal networks should be balanced, but there is often no practical advantage to be gained from symmetry and there may be disadvantages. This may be illustrated by reference to the method of Bower and Ordung[7] for the synthesis of a symmetrical lattice.

The voltage transfer function of the network of Fig. 2 is given in Equ. (2), where the admittance parameters are the short-circuit driving-point and transfer admittances[3].

$$\frac{E_2}{E_1} = \frac{-y_{12}}{y_{22} + Y_L} \qquad (2)$$

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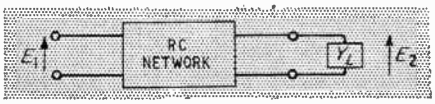

Fig. 2. *General RC network connected to load of admittance* Y_L

These admittances and y_{11} may be expanded[3] in the form shown in Equs (3), (4) and (5).

$$-y_{12} = a_\infty p + a_0 + p\sum_{1}^{j=k}\frac{a_j}{p+p_j} \quad \ldots \quad (3)$$

$$y_{22} = b_\infty p + b_0 + p\sum_{1}^{j=k}\frac{b_j}{p+p_j} \quad \ldots \quad (4)$$

$$y_{11} = c_\infty p + c_0 + p\sum_{1}^{j=k}\frac{c_j}{p+p_j} \quad \ldots \quad (5)$$

There are restrictions on the coefficients of y_{12} at the various poles, to the effect that $|a_j|^2 \leqslant b_j c_j$ for all j. Furthermore, $|a_\infty|$ must not be greater than the smaller of b_∞ and c_∞, and similarly $|a_0|$ must not be greater than the smaller of b_0 and c_0. If the network is symmetrical, then $y_{22} = y_{11}$, and the relationships between the residues reduce to

$|a_j| \leqslant b_j$ for all j, including $j = 0$ and $j = \infty$.

In the method of Bower and Ordung, the function E_2/E_1 is treated as in Equ. (2), and the simplest case is when $Y_L = 0$. Then

$$\frac{E_2}{E_1} = h\cdot\frac{N(p)}{D(p)} = \frac{-y_{12}}{y_{22}} \quad \ldots \quad (6)$$

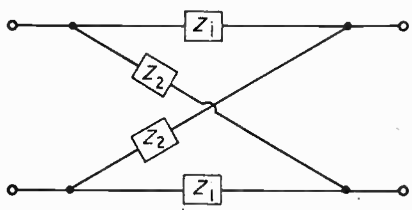

Fig. 3. *Symmetrical lattice section*

For the lattice of Fig. 3,

$$-y_{12} = (Y_1 - Y_2)/2 \quad \ldots \quad (7)$$

$$y_{11} = y_{22} = (Y_1 + Y_2)/2 \quad \ldots \quad (8)$$

By the creation of a suitable polynomial $G(p)$, y_{12} and y_{22} may be indentified as

$$-y_{12} = h\cdot\frac{N(p)}{G(p)} \quad \ldots \quad (9)$$

$$y_{22} = \frac{D(p)}{G(p)} \quad \ldots \quad (10)$$

The maximum value of h will then be determined by the necessity to satisfy the relationships between the coefficients of y_{12} and y_{22} discussed earlier; i.e., $|a_j| \leqslant |b_j|$. Bower and Ordung show that this maximum value may be calculated by evaluating $|N(p)/D(p)|$ for all negative real values of p, and obtaining the maximum of the minima of this modulus. This maximum is then $1/h$.

It would often be possible to obtain far higher values of h than this process will generally yield, by making the network asymmetrical. If the maximum value of the minima $|N(p)/D(p)|$ occurs for any value of p other than zero or infinity, then the conditions may be satisfied with a greater value of h by creating a network in which the coefficients of y_{11} (c_j) are made sufficiently large.

It is, however, only possible to make the lattice asymmetrical by making it unbalanced and, as argued above, this is undesirable in most cases. This article is concerned with outlining a method of synthesis which leads to a balanced but asymmetrical four-

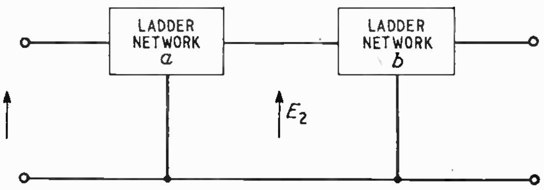

Fig. 4. *Unbalanced ladder networks (a) and (b) in series*

terminal RC network, having no restrictions on the location of the zeros of the voltage transfer function. The gain constant obtainable from the network will not be less than that obtainable from a symmetrical lattice, and may be considerably greater.

Outline of the Method of Synthesis

It may be deduced from Equ. (2) that the voltage ratio of the circuit of Fig. 4 is given by

$$\frac{E_2}{E_1} = -\left[\frac{y_{12}^{(a)} - y_{12}^{(b)}}{y_{22}^{(a)} + y_{22}^{(b)}}\right] \quad \ldots \quad (11)$$

Let this ratio be given also by Equ. (1) in which $D(p)$ has roots $-d_1$, $-d_2$, $-d_3 \ldots . -d_n$. Consider a polynomial $F(p)$, with positive real coefficients and negative real simple roots.

$$F(p) = c_r p^r + c_{r-1} p^{r-1} \ldots . + c_0 \quad \ldots \quad (12)$$

Let $r = n$ or $r = n - 1$, and let the roots of $F(p)$ occur for $p = -f_1, -f_2 \ldots . -f_r$. The polynomial must be such that its roots alternate with those of $D(p)$; i.e., $d_1 < f_1 < d_2 < f_2 \ldots . d_{n-1} < f_{n-1} < d_n < f_n$. Clearly, if $r = n - 1$, the final inequality does not apply. The ratio $D(p)/E(p)$ now has the properties of an RC driving-point admittance, and may therefore be identified with $(y_{22}^{(a)} + y_{22}^{(b)})$. The ratio $hN(p)/F(p)$. may be identified with $-(y_{12}^{(a)} - y_{12}^{(b)})$ and, in particular, it will always be possible to split the numerator function $N(p)$ so that $y_{12}^{(a)}$ and $y_{12}^{(b)}$ have negative real roots, and may thus be associated with ladder networks. That is

$$\frac{h\ N(p)}{F(p)} = h\cdot\frac{(N_1(p) - N_2(p))}{F(p)} \quad \ldots \quad (13)$$

$$-y_{12}^{(a)} = h\cdot\frac{N_1(p)}{F(p)} \quad \ldots \quad (14a)$$

$$-y_{12}^{(b)} = h\cdot\frac{N_2(p)}{F(p)} \quad \ldots \quad (14b)$$

To show that such a division of $N(p)$ is possible, it is only necessary to consider the partial fraction expansion of $N(p)/F(p)$.

$$\text{Let } \frac{N(p)}{F(p)} = s_\infty p + s_0 + p\sum^{r} \frac{s_k}{p+f_k} \quad \cdots \quad \cdots \quad (15)$$

In this expansion, some of the coefficients s, will be positive and some negative. Collect together those of like sign, so that

$$\frac{N(p)}{F(p)} = \frac{N_1(p)}{F(p)} - \frac{N_2(p)}{F(p)} \quad \cdots \quad \cdots \quad \cdots \quad (16)$$

in which the coefficients of the partial fraction expansions of $N_1(p)/F(p)$ and $N_2(p)/F(p)$ are now all positive. Each of these ratios will have the form of a RC driving-point admittance. Therefore, the functions $N_1(p)$ and $N_2(p)$ have negative real roots.

It is not suggested that this particular method of obtaining $N_1(p)$ and $N_2(p)$ is necessarily the most useful, although, as will be argued later, it is generally the method yielding a network containing the minimum number of components. The method is clearly unduly restrictive since, in this case, not only do $N_1(p)$ and $N_2(p)$ have negative real roots, but they have a special relationship to the roots of $D(p)$.

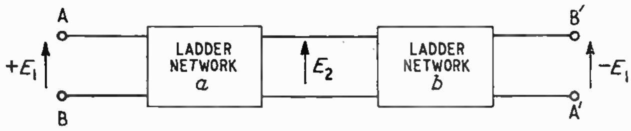

Fig. 5. *Balanced ladder networks (a) and (b) in series*

It will be possible now to share $D(p)/F(p)$ between $y_{22}{}^{(a)}$ and $y_{22}{}^{(b)}$, and proceed to the synthesis of two ladder networks by any of the established methods[8,9,10]. The two ladder networks are made balanced, as indicated by Fig. 5 and, by virtue of this balancing, there is no potential difference between terminals A and A′, or between terminals B and B′. These pairs of terminals may therefore be joined, to yield the network of Fig. 6, consisting of two balanced ladder networks connected in parallel at one end, and crossed at the other.

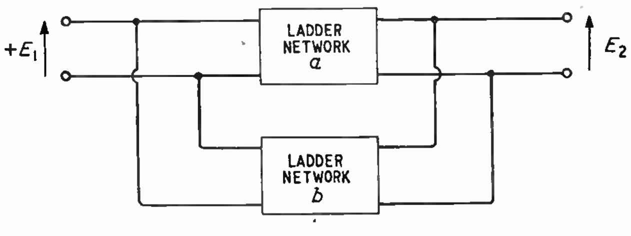

Fig. 6. *Modification of Fig. 5 by strapping terminals AA′ and BB′*

The procedure outlined may be extended to yield a network which will give the required transfer function when working into any RC impedance. In such a case Equ. (11) is replaced by

$$\frac{E_2}{E_1} = \frac{-[y_{12}{}^{(a)} - y_{12}{}^{(b)}]}{y_{22}{}^{(a)} + y_{22}{}^{(b)} + Y_L} \quad \cdots \quad \cdots \quad \cdots \quad (17)$$

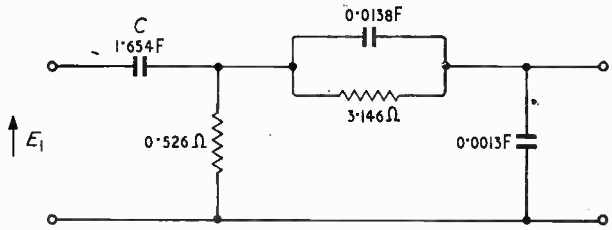

Fig. 7. *First stage in synthesis of Fig.* 10

where Y_L is the admittance of the required load. The procedure is similar to that developed by Bower and Ordung for the loaded lattice network.

An Illustrative Example

$$\text{Let } \frac{E_2}{E_1} = h \cdot \frac{(p-1)\,(p-2)\,(p-3)}{(p+1)\,(p+2)\,(p+3)} \quad \cdots \quad \cdots \quad (18)$$

The polynomial $F(p)$ will be chosen to be a quadratic, the roots of which lie between -1 and -2, and between -2 and -3. In order to minimize the ratio of the comparable coefficients in the expansions of $N(p)/F(p)$ and of $D(p)/F(p)$, the polynomial is chosen to be $(3p+4)\,(2p+5)$. Then

$$-(y_{12}{}^{(a)} - y_{12}{}^{(b)}) = h \cdot \left(\frac{p}{6} - \frac{3}{10} - \frac{99p}{10(2p+5)} + \frac{65p}{6(3p+4)}\right) \quad \cdots \quad (19)$$

In order to make the final network as simple as possible, the two transfer admittances are identified as follows:

$$-y_{12}{}^{(a)} = h \cdot \left(\frac{p}{6} + \frac{65p}{6(3p+4)}\right) \quad \cdots \quad \cdots \quad (20a)$$

$$-y_{12}{}^{(b)} = h \cdot \left(\frac{3}{10} + \frac{99p}{10(2p+5)}\right) \quad \cdots \quad (20b)$$

It is assumed that the output voltage is to be developed across a finite load resistance, but Y_L is assumed to be

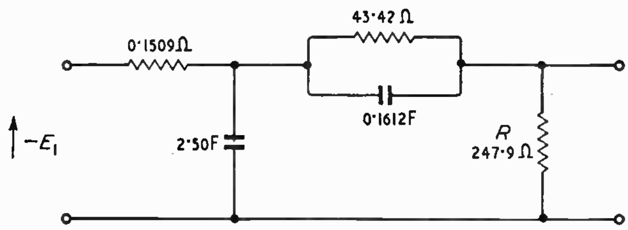

Fig. 8. *Second stage in synthesis of Fig.* 10

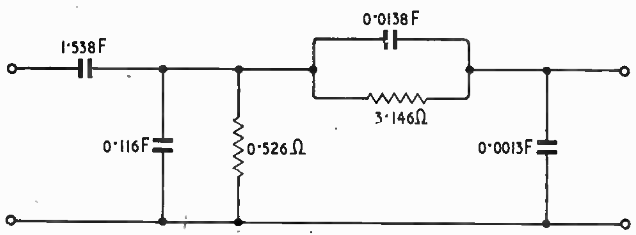

Fig. 9. *Third stage in synthesis of Fig.* 10

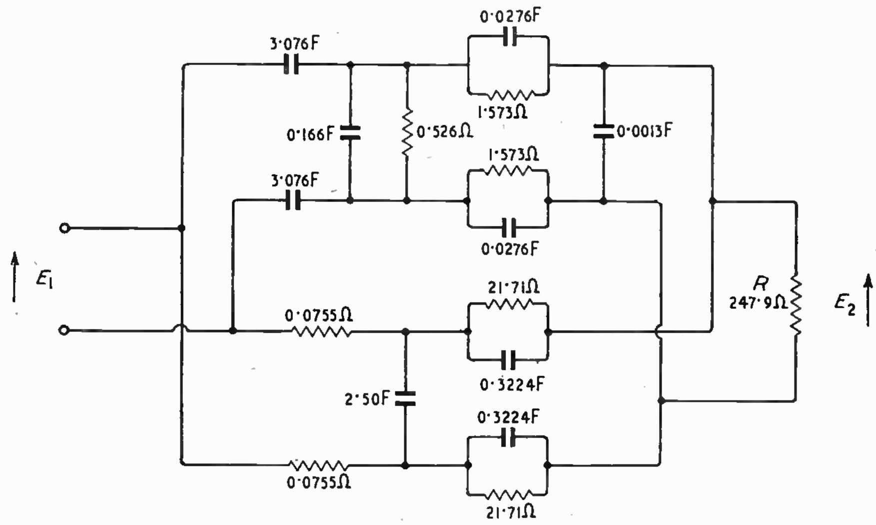

Fig. 10. *Final balanced RC network with resistance R of Fig. 8 removed to act as the load resistance*

zero since it will be possible to extract a conductance, from one of the ladder networks, to act as a load impedance.

The partial fraction expansion of $D(p)/F(p)$ now gives

$$y_{22}{}^{(a)} + y_{22}{}^{(b)} = \frac{p}{6} + \frac{3}{10} + \frac{5p}{42(3p+4)} + \frac{3p}{70(2p+5)} \quad (21)$$

Since the roots of $F(p)$ have been located in the positions to minimize the ratios of comparable coefficients in Equs (19), (20a) and (20b), it follows that the Bower and Ordung method would yield a maximum value of h of $3/70 \times 10/99 = 1/231$. That is, the pad loss would be 47·3 dB. In order to obtain a lower pad loss, although at the cost of some increase in the number of elements in the network, $y_{22}{}^{(a)}$ and $y_{22}{}^{(b)}$ are defined by

$$y_{22}{}^{(a)} = \frac{5p}{42(3p+4)} + \frac{p}{60} + \frac{27}{100} \quad \text{.. .. (22a)}$$

$$y_{22}{}^{(b)} = \frac{3p}{70(2p+5)} + \frac{9p}{60} + \frac{3}{100} \quad \text{.. .. (22b)}$$

This is clearly not the simplest way of dividing the right-hand side of Equ. (21) but, if the driving-point admittances were made to contain only terms corresponding to the allocation in Equ. (20), the pad loss would be 47·3 dB, as in the other method mentioned. The maximum value of h that could be obtained for a pair of admittances, say $y_{22}{}^{(a)}$ and $y_{12}{}^{(a)}$, may be evaluated by the techniques discussed by Fialkow and Gerst[4]. To do this in a general case may be tedious and lengthy, however, and the division suggested is a compromise between the maximization of h, the minimization of the complexity of the final network, and the curtailing of the synthesis process to a reasonable length.

A circuit to realize $y_{12}{}^{(a)}$ and $y_{22}{}^{(a)}$ is shown in Fig. 7, and the circuit of Fig. 8 will realize $y_{12}{}^{(b)}$ and $y_{22}{}^{(b)}$. The values of h obtained in the two cases are 0·0822 and 0·0765, respectively. In order to reduce h for network (a), the capacitance C is divided to give the network of Fig. 9, for which h is also 0·0765. The final balanced network is shown in Fig. 10, with a resistance removed to act as the load impedance. The network itself contains 16 elements, and has a pad loss of 22·3 dB. Orchard[11] has synthesized a symmetrical lattice to give the transfer function of Equ. (18) by a method which is different from that of Bower and Ordung. The lattice in this case contains 12 elements, and the pad loss is 36·1 dB.

It will be seen, therefore, that the proposed method compares favourably with the methods for the synthesis of a symmetrical lattice. It is more general than the method of Bower and Ordung, which is included as a special case, since the lattice may be regarded as two balanced ladder structures connected as shown in Fig. 6, the unbalanced ladder networks containing only one series impedance each.

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NEW YEAR HONOURS

Col. D. McMillan, O.B.E. (Director, External Communications, G.P.O.) becomes a C.B.

J. A. Ratcliffe, F.R.S. (Chairman, Radar and Signals Advisory Board, Ministry of Supply Scientific Advisory Council) and W. A. S. Butement (Chief Scientist, Department of Supply, Australia) are appointed C.B.Es.

R. F. Ballard (General Manager, Kolster-Brandes), Dr. L. Essen (N.P.L.) and E. F. Wheeler (Superintendent Engineer, Transmitters, B.B.C.) are appointed O.B.Es, while R. D. Petrie (B.B.C. Designs Department), P. C. Ruggles (Senior Engineer, English Electric) and D. C. Walker (Senior Executive Officer, Post Office Research Station) receive M.B.Es.

SYMPOSIUM ON HIGH VACUA

The Institute of Physics has arranged a one-day symposium on "Current Developments in the Production of High Vacua" to be held in London on 17th April 1959.

Further information may be obtained from the Secretary, The Institute of Physics, 47 Belgrave Square, London, S.W.1.

Sampling of Signals Without D.C. Components

By A. R. Billings, B.Sc., Ph.D., A.M.I.E.E.*

SUMMARY. *It is shown that a signal contained in a finite frequency band can only be sampled unambiguously at sampling frequencies lying within certain permitted bands, where the last of these bands extends to infinity. This leads to a more general statement of the Shannon-Hartley law restricting the permissible sampling frequencies, which is then applicable to all band-limited signals.*

It is well known[1,2,3] that any signal contained in a frequency band 0–Δf c/s can be sampled unambiguously at any frequency equal to or greater than $2\Delta f$. However, if the signal is contained in a frequency band of Δf which does not include zero, the minimum and permitted sampling frequencies are not given by this simple law. It is possible, by the use of modulators, to produce a modified signal which does occupy the band 0–Δf c/s and, in this case, the Shannon-Hartley sampling law is applicable to the modified signal. However, it is not always desirable to employ this frequency shifting, and this article is concerned with the calculation of permitted sampling frequencies for the unshifted signal. It will be shown that if the lowest signal frequency is f_0, and

$$f_0 = (N + K)\Delta f \quad \ldots \quad (1)$$

where N is an integer and K is less than unity, then there are $(N + 1)$ bands of permissible sampling frequencies.

Conditions for Non-Ambiguous Sampling

When a continuous time function or signal is sampled, the samples themselves constitute a new time function; and when sampling occurs regularly at a frequency f_s, this new time function is the product of the original signal and a train of narrow pulses spaced $1/f_s$ seconds apart. In consequence, the spectrum of the samples is the convolution of the spectrum of the pulses with the spectrum of the original signal. Examples of such spectra, when the original signal is contained in a band f_0 to $f_0 + \Delta f$, are shown in Figs. 1 and 2. These diagrams are intended only to show frequency occupation and not relative amplitudes. In fact, the amplitudes of the sample spectra are much less than those of the original signal spectra. It will be seen that, except for this reduction of amplitude, the spectrum of the samples contains that of the original signal, together with a set of sidebands produced by the sampling process. The reason for this can be simply stated. The pulse train has a spectrum consisting of lines spaced f_s apart starting at zero frequency and, therefore, any frequency f in the original signal appears as a set of frequencies $|f + nf_s|$ in the sample spectrum, where n takes all integral values from $+\infty$ to $-\infty$. Similarly, a band of frequencies f_0 to $(f_0 + \Delta f)$ in the original signal appears in the sample spectrum as a set of bands each extending from $|f_0 + nf_s|$ to $|f_0 + nf_s + \Delta f|$.

A signal is said to be sampled unambiguously if the original signal, or a reduced amplitude version of the original signal, can be extracted from the samples by simple filtering. In terms of spectra, this implies that it is necessary that the sidebands produced by the harmonics of the sampling frequency shall not overlap the frequency band containing the original signal. This condition is both necessary and sufficient, since the portion of the samples contained in the band f_0 to $(f_0 + \Delta f)$ is simply the product of the original signal and the zero frequency component of the sampling pulses.

Bands of Permitted Sampling Frequencies

The nth harmonic of the sampling frequency produces two sidebands, the first extending from $(f_0 + nf_s)$ to $(f_0 + nf_s + \Delta f)$ and the second extending from $|f_0 - nf_s|$ to $|f_0 - nf_s + \Delta f|$. If, as is necessary, f_s is greater than Δf, then the first of these sidebands lies above the signal band and does not overlap it. This means that attention can be confined to the second, or difference-frequency, sideband. Let n be chosen such that the difference-frequency sideband is the first band lying below the signal band and, in consequence, that the sideband produced by the $(n + 1)$th harmonic is the first band above the signal band. These correspond to bands D and E in Fig. 1(c) and (d). The highest frequency produced by the nth harmonic is

$$f = nf_s - f_0 \quad \ldots \quad (2)$$

and the lowest frequency produced by the $(n + 1)$th harmonic is

$$f = (n + 1)f_s - f_0 - \Delta f \quad \ldots \quad (3)$$

Thus, the conditions necessary if these frequencies are not to lie in the signal band, are given by

$$nf_s - f_0 \leqslant f_0 \leqslant (n + 1)f_s - f_0 - 2\Delta f \quad \ldots \quad (4)$$

To discover the permitted bands of sampling frequencies it is now necessary to determine the permissible values of n and to insert these in Equ. (4).

As a step towards this, consider the condition for no overlap between the sideband produced by the nth harmonic and the band immediately below it in

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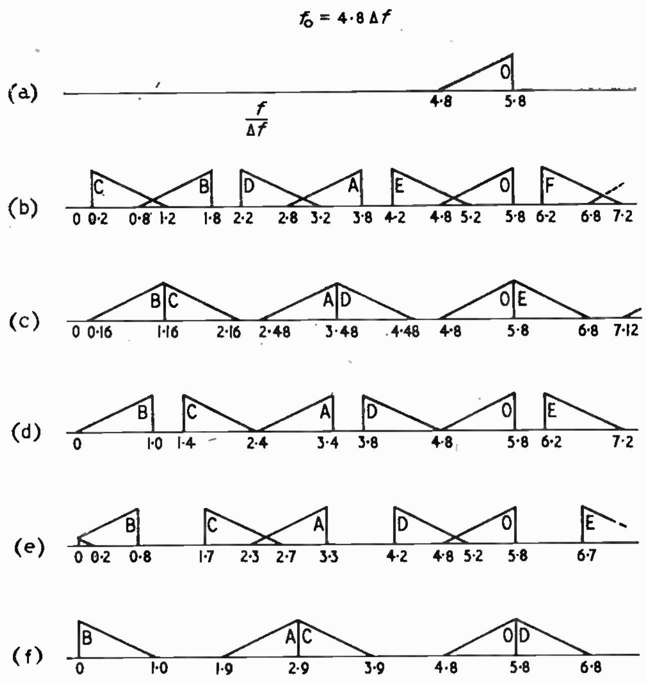


Fig. 1. *Spectra of samples when $f_0 = 4{\cdot}8\Delta f$ (N even). (a) Original signal spectrum. Sample spectra when (b) $f_s < f_{s1}$, (c) $f_s = f_{s1}$, (d) $f_s = f'_{s1}$ (e) $f_{s2} > f_s > f'_{s1}$, (f) $f_s = f_{s2}$.*

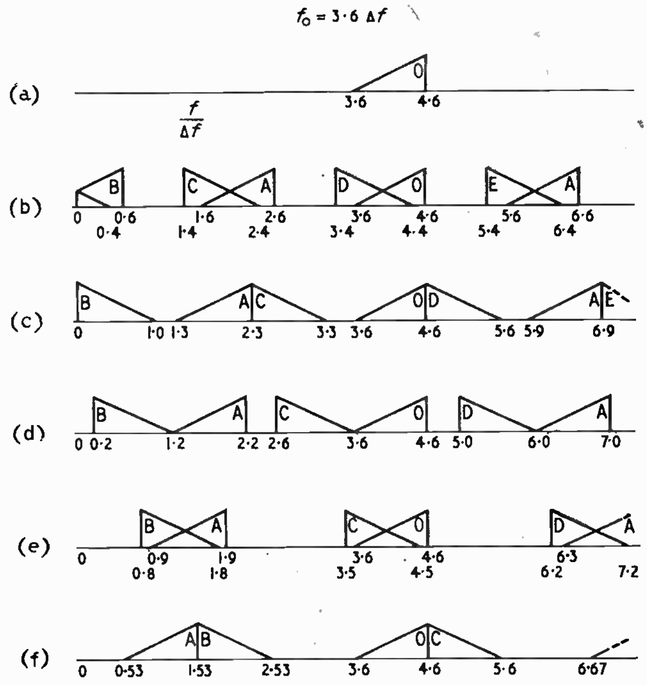


Fig. 2. *Spectra of samples when $f_0 = 3{\cdot}6f$ (N odd). (a) Original signal spectrum. Sample spectra when (b) $f_s < f_{s1}$, (c) $f_s = f_{s1}$, (d) $f_s = f'_{s1}$, (e) $f_{s2} > f_s > f'_{s1}$, (f) $f_s = f_{s2}$.*

frequency, which is produced by the fundamental component of the sampling impulses. These correspond to bands D and A in Fig. 1(c) and (d). The lowest frequency in the band produced by the nth harmonic is

$$f = nf_s - f_0 - \Delta f \quad .. \quad .. \quad .. \quad .. \quad (5)$$

and the highest frequency in the band produced by the fundamental is

$$f = f_0 + \Delta f - f_s \quad .. \quad .. \quad .. \quad .. \quad (6)$$

whence the condition for overlap between these bands is

$$f_0 \leqslant (n+1) f_s - f_0 - 2\Delta f \quad .. \quad .. \quad .. \quad (7)$$

Finally, consider the condition for no overlap between the sideband produced by the fundamental and the band immediately below it produced by the $(n - 1)$th harmonic; i.e., bands A and C in Fig. 1(c) and (d). The highest frequency in the band produced by the $(n - 1)$th harmonic is

$$f = (n-1) f_s - f_0 \quad .. \quad .. \quad .. \quad .. \quad (8)$$

and the lowest frequency in the band produced by the fundamental is

$$f = f_0 - f_s \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

from which the condition of no overlap is

$$f_0 \geqslant nf_s - f_0 \quad .. \quad .. \quad .. \quad .. \quad (10)$$

whence, the conditions necessary if the band produced by the fundamental is not to overlap any other band are

$$nf_s - f_0 \leqslant f_0 \leqslant (n+1) f_s - f_0 - 2\Delta f \quad .. \quad (11)$$

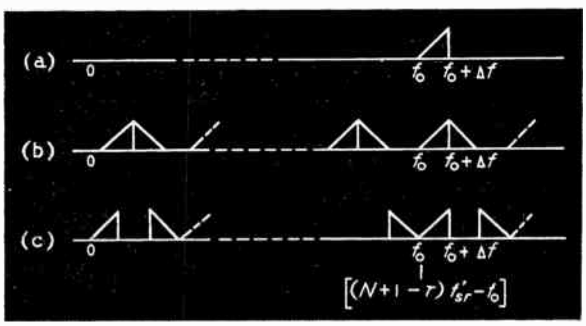
It will be seen that Equs (4) and (11) are identical. Also an extension of the above argument leads to the same conditions for non-overlap of any band with its neighbours, or with itself folded back about zero frequency.

It is now possible to discover the permitted bands of sampling frequencies. Since the conditions for non-overlap of one sideband with any other sideband are the same as for the non-overlap of any sideband with the signal band, it follows that the maximum value which n can assume is N, where N is the next integer below $f_0/\Delta f$. Thus, the first band of permitted sampling frequencies is defined by

$$Nf_s - f_0 \leqslant f_0 \leqslant (N+1) f_s - f_0 - 2\Delta f \quad .. \quad (12)$$

If the upper and lower limits of this band are f_{s1} and

Fig. 3. *Spectrum samples when sampling frequency is in rth sampling band. (N even). (a) Original signal spectrum. Sample spectra when (b) $f_s = f_{sr}$, r odd, (c) $f_s = f'_{sr}$, r odd.*



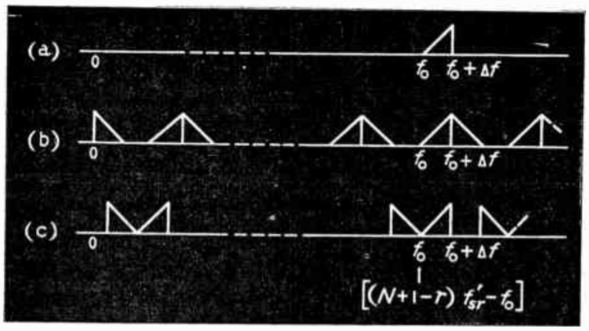


Fig. 4. *Spectrum of samples when sampling frequency is in rth sampling band.* (N *odd*). (*a*) *Original signal spectrum. Sample spectra when* (*b*) $f_s = f_{sr}$, r *odd*, (*c*) $f_s = f'_{sr}$, r *odd.*

f'_{s1} respectively, then

$$f_{s1} = \frac{2\Delta f\,(N+K+1)}{(N+1)} \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

and

$$f'_{s1} = \frac{2\Delta f\,(N+K)}{N} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

where K is as defined in Equ. (1).

If N is greater than zero, there are further bands of sampling frequencies above this. In particular, for the rth of these bands.

$$(N+r-1)f_s - f_0 \leqslant f_0 \leqslant (N+r)f_s - f_0 - 2\Delta f \quad (15)$$

whence

$$f'_{sr} = \frac{2\Delta f\,(N+K+1)}{(N+2-r)} \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

and

$$f'_{sr} = \frac{2\Delta f\,(N+K)}{(N+1-r)} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Examples of spectra obtained when sampling at frequencies f_{sr} and f'_{sr} are shown in Figs 3 and 4, and the permitted sampling frequencies for low values of N are shown graphically in Fig. 5.

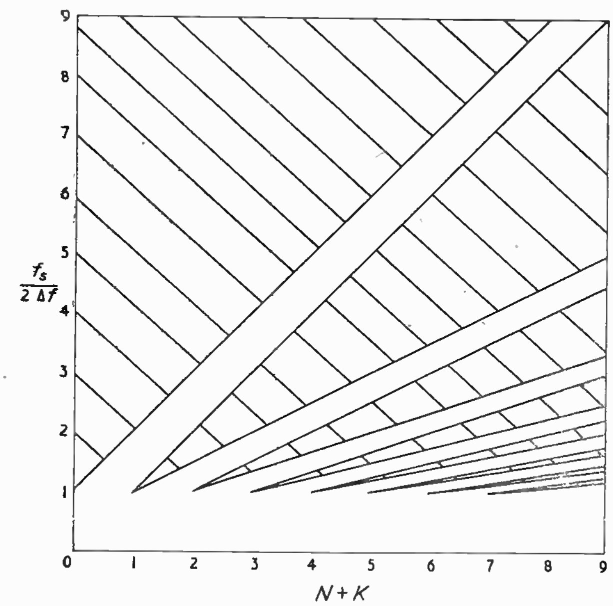


Fig. 5. *Permitted sampling frequencies for low N. (Bands of permitted sampling frequencies shown shaded).*

Interpretation of Results

From Equs (16) and (17) it can be seen that there are $(N+1)$ bands of permissible sampling frequencies, where the last band has an upper limit of infinity and a lower limit equal to twice the upper signal frequency. This lower limit in the $(N+1)$th band corresponds to the minimum sampling frequency given by the Shannon-Hartley law when the fact that there is no signal in the band 0–f_0 c/s is ignored.

When the ratio of the lowest signal frequency to the bandwidth is an integer; in other words, when $K = 0$, the first sampling band contracts to a single frequency $2\Delta f$, which is the same as the minimum sampling frequency given by the Shannon-Hartley law for a signal of the same bandwidth starting at zero frequency. When $K \neq 0$ the minimum sampling frequency is greater than $2\Delta f$. The worst case occurs when $N = 0$ and K tends to unity, when the minimum sampling frequency tends to $4\Delta f$.

From the results obtained above, it is possible to restate the sampling law as follows. When sampling a signal of bandwidth Δf having a lowest frequency f_0 which is not equal to zero, the minimum sampling frequency lies between $2\Delta f$ and $4\Delta f$ and permitted sampling frequencies lie in one or more bands, where the number of bands is the next integer above $(f_0/\Delta f)$.

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Lens Aerial Design

SIMPLE GEOMETRICAL METHOD

By G. P. Foldes* and L. Solymar†

SUMMARY. *A graphical method is described of designing a lens for an aerial that will realize a prescribed amplitude and phase distribution function in an aperture, when fed by a given primary source.*

It is often required to design a shaped beam aerial for a prescribed radiation pattern. If this radiation pattern is a slowly varying function, this problem can be solved on the basis of geometrical optics, where it is supposed to be valid in the whole space filled by radiation.[1] However, if there are sudden changes in the radiation pattern (for example the sharp cut-off of a cosecant aerial near the ground surface) the geometrical optics cannot be used in the far field region. In this case the aperture distribution function must be computed. It may be obtained from the prescribed radiation pattern by one of the approximation methods[2,3]. Hence the problem is equivalent to the realization of a prescribed amplitude and phase distribution function in a given aperture.

Let us suppose that the radiation pattern of the primary source is given. Then the problem to be solved is how to design a suitable device which will transform the radiation pattern of the primary source into the desired amplitude and phase distribution function in the aperture. The treatment will be restricted to two-dimensional cases.

Outline of the Problem

For the realization of the complex distribution function two independent surfaces (two independent mirrors or the two refracting surfaces of a lens) are necessary. The lens promises an easier solution and even for constant refractive index, there are two degrees of freedom.

The solution of the problem generally cannot be obtained in closed mathematical form. The approximate numerical calculations are very lengthy, therefore a relatively simple graphical method will be applied.

For simplicity, the solution will be obtained only for a symmetrical arrangement, but any asymmetrical distribution function may be realised on the same principle.

Within the aerial region geometrical optics is valid, if the prescribed amplitude and phase distribution functions do not vary too quickly.‡

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* *Now with 'RCA Montreal', Canada.*

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‡ This restriction does not imply that the radiation pattern must also be varying slowly.

The function of the lens is to direct the energy between any two rays of the primary source to the appropriate part of the aperture and ensuring, at the same time, that the emergent rays should be perpendicular to the prescribed phase front.

Realization of the Amplitude Distribution Function

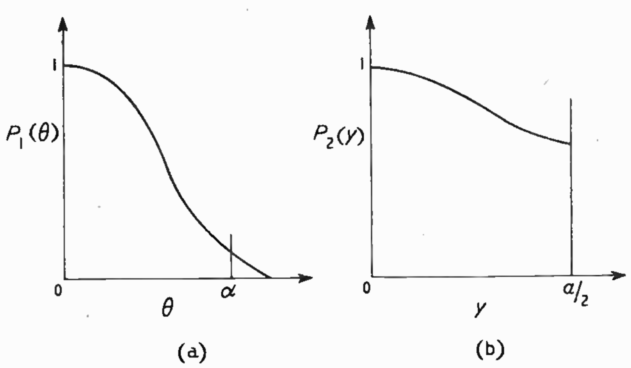
Let us plot in a rectangular co-ordinate system $P_1(\theta)$ the power pattern of the primary source, and $P_2(y)$ the power distribution function (Fig. 1), where θ is the angle between a ray of the primary source and the axis of symmetry, and y is the co-ordinate along the aperture.

The power radiated by the primary source will be utilized only up to an angle α, where 2α is the angle which the lens subtends at the centre of the primary source. The power radiated beyond this angle will be lost, therefore the useful power is given by

$$P_u = \int_0^\alpha P_1(\theta)\, d\theta \quad \ldots \quad (1)$$

Depending on the required accuracy let us divide the region between 0 and α into N not necessarily equal intervals. Let $\theta_1, \theta_2, \ldots \theta_k, \ldots$ be the dividing points in the primary radiation pattern; and let y_1,

Fig. 1. (*a*) *The area contained within the co-ordinates* (0, 1) *and* (0, α) *represents the useful power radiated by the primary source;* (*b*) *The area contained under this curve and the y axis represents the power directed into an aperture of size a/2*



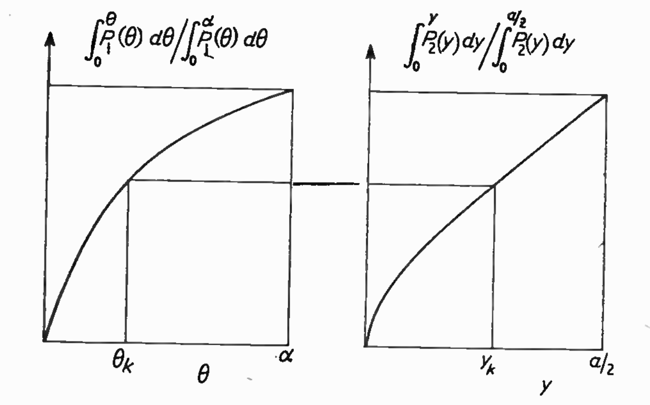


Fig. 2. *Illustrating how the interval y_k (corresponding to a particular value of θ_k) can be obtained by plotting the integrals against θ and y respectively*

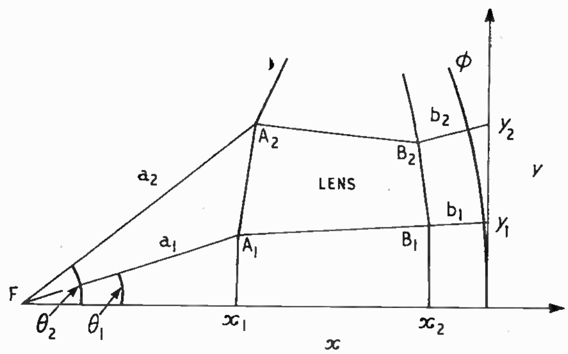


Fig. 4. *Construction of the contours of the lens*

$y_2, \ldots y_k, \ldots$ the corresponding co-ordinates in the aperture.

The power radiated between the angles 0 and θ_k is

$$\int_0^{\theta_k} P_1(\theta)\, d\theta \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (2)$$

Let us direct this power into the interval $(0, y_k)$ in the aperture, where y_k may be determined from the mathematical condition

$$\frac{\int_0^{\theta_k} P_1(\theta)\, d\theta}{\int_0^{\alpha} P_1(\theta)\, d\theta} = \frac{\int_0^{y_k} P_2(y)\, dy}{\int_0^{a/2} P_2(y)\, dy} \qquad \ldots \quad \ldots \qquad (3)$$

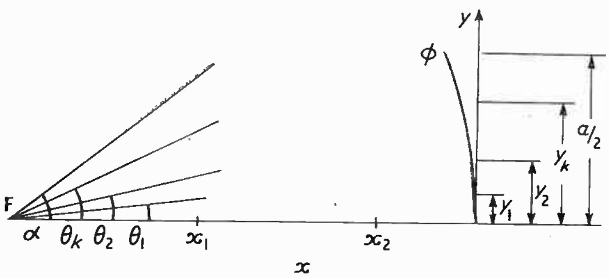
where a is the size of the aperture.

The value of y_k corresponding to a particular value of θ_k can be obtained by plotting both sides of Equ. (3) as indicated in Fig. 2.

Realization of the Phase Distribution Function

Thus, with the aid of Fig. 2 we have for an arbitrarily chosen set of interval angles θ_k, a set of intercepts y_k. These are shown in Fig. 3. Let x_1 and x_2 represent the points at which the lens cuts the symmetry axis x; and let ϕ represent the prescribed constant phase-front. Clearly, a ray in the direction θ_k (after being refracted twice by the lens) should reach y_k in a direction normal to the curve ϕ. This can be utilized to determine the approximate contours of the required lens in a series of steps.

Fig. 3. *The set of interval angles θ_k and intercepts y_k*



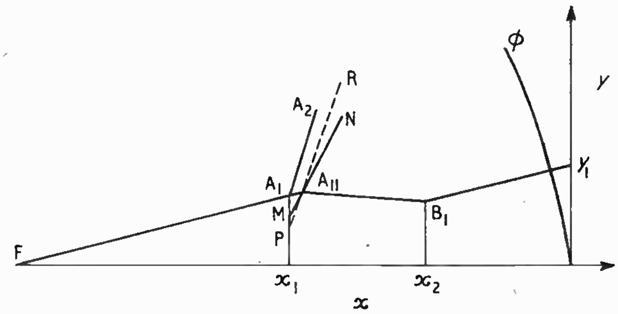


Fig. 5. *Modified method for constructing the contours of the lens*

Construction of the Contours of the Lens

First we construct (as shown in Fig. 4) the ray 'a₁' making an angle θ_1 with the x axis, and the ray 'b₁' from y_1 perpendicular to the curve ϕ. These are the incident and emergent rays. The perpendiculars to the axis x at x_1 and x_2 meet these rays at A_1 and B_1 respectively. A_1B_1 is the direction of the ray within the lens. Using a construction described in the appendix we can find the directions of the interfaces at A_1 and B_1 which will satisfy Snell's laws of refraction. Let these be A_1A_2 and B_1B_2. The second incident and emergent rays 'a_2' and 'b_2' cut these at A_2 and B_2 respectively. The construction is repeated and a series of points $A_1, A_2, \ldots A_N$ and $B_1, B_2, \ldots B_N$ serve as approximate surfaces of the lens. The surfaces so obtained will depend on the original choice of x_1, x_2 the thickness and position of the lens.

If the thickness is chosen too small, the lens may terminate before it intercepts all the rays up to α. If it is chosen too large, excess losses may occur. In the case of a lens of high dielectric constant, x_1 and x_2 should be chosen in such a way that multi-reflections within the lens would be kept at a low level.

The lens must be kept well within the Fraunhofer region of the primary source. Consideration of bulk

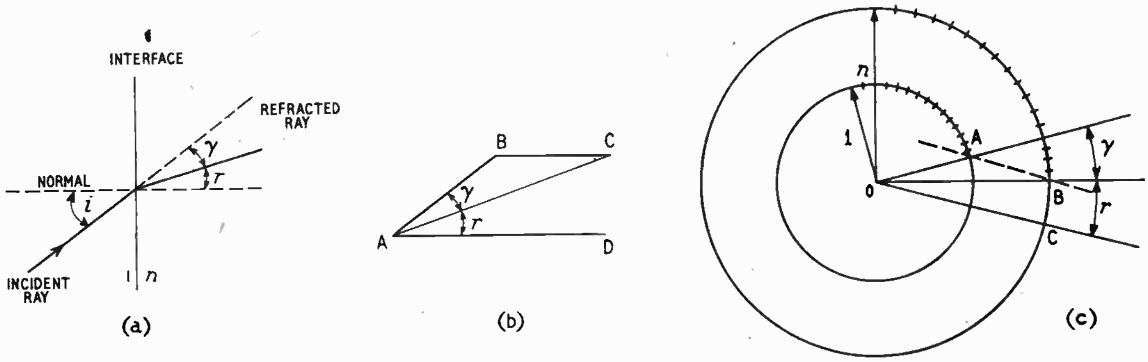


Fig. 6. *Construction of an interface for given incident and refracted rays; (a) showing the deviation of a single ray through a medium of refractive index n; (b) illustrating basis of solution to satisfy Snell's law, and (c) practical construction to find the angle of refraction r*

and mechanical stability will limit the maximum distance of the lens from the source.

With some experience, an appropriate choice of position and thickness of the lens can be obtained in most cases. In the cases where this is not so, the nature of the failure would indicate the necessary alterations.

Improved Method

The method of construction described yields a polygon which is an approximation to the contour of the lens. It is possible to obtain a somewhat better approximation with a modified method of construction.

As before, the rays 'a_1' and 'b_1' meet the perpendiculars from x_1 and x_2 at A_1 and B_1 respectively. Through P (Fig. 5), halfway along A_1x_1 draw PR parallel to A_1A_2. Produce 'a_1' to meet PR at A_{11}, join $A_{11}B_1$. At A_{11} construct MN as described in the Appendix. Produce 'a_2' to meet MN and repeat the procedure. The polygon so obtained is a better approximation, because the point A_{11} is a point of the continuous contour and MN is its tangent at A_{11}.

APPENDIX

Construction of an interface when the incident and refracted rays are given.

The problem is illustrated in Fig. 6(a) where the incident ray is shown on the left, and the refracted ray (making an angle γ with the incident ray) is on the right. It is required to find an interface which will satisfy Snell's law

$$n \sin \gamma = \sin (\gamma + r)$$

Fig. 6(b) illustrates the basis of the solution found. In the triangle ABC, AB = 1, AC = n and $\angle$ BAC = γ. AD is drawn parallel to BC. The angle CAD = r.

For quick practical construction one may employ the following method. Construct two circles of radius 1 and n respectively and mount on a common centre. Round the periphery mark out degrees on each circle. Rotate one circle relative to the other until the zero lines make an angle γ between them, as shown in Fig. 6(c). Mark OC parallel to AB. The required angle r is given by BOC.

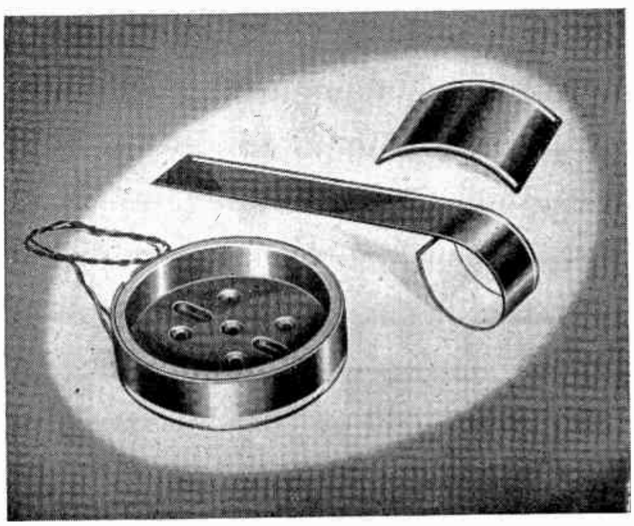
REFERENCES

[1] S. Silver, "Microwave Antenna Theory and Design". McGraw-Hill, New York, 1949.
[2] P. M. Woodward, "A Method of Calculating the Field over a Plane Aperture required to produce a given Polar Diagram", *J. Instn elect. Engrs*, Part IIIA, 1946, Vol. 93, p. 1554.
[3] R. Kovacs and L. Solymar, "Theory of Aperture Aerials Based on the Properties of the Entire Functions of the Exponential Type", *Acta phys. hungarica*, 1956, p. 161.

CURVED PHOTOELECTRIC CELLS

'Contour photocells' are specially-processed pliable selenium photocells developed by the International Rectifier Corporation, California, U.S.A. According to the manufacturers, these cells (which consist of a metal base on which layers of selenium, cadmium and gold are deposited in turn) may be used in a wide variety of electronic photoelectric devices for control applications, production flow processes, automatic inspection and sorting, and similar functions. For instance, they can be mounted on a rotating shaft in a position control servo-mechanism, or when formed into a photo-sensitive cam they may be used as a type of non-linear function generator.

The cells can be produced to any requirement in three-dimensional shapes with as little as one-inch radius of curvature and are available in sizes ranging from a minimum of ¼ in. × ¼ in. to a maximum of 10 in. × 10 in.



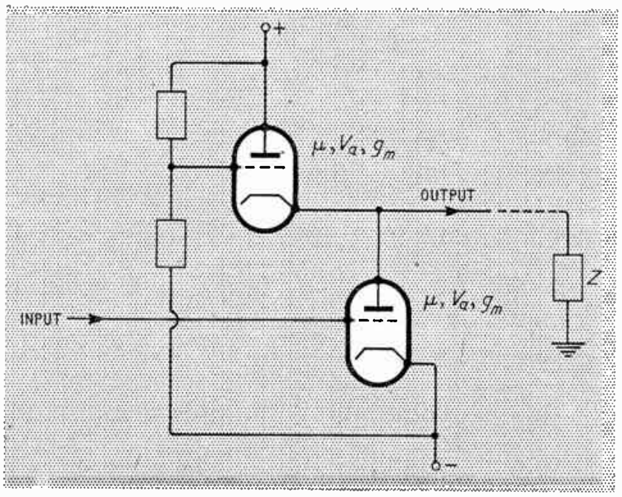
Correspondence

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.

New Anode Follower

SIR,—In his article in your September issue, Mr. Keen gives an excellent account of the bootstrap circuit technique.

A circuit, which has the advantages of the anode follower, without its disadvantages, is derived from the cascode amplifier which has as output terminal not the anode, but the common point of both valves.



As shown in the figure, the first valve has an earthed cathode and the second has earthed grid and anode.

It can be shown, that the output impedance is

$$Z_{out} \approx 1/g_m$$

and the amplification

$$\frac{V_{output}}{V_{input}} = \frac{-\mu}{\mu + 2 + r_a/Z}$$

The advantage of the circuit is the high input impedance. There is no parallel voltage feedback to the input terminals to decrease the input impedance, as in the usual anode follower.

J. HAJEK.

Laboratory of Industrial Electronics of CSAV,
Brno, Czechoslovakia.
23rd November 1958.

Step Detection

SIR,—Step detectors, or pulse-lengtheners as they are sometimes called, have been used in pulse-amplitude modulated time-division multiplex systems for several years and it is well known that they introduce attenuation distortion. This result is implicit in the classic paper by Moss[1] and an explicit expression for the distortion has been derived by Kleene[2].

In a paper in the December issue of *Electronic and Radio Engineer*, Dr. Billings considers both attenuation distortion and inter-channel crosstalk. If the transmission path has a step response which has an overshoot, it is possible to obtain zero crosstalk from the preceding channel even when a conventional low-pass filter is used for demodulation[3]. The pulse of the receiving gate can be so timed with respect to the oscillatory response that the mean received crosstalk voltage is zero. When a pulse-lengthener is used, its effect on crosstalk depends on the type of circuit. In one type, the lengthener capacitor can only be charged by the received pulse and is discharged through a resistor or by another pulse. Alternatively, the received pulse can be applied to the capacitor by a gating circuit that enables it to charge or discharge according to the amplitude of the received pulse. It appears from his Fig. 2(h) that the author is considering the latter case and that the resistance of the gate is assumed to be negligible so that the p.d. across the capacitor can equal the instantaneous voltage of the pulse. If the gate closes at the instant when the received pulse from the preceding channel is zero, the adjacent-channel crosstalk will be finite, as stated by Dr. Billings. However, if the gate is closed slightly later, the crosstalk voltage left across the capacitance can be small and of opposite sign to that existing during the previous part of the received pulse so that the mean voltage during the cycle is zero. In practice, the situation will be complicated by a finite charging resistance causing the p.d. across the capacitor to lag behind the voltage of the received pulse.

Most methods of cancelling crosstalk in narrow-band t.d.m. systems, whether they depend on an oscillatory response (with or without step detection) or on a monotonic response[4,5], require extreme accuracy to be maintained in the timing waveforms used. In order to ensure that crosstalk is inaudible, an attenuation of at least 70 dB is needed and this requires a timing tolerance of the order of $\pm$ 0·01 per cent of the pulse length. A stability of this order on the delay of the transmission path and the synchronizing and pulse-generating equipment is usually impracticable. If adequate crosstalk attenuation is required, there must be a considerable increase in the bandwidth of the transmission path to ensure that small changes in timing do not cause large increases in crosstalk. The crosstalk attenuation will then be adequate whether the pulse response of the transmission path is oscillatory or monotonic. Incidentally, a pulse-lengthener can provide some improvement in crosstalk when the pulse response is monotonic[6]. The improvement is greater when the charging time-constant of the pulse-lengthener is finite than when it is zero.

Siemens Edison Swan Ltd.,
Blackheath, London, S.E.3.
22nd December 1958.

J. E. FLOOD.

REFERENCES

[1] S. H. Moss, "Frequency Analysis of Modulated Pulses," *Phil Mag.*, September 1948, Vol. 39, p. 663.
[2] S. C. Kleene, "Analysis of Lengthening of Modulated Repetitive Pulses," *Proc. Inst. Radio Engrs*, October 1947, Vol. 35, p. 1049.
[3] J. E. Flood and J. R. Tillman, "Crosstalk in Time-Division Multiplex Systems," *Proc. Instn elect. Engrs*, July 1951, Vol. 98, Part III, p. 279.
[4] W. P. Boothroyd and E. M. Creamer, "A Time-Division Multiplexing System," *Elect. Engng, N.Y.*, July 1949, Vol. 68, p. 583.
[5] Thomas Roddam, "Narrow-Band Pulse Communication," *Wireless World*, June 1950, Vol. 56, p. 202.
[6] J. E. Flood, Ph.D. Thesis, University of London, 1951.

Relative Speeds of Telegraphic Codes

SIR,—As stated by Dr. Bell and Mr. Duggan (*Electronic & Radio Engineer*, December 1958, p. 476), a comparison of the speeds of telegraphic codes requires a careful statement of the basis of comparison. Even the term "English language" requires some qualification since, if a clear rendering of the text is to be given, the word-space and possibly some punctuation symbols must be transmitted in addition to the 26 letters of the alphabet. In the 5-unit code, the word-space is, of course, transmitted as a distinct character and the average length of a character remains at 5 units for synchronous working. For unequal-length codes, the inclusion of the word-space has a considerable effect on the average weighted length of character since it occurs more frequently than any other character. Punctuation is not so important since the frequencies are much lower. For the Morse code, the effect of including the word-space may be estimated as follows:

If the average number of letters in a word is taken to be four, the frequency of the word-space is 0·2. With character lengths reckoned as in the article, the word-space has a length of one digit, so that with its associated inter-character space and the inter-character space for the preceding character, there is a 7-digit space between the last 'marking' digit of one word and the first 'marking' digit of the next. The average weighted length of character for an English language text is then shown in Table 1 to be 8·01 digits (including

Table I

Morse letter	Length (digits)	Letter frequency	Weighted length
word space	1	0·200	0·200
E	1	0·105	0·105
T	3	0·084	0·252
I	3	0·050	0·150
A	5	0·066	0·330
N	5	0·057	0·285
S	5	0·049	0·245
R	7	0·054	0·378
M	7	0·020	0·140
H	7	0·042	0·294
D	7	0·030	0·210
U	7	0·020	0·140
F	9	0·023	0·207
B	9	0·011	0·099
G	9	0·016	0·144
K	9	0·003	0·027
L	9	0·027	0·243
V	9	0·007	0·063
W	9	0·012	0·108
C	11	0·022	0·242
O	11	0·064	0·704
P	11	0·016	0·176
X	11	0·002	0·022
Z	11	0·001	0·011
J	13	0·001	0·013
Q	13	0·001	0·013
Y	13	0·016	0·208
Total		0·999	5·009

the inter-character space) instead of 9·02 digits if the word-space is ignored.

Measurements made from sample batches of messages transmitted on public telegraph systems [1,2] show that the mean length of character is about 9·5 digits for the Morse code. The difference between this figure and the result obtained with an English language text may be explained partly by the omission of the more commonly used words from telegraphic texts and partly by the fact that a telegraphic alphabet of about 45 characters is normally used. Under these conditions, the 5-unit code has to make use of 'shift' characters, so that the average length of an effective character is about 5·3 digits. A Huffman code for such a message source has been shown to require an average of 4·6 digits per character.[3]

Post Office Research Station,
Dollis Hill, London, N.W.2.
C. J. Hughes.
31st December 1958.

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[1] I. S. Coggeshall, "Submarine Telegraphy in the Post-War Decade", *Trans. Amer. Inst. elect. Engrs*, 1930, Vol. 49, p. 476.
[2] T. Hayton, C. J. Hughes and R. L. Saunders, "Telegraph Codes and Code Convertors", *Proc. Instn elect. Engrs*, 1954, Vol. 101, Part III, p. 137.
[3] Reply to the discussion on the paper given in Reference 2.

New Books

Magnetic Recording Techniques

By W. Earl Stewart. Pp. 272 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 66s.

During the last ten years considerable advances have been made in the sphere of magnetic recording and these advances have been described in a correspondingly large number of published papers. From 1948, however, when S. J. Begun's book appeared, up to a year ago, there was a dearth of new books to integrate the new ideas and techniques. The work under review is one of a number which have appeared during the last twelve months and makes a welcome contribution towards filling the gap.

The book consists of six chapters and a series of nine appendixes. The first chapter deals with the recording process and discusses in some detail not only standard but also non-standard techniques as, for example, boundary-displacement recording. Chapter 2, entitled Magnetic Recording Media, refers briefly to these materials and provides a table of relevant data on the magnetic constituents of tape, wire and discs. But it also includes special formulae for dispersions for coating on cine film and discusses at length transfer, both unwanted and deliberate, and erasure. The third chapter describes the reproducing process and this, like Chapter 1, deals with the unconventional as well as the conventional. Chapter 4, Magnetic Recording Mechanisms, is probably the most valuable of all as it deals with the mechanical side of the recorder with a thoroughness all too rare in works on communications equipment. Chapter 5 is entitled Ferromagnetism and provides a conventional review of the subject with useful tables covering the properties of hard and soft magnetic materials. Chapter 6 reproduces American standards for tape and magnetically-coated cine film.

The nine appendixes, which occupy a total of 66 pages out of 272, are mainly verbatim extracts from published analytical papers on such subjects as self-demagnetism, accidental printing, reproduction of recorded signals and spacing and thickness loss formulae. This gives the book a slightly unbalanced appearance and it is felt that the work would have been more presentable if some at least of this material had been condensed and introduced into the body of the text.

Each chapter is supported by an extensive bibliography, but it is regrettable that, out of the 130 listed, there are only two Dutch, one German and two English. The remaining 125 are, of course, American! However, the non-Americans come into their own in the appendixes.

H.G.M.S.

Liquid Scintillation Counting

Edited by Carlos G. Bell, Jr. and F. Newton Hayes. Pp. 292. Pergamon Press Ltd., 4 & 5 Fitzroy Square, London, W.1. Price 70s.

The thirty-five articles in this report of a conference held at North-Western University in 1957 survey the theory, techniques, and applications of this relatively new method of radiation measurement. Versatile in its uses, which range from the detection of low-energy beta-emission from sources suspended within the liquid, to the counting of fast neutrons and the revelation of neutrinos, the method itself poses a wide range of physical and chemical problems. The instrumentation side is a challenge to the electronic engineer, and this aspect of the subject is discussed fully in a number of the contributions.

Switching Circuits and Logical Design

By Samuel H. Caldwell. Pp. 686 + xvii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2, for John Wiley & Sons Inc., New York. Price 112s.

The logical design of switching circuits seems to have received little attention until comparatively recently, and the subject is still in its infancy. Symbols and notation are in need of standardization and it is freely admitted that in the present state of the art it is not possible to express in mathematical terms *all* the conditions which must be satisfied to obtain the best solution to a switching problem.

In this book however, Dr. Caldwell has succeeded in explaining the principles of logical switching design in a manner easily understood by a newcomer to the subject. By initially treating Boolean Algebra independently of its application to switching circuits, the author is able to demonstrate its use in the complementary functions which he terms 'hindrance' and 'transmission' without confusing the reader. This introduction to switching algebra is followed by very lucid chapters on series-parallel networks, minimization methods, multiterminal networks, symmetrical networks, non-series-parallel networks and iterative networks. The chapter on minimization methods introduces the well-known methods, and also includes ones which have not previously been readily available. Chapters are also devoted to the problems peculiar to the use of electronic and solid-state elements and to the switching aspects of codes.

The last four chapters are devoted to the various aspects of sequential switching operations, two being in general terms and the others covering electronic and solid-state elements and pulsed sequential circuits. The last chapter on pulsed sequential circuits is the only one primarily written with the digital computer in mind. Wisely, the author does not attempt the problem of computer design, but seeks simply to show how the principles enunciated in earlier chapters are applicable to the special case of the computer.

At the end of each chapter there is a good selection of problems for the reader to attempt, but it is a pity that answers are not given, even to the problems which yield unique answers.

Criticism of this book can only be on minor points. For example, the contact coding described on page 29 to be that used in the remainder of the book is replaced by another only 2 pages later.

Then again, the author introduces the symbol A' to represent the complement instead of the more usual $\bar{A}$, the argument being vaguely reminiscent of that between engineers and mathematicians about j and i. The reviewer also considers that in article 10.1 a statement is made about binary coded decimal in relation to computers which many would question.

Summarizing, "Switching Circuits and Logical Design" is not written for any specific field, but for all who are concerned with switching problems, and can be recommended to students and experts alike.

R.G.

F.B.I. Register of British Manufacturers 1959

Pp. 1,140. Published for the Federation of British Industries by Kelly's Directories Ltd. and Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 42s. (post free).

Classified buyers' guide, addresses of companies, information about trade associations, brands and trade names, trade marks, etc. French, German and Spanish glossaries.

MEETINGS

I.E.E.

6th *February*. "Problems of Storing Transient Phenomena for Subsequent Analysis", discussion to be opened at 6 o'clock by Dr. P. Bauwens, M.R.C.S., M.R.C.P., and P. Styles.

9th *February*. "Dissemination and Assimilation of Technical Literature—A Growing Problem", discussion to be opened by J. K. Webb, M.Sc.(Eng.), B.Sc.Tech.

16th–17th *February*. Specialist discussion meetings on New Digital Computer Techniques. All wishing to attend must register; forms available on application to I.E.E.

18th *February*. "Ultrasonic Iconoscopes", by C. N. Smyth, M.A., B.Sc.(Eng.), M.B., B.Ch., and J. Sayers.

20th *February*. "The Measurement of Errors in Data Transmission for the Design of Detecting and Correcting Equipment", by V. J. Terry and E. P. G. Wright.

23rd *February*. "A Simple Investigation of the Cross-Modulation Distortion arising from the Pulling Effect in a Frequency-Modulated Klystron", by D. Gjessing. "Theory and Behaviour of Helix Structures for a High-Power Pulsed Travelling-Wave Tube", by G. W. Buckley, M.A., and J. Gunson, B.A. "A Multi-Cavity Klystron with Double-Tuned Output Circuit" by H. J. Curnow, B.Sc., and L. E. S. Mathias, M.Sc. "A Method for the Measurement of Very-High Q-Factors of Electromagnetic Resonators", by F. H. James.

5th *March*. "The Reliability and Life of Impregnated Paper Capacitors", by J. P. Pitts, B.Sc.(Eng.).

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30 *except the meetings on* 6th *and* 16th–17th *February.*

Brit. I.R.E.

13th *February*. "Some Instrumentation Problems in Medical Electronics with particular reference to Electromyography", by Peter Styles.

25th *February*. "Patents and the Radio Engineer", by E. D. Swann.

These meetings will be held at 6.30 *at the London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.*1.

The Television Society

6th *February*. "Master Control Room Techniques", by B. Marsden.

26th *February*. "A Colour Signal Encoder for Laboratory Use", by S. H. Cohen and P. C. Kidd.

These meetings will be held at 7 *o'clock at the Cinematograph Exhibitors' Association,* 164 *Shaftesbury Avenue, London, W.C.*2.

RUSSIAN TECHNICAL LITERATURE

English translations of some Russian journals are being published by Pergamon Press Inc. under the editorial auspices of the Massachusetts Institute of Technology and assisted by a grant-in-aid from the National Science Foundation. The journals being so published include *Telecommunications*, *Radio Engineering* and *Radio Engineering & Electronics*.

A specimen copy of *Telecommunications* and another of *Radio Engineering & Electronics* have been received. These copies have 94 and 183 pages respectively. The total number of pages per year is stated to be about 1,000 and 2,500.

The journals are available from Pergamon Press Ltd., 4 & 5 Fitzroy Square, London, W.1, by yearly subscriptions of £10.14s. and £16 respectively.

RADIO SHOW 1959

The Radio Industry Council announces that the 26th National Radio and Television Exhibition will be held at Earls Court, London, from Wednesday, 26th August to Saturday, 5th September (excluding Sunday).

TELEVISION SOCIETY EXHIBITION

From 3rd to 5th March inclusive, the annual Exhibition organized by the Television Society is to be held at the Royal Hotel, Woburn Place, London, W.C.1. The opening hours on the three days are respectively 11.30 a.m. to 8 p.m., noon to 8 p.m., and noon to 7 p.m.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for December* 1958

Date 1958 December	MSF 60 kc/s 1500 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	+ 1	− 2
2	+ 1	− 2
3	+ 1	− 2
4	+ 1	− 1
5	+ 1	− 1
6	+ 1	N.M.
7	+ 1	N.M.
8	+ 1	0
9	+ 1	0
10	+ 1	0
11	+ 1	0
12	+ 1	0
13	+ 1	N.M.
14	+ 1	N.M.
15	+ 1	+ 1
16	+ 1	+ 1
17	+ 1	+ 2
18	+ 1	+ 2
19	+ 1	+ 2
20	+ 1	N.M.
21	+ 1	N.M.
22	+ 1	+ 3
23	+ 2	+ 3
24	+ 2	+ 3
25	+ 2	N.M.
26	+ 1	N.M.
27	+ 2	N.M.
28	+ 2	N.M.
29	+ 1	+ 6
30	+ 2	+ 6
31	+ 2	+ 6

* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

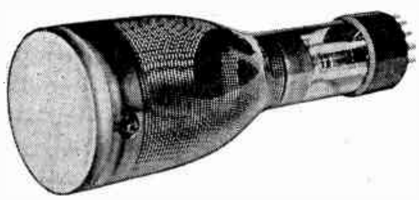
New Products

New Instrument Cathode-Ray Tube

A new high-brightness, high-sensitivity instrument cathode-ray tube, type DH10-78, employing helical post-deflection acceleration has been announced by Mullard Ltd. It has a 4 in. diameter flat screen and an overall length of only 12 in. It uses electrostatic focusing and deflection, and is suitable for double symmetrical operation.

The post-deflection accelerator consists of a high-resistance coating applied in the form of a helix to the inside of the bulb. It provides a continuous gradient of acceleration voltage, rising from a relatively low value near the deflector plates to a high final value, in order to minimize pattern distortion. Thus, pre-deflection acceleration may be kept low to ensure high sensitivity, while the final accelerating voltage may be made high to obtain maximum brightness. Under typical operating conditions, X and Y sensitivities of 30 and 10 V/cm respectively can be achieved with a post-acceleration voltage as high as 4 kV.

An additional feature of the tube is that the shield between the two pairs of deflection plates is brought out separately, so that the



shield voltage may be adjusted to correct pin-cushion and barrel distortion.

The screen of the DH10-78 is designed specifically for oscilloscopes and uses a high-sensitivity sulphide phosphor giving a bright actinic display. It has blue and yellow components giving, under normal excitation conditions, a green/white fluorescence of medium persistence.

Other points from the makers' specification are as follows: Heater rating, 6·3 V, 300 mA (a.c. or d.c.); $-V_g$(max) 200 V; $+V_g$(max) 0 (2-V peak); V_{a1+a3} (max) 3 kV; V_{a1+a3} (min) 1 kV; V_{a2}(max) 8 kV; base, B14A diheptal.

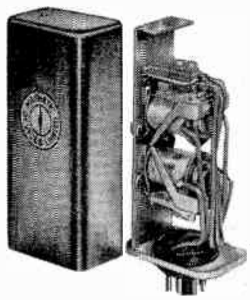
Mullard Ltd.,
Torrington Place, London, W.C.1.

Interlocking Relay

This Series 593 relay is an interlocking unit suitable for use in equipments such as flow-control motors and sorting mechanisms where it is desirable to switch between alternate circuits at regular or irregular intervals.

It comprises a pair of relays that are mutually interlocking so that one or the other is always locked in. Either or both coils can be supplied for a.c. or d.c. operation, and either or both relays may carry multiple contact assemblies.

The unit is available as an open relay on a two-position mounting bracket or, alternatively, enclosed or hermetically sealed. The enclosed and hermetically-sealed versions are available on 8, 9 or 11-pin plug-in octal-style headers and alternate



headers, such as the hook type, are also available.

The contacts are rated at 5 A 30 V d.c. or 250 V a.c. (resistive) with a maximum nominal coil voltage of 140 V d.c. or 250 V a.c.

Magnetic Devices Ltd.,
Exning Road, Newmarket, Suffolk.

Transistorized Level Control

The 'Pneutronic' on-off level controller, type PL1B, manufactured by Fielden Electronics Ltd., has been designed to control the level of all liquids or free-flowing solids which may be at high temperatures, high pressures or under corrosive conditions. It provides an output air pressure suitable for operating a standard pneumatic diaphragm control valve.

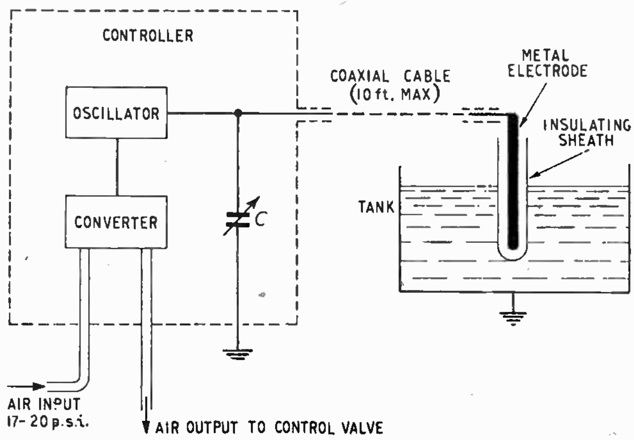
The instrument functions by measuring the change of electrical capacitance of an electrode (shown in the schematic diagram as a metal rod with an insulating sheath) mounted in the tank containing the liquid or granular solid. This electrode has a certain electrical capacitance to earth (when the tank is empty) which increases as the level of the material in the tank rises. The relative values of this capacitance and that of the variable capacitor *C* determine whether or not the circuit oscillates. (*C* is the control by which the operating level can be changed.) When the material in the tank rises to the desired level, the circuit oscillates and a d.c. signal is fed to the converter. This converter, which comprises a small electrically-operated valve, either permits air to pass to the control valve or reduces the air pressure of the control valve to atmospheric. This, of course, depends on whether the instrument is arranged as a direct-acting or reverse-acting controller.

If the material falls below the desired level, then the d.c. signal is removed from the shuttle valve and the output air pressure is reversed.

The frequency of operation is determined by several factors; the air capacity of the control-valve diaphragm motor, the air lines to this motor, and the rate at which the level of the material in the tank is changed.

Details from the makers' specification include: Operating voltage, 12 V d.c.; output, air pressure, zero or line pressure; dimensions, 11 in. × 12½ in. × 6¼ in.; weight, 17 lb. net.

Fielden Electronics Ltd.,
Wythenshawe, Manchester, Lancs.



Transistor Frequency Meter

This instrument, type 715, provides a simple yet accurate means of measuring the rotational speeds of modern high-speed mechanisms. When used in conjunction with the makers' photoelectric pickup, type 716, measurements can be made without any mechanical contact with the moving member.

When used as a tachometer, rotational speeds from 600 to 1,800,000 rev/min can be registered. Alternatively, cyclic events occurring at speeds between 10 and 30,000 c/s can also be investigated. The overall frequency coverage is divided into six ranges, each range having an accuracy of $\pm 2\%$ f.s.d. The meter can be operated from either self-contained dry batteries or from the mains. The switch provided for the selection of the appropriate power supply has an additional test position for checking the battery voltage.

In service, the signal is applied to the

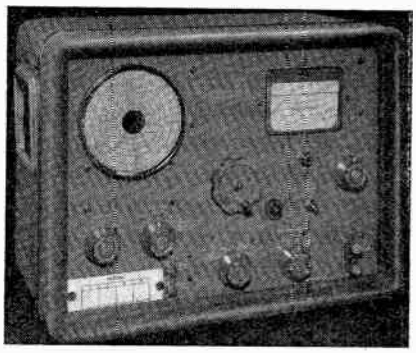
high-impedance earthed-collector input stage, followed by a shaping and a two-transistor bi-stable circuit, a resistance-capacitance differentiating network, a rectifier and, finally, to a moving-coil meter which is calibrated directly in terms of frequency.

Dawe Instruments Ltd.,
99 *Uxbridge Road, London, W.*5.

Carrier Deviation Meter

The TF 791D is the latest model in the Marconi Instruments series of direct-reading f.m. deviation meters. Suitable for both communication and broadcast systems, it has four deviation ranges of 5, 25, 75 and 125 kc/s full-scale, and a modulation-frequency range of 50 c/s to 35 kc/s. The calibrated carrier-frequency range extends from 4 to 1,024 Mc/s; freedom from microphony is ensured by the crystal lock facilities, which bring the internal f.m. noise down to -55 dB relative to 5-kc/s deviation, allowing reliable measurement of f.m. hum and noise in transmitters for v.h.f. sound broadcasting and mobile close-channel working. To aid measurement of asymmetric modulation, positive or negative deviation is displayed at the turn of a switch without the need for retuning.

The local oscillator operates on harmonics over six of its eight bands; this avoids the danger of frequency-pulling at the higher frequencies and ensures positive locking at any harmonic frequency of crystals within the relatively small fundamental range of



2 to 10 Mc/s. Crystals, up to a maximum of four, can be supplied to suit any specified carrier frequencies. Where the crystal-oscillator harmonic power is adequate, the local oscillator can be switched out of circuit and the crystal oscillator coupled direct to the mixer stage.

The instrument is basically an f.m. receiver using a counter-type discriminator for high stability; its demodulated output energizes a meter calibrated directly in deviation, and is also available externally at a separately-buffered outlet for aural monitoring or measurement of very low deviation.

The f.m. input is applied to a frequency-changer comprising a tunable local oscillator and a crystal mixer. The resultant i.f. signal is amplified and then converted to a square wave by a limiter circuit. From the square wave, unidirectional pulses are produced by rectification and differentiation. The pulses are integrated to give an output whose amplitude is proportional to deviation and whose frequency is that of the original modulation.

A high order of measurement accuracy is ensured by the provision of a crystal-controlled deviation standard; when the instrument is switched to 'Set Deviation', the r.f. and first i.f. stages are disconnected from the circuit and the remaining i.f. stages generate a crystal-controlled test signal of constant effective deviation against which the meter scale may be standardized.

Marconi Instruments Ltd.,
St. Albans, Herts.

Fine Wire Feeder

A device for unwinding fine wire at a constant tension from a reel, while the wire is being wound on to delicate apparatus, has been produced by Kinetrol Ltd.

The feeder, which stands 15 in. high, is said to be capable of handling the thinnest of wires, such as copper wire down to 0.0008 in. diameter or nichrome wire down to 0.0006 in. diameter, while the tension may be adjusted from a minimum of 5 gm to a maximum of 15 gm.

Two models are available, one which will feed 46 to 52 gauge wire at a speed of 60 in./sec, and another (a spring-controlled version) which will feed 42 to 46 gauge wire at a speed of 240 in./sec.

Basically, the wire is unwound from its reel by the action of winding it on to the article being manufactured. However, the following description gives a more detailed account of the mechanical operation.

The reel is stationary on the machine. An arm is pulled by the wire, rotating round the reel, to lead it away over a series of pulleys. Two of these pulleys are mounted on a beam and the wire hanging in a loop between them supports an adjustable jockey-weight attached to a jockey pulley. This jockey-weight, which applies the correct tension to the wire, is poised over a pan attached to the end of a lever. The lever actuates a brake controlling the spinning of the rotating arm. The slightest slackening of tension in the wire causes the jockey-weight to settle in the pan, applying the brake to the rotating arm and restoring the tension. As the tension increases, the jockey-weight tends to rise, releasing the brake until the tension is correct again. If the wire jams on the reel, the jockey-weight can rise several inches without increasing the tension and gives warning of trouble.

The makers claim that the machine allows a high winding speed at constant tension, without straining or breaking the wire, when winding on to an irregularly-shaped object. The rotating arm, which is the only moving part, is so light that its inertia will not break the wire.

Kinetrol Ltd.,
Trading Estate, Farnham, Surrey.

Magnetic Tape and Accessories

A wide range of Sonocolor magnetic-recording tape may now be obtained from Tape Recorders (Electronics) Ltd. A Sonocolor jointing and editing accessory kit is also available. It contains a splicer, three spools of 165-ft leader tape, jointing tape, jointing compound, ten safety clips, spare parts for splicer, etc. These items may be purchased separately.

Tape Recorders (Electronics) Ltd.,
784–788 *High Road, Tottenham, London, N.*17.

High-Vacuum Pump

A new general purpose high-vacuum pumping unit, type CU.100, has been produced by N.G.N. Electrical Ltd.

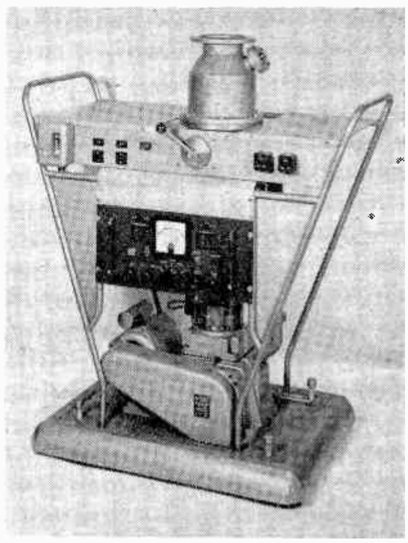
It is operated by means of a single-lever control, to obtain rough pump, fine pump, isolation and air admittance. Access to the high-vacuum chamber can be made, with a minimum of delay, when required.

The diffusion pump (which can be backed by a rotary pump of 3, 6 or 12 cu. ft per minute) has a pumping speed of 100 litres per second and can attain a pressure in excess of 5×10^{-6} mm of mercury.

Electrical failure is safeguarded by means of a protective unit which operates solenoid and air admittance valves. A water relay is also fitted to protect the diffusion pump if the water supply fails. The unit will re-start automatically.

Provision is made for fitting a liquid air trap and a Pirani direct-reading control.

N.G.N. Electrical Ltd.,
Avenue Parade, Accrington, Lancs.



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 selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publishers concerned.*

ACOUSTICS AND AUDIO FREQUENCIES

534.22-14 **328**

Viscosity Correction to the Velocity of Sound Measured by a Resonance Method.—C. Sălceanu & M. Zăgănescu. (*C. R. Acad. Sci., Paris*, 5th May 1958, Vol. 246, No. 18, pp. 2589–2591.) The velocity of sound in liquids is considered theoretically.

534.23+621.396.677.3 **329**

Signal/Noise Performance of Super-directive Arrays.—D. G. Tucker. (*Acustica*, 1958, Vol. 8, No. 2, pp. 112–116.) A super-directive array is defined as one whose effective aperture exceeds its physical aperture. Such an array has a directivity index somewhat better than that of an ordinary array, but its noise factor is much worse. The concept of super-directivity is extended to bearing-determining arrays which give a null response on the axis. See also 2 of January.

534.232.089.6 **330**

Calibration of Electroacoustic Transducers Operating under Increased Pressures of the Ambient Medium.—I. P. Neroda. (*Elektrosvyaz'*, Oct. 1957, No. 10, pp. 57–61.) Absolute calibration of transducers by the reciprocity method using a tube is considered. The theory of the method is discussed and a suitable experimental procedure is suggested.

534.78 : 621.395 **331**

Artificial Auditory Recognition in Telephony.—E. E. David, Jr. (*IBM J. Res. Developm.*, Oct. 1958, Vol. 2, No. 4, pp. 294–309.) A discussion of the possibility of machine recognition of acoustic patterns such as spoken commands. Techniques involve time division into recognizable discrete units and the study of spectrograms. Rules for interpreting these were successful in experiments of speech recognition.

534.845 **332**

A Note on Reciprocity in Linear Passive Acoustical Systems.—J. H. Janssen. (*Acustica*, 1958, Vol. 8, No. 2, pp. 76–78.) "It is shown theoretically that linear acoustical passive systems can behave as so-called reciprocity-violating systems; an example of such a system is a sound-absorbing material that is porous as well as flexible."

621.395.623.7 **333**

A Survey of Performance Criteria and Design Considerations for High-Quality Monitoring Loudspeakers.—D. E. L. Shorter. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 607–623. Discussion.)

621.395.623.73 **334**

Rigidity of Loudspeaker Diaphragms.—D. A. Barlow. (*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 564–569.) The advantages of a sandwich construction are discussed.

AERIALS AND TRANSMISSION LINES

621.315.2.029.5/.6 **335**

High-Frequency Cables.—R. Goldschmidt. (*Bull. schweiz. elektrotech. Ver.*, 2nd Aug. 1958, Vol. 49, No. 16, pp. 708–716.) General theory, materials, and methods of construction and testing are considered.

621.315.212.002.2 : 621.316.992 **336**

Solderless Grounding for Braided Shields.—F. C. March. (*Electronic Equipm. Engng*, June 1958, Vol. 6, No. 6, pp. 48–50.) A technique is described for splicing sections of coaxial cable or connecting earth leads to braid, using a mechanical crimping tool.

621.315.212.029.64 **337**

Factors affecting Attenuation of Solid-Dielectric Coaxial Cables above 3 000 Megacycles.—J. R. Hannon. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CP-3, No. 3, pp. 99–105. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 573.)

621.372.2.029.6 : 621.372.43 **338**

A Very-Wide-Band Balun Transformer for V.H.F. and U.H.F.—T. R. O'Meara & R. L. Sydnor. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1848–1860.) The structure described may be used as a phase inverter, a differential transformer, or as a balun transformer. The insertion loss varies from 1 to 2 dB, between 5 and 1 000 Mc/s.

621.372.8 : 621.372.2.092 **339**

Waveguide Coils make Compact Delay Lines.—R. R. Palmisano & A. Sherman. (*Electronics*, 24th Oct. 1958, Vol. 31, No. 43, pp. 88–89.) A 240-ft delay-line unit consisting of six tightly wound 40-ft coils of rectangular waveguide has an insertion loss of 21 dB with maximum voltage s.w.r. of 1·5.

621.372.85 **340**

Low-Loss Structures in Waveguides.—M. F. McKenna. (*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 470–472.) A method is described for evaluating the equivalent-circuit parameters of obstacles in waveguides by measuring the field in the guide terminated by a variable reactance.

621.372.852.22 **341**

Electromagnetic Wave Propagation in Cylindrical Waveguides containing Gyromagnetic Media.—R. A. Waldron. (*J. Brit. Instn Radio Engrs*, Oct.–Dec. 1958, Vol. 18, Nos. 10–12, pp. 597–612, 677–690 and 733–746.) A comprehensive analysis is given, with a large number of computed results of cut-off points and phase constants for a guide containing a concentric ferrite rod of arbitrary radius. Faraday rotation, power flow, and losses are discussed.

621.396.677.3 + 534.23 **342**

Signal/Noise Performance of Super-directive Arrays.—Tucker. (See 329.)

621.396.677.3 **343**

Methods of Calculating the Horizontal Radiation Patterns of Dipole Arrays around a Support Mast.—P. Knight. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 548–554.) A discussion of some theoretical methods and their limitations. The methods include the technique developed by Carter (1928 of 1944), the infinite-plane method, the induced-current method, and the diffraction method. The results are compared with experimental patterns.

621.396.677.81 **344**

A Concentric-Feed Yagi.—C. R. Graf. (*QST*, Nov. 1958, Vol. 42, No. 11, pp. 24–25.) An impedance matching technique is described in which a $3\lambda/4$ coaxial line is inserted in one side of the folded-dipole driven element.

AUTOMATIC COMPUTERS

681.142 **345**

Computation in the Presence of Noise.—P. Elias. (*IBM J. Res. Developm.*, Oct. 1958, Vol. 2, No. 4, pp. 346–353.) An analysis of the problem of performing reliable computation with elements which are themselves unreliable. The effects of the coding procedures on the reliability are discussed.

681.142 **346**

Control Apparatus for a Serial Drum Memory.—D. S. Kamat. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 634–639.) A detailed circuit description of equipment used for obtaining design data for fast track switching on a serial drum magnetic storage device.

681.142 **347**

Matrix Programming of Electronic Analogue Computers.—R. E. Horn & P. M. Honnell. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 420–426. Discussion, pp. 426–428.) The technique is based on establishing a correspondence between the matrically formed differential equations and the computer networks. This simplifies the setting up of the computer and reduces the possibility of errors. Examples are given of the use of the method.

681.142 **348**

A Magnetic-Drum Store for Analogue Computing.—J. L. Douce & J. C. West. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 577–580.) Large storage capacity combined with relatively short access time is obtained. The drum can be used for generating nonlinear functions and as a delay network.

681.142 : 538.632 **349**

An Electrical Multiplier utilizing the Hall Effect in Indium Arsenide.—R. P. Chasmar & E. Cohen. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 661–664.) Details are given of the construction and performance of a multiplier for computer applications in which an InAs Hall plate is mounted in the gap in a ferrite core.

681.142 : 538.632 : 537.311.33 **350**

Multiplication by Semiconductors.—C. Hilsum. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 664–666.) The modes of operation of analogue computer multipliers using the Hall effect in semiconductors are discussed. Some experimental multipliers and the results obtained with them are described. The application of the magnetoresistive effect is outlined.

681.142 : 621.314.7 : 621.318.134 **351**

The Design of Logical Circuits using Transistors and Square-Loop Ferrite Cores.—A. F. Newell. (*Mullard tech. Commun.*, Oct. 1958, Vol. 4, No. 34, pp. 110–120.) Some basic circuits are reviewed and a range of practical designs and design procedures is given.

681.142 : 621.318.5 **352**

Some Aspects of the Network Analysis of Sequence Transducers.—J. M. Simon. (*J. Franklin Inst.*, June 1958, Vol. 265, No. 6, pp. 439–450.) An algebraic formulation is presented for use in the analysis of networks of the synchronous type of sequence transducer. See also 28 of 1956 (Mealy).

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049 : 621.3-71 **353**

Design and Performance of Air-Cooled Chassis for Electronic Equipment.—M. Mark & M. Stephenson. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. CP-3, No. 2, pp. 38–44. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, pp. 112–113.)

621.3.049.7 **354**

Future Electronic Components.—G. W. A. Dummer. (*Wireless World*, Dec. 1958 Vol. 64, No. 12, pp. 591–593.) To obtain substantial reduction in size, components will probably consist of thin films of resistive, dielectric, magnetic and conductive materials, and film-type or flat-plate semiconductor devices will be used.

621.3.049.75 **355**

Foil-Clad Laminates in Printed Circuitry.—D. K. Rider. (*Metal Progr.*, Sept. 1958, Vol. 74, No. 3, pp. 81–85.)

621.314.21.001.2 **356**

Simplified Mains-Transformer Design.—H. D. Kitchen. (*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 582–585.) The power-handling capabilities of various laminations are tabulated and the design procedure is illustrated by an example.

621.314.22 : 621.3.018.7 **357**

Measurement of Parameters Controlling Pulse Front Response of Transformers.—P. R. Gillette, K. Oshima & R. M. Rowe. (*Trans. Inst. Radio Engrs*, March 1956, Vol. CP-3, No. 1, pp. 20–25. Abstract, *Proc. Inst. Radio Engrs*, June 1956, Vol. 44, No. 6, Part 1, p. 832.)

621.318.57 : 621.327.42 : 681.142 **358**

A Study of the Neon Bulb as a Non-linear Circuit Element.—C. E. Hendrix. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. CP-3, No. 2, pp. 44–54. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.319.45 **359**

A Method of Electrolytic Etching of Tantalum for Capacitor Use.—I. Sanghi. (*Curr. Sci.*, Aug. 1958, Vol. 27, No. 8, pp. 297–298.) Etching of foil in a bath of trichloroacetic acid, sodium trichloroacetate and methanol produced an increase of 300–400 % in capacitance relative to a capacitor formed of unetched foil.

621.372 : 531.314.2 **360**

The Lagrange Equations in Electrical Networks.—F. L. Ryder. (*J. Franklin Inst.*, July 1958, Vol. 266, No. 1, pp. 27–38.)

621.372 : 534.213-8 **361**

Ultrasonic Mercury Delay Lines.—C. F. Brockelsby. (*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 446–452.) The construction and characteristics of lines and transducers are discussed.

621.372.029.6 : 621.374 **362**

Analysis of Milimicrosecond R.F. Pulse Transmission.—M. P. Forrer. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1830–1835.) The analysis assumes a quadratic approximation for the complex propagation constant and a transmitted pulse with Gaussian envelope. The theory is applicable to uniform microwave transmission systems and relates pulse shapes with the c.w. properties of the system.

621.372.413.029.64 : 621.3.049.75 **363**

Fabrication Techniques for Ceramic X-Band Cavity Resonators.—M. C. Thompson, Jr, F. E. Freethey & D. M. Waters. (*Rev. sci. Instrum.*, Oct. 1958, Vol. 29, No. 10, pp. 865–868.) Techniques are described for constructing cavity resonators for the X-band from low-thermal-expansion ceramics. A variety of mechanical arrangements is discussed. *Q* values as high as 14 000 and frequency/temperature coefficients as low as 1 part in 10^8 per °C have been obtained using simple processes.

621.372.5 **364**

A Correlation between Stagger-Tuned and Synchronously Tuned Coupled Circuits.—J. B. Rudd. (*A. W. A. tech. Rev.*, 1958, Vol. 10, No. 3, pp. 101–109.) Application of a network theorem enunciated by Green (54 of 1958).

621.372.54 **365**

Time-Symmetric Filters.—L. R. O. Storey & J. K. Grierson. (*Electronic Engng*, Oct. & Nov. 1958, Vol. 30, Nos. 368 & 369, pp. 586–592 & 648–653.) Methods are described for simulating filters that have impulse responses symmetrical in time, and their application to the analysis of gliding tones is discussed. The response of a narrow-band time-symmetric filter to a gliding tone is evaluated and shown to consist of an oscillation bounded by a slowly varying envelope. The shape is governed by a parameter involving the bandwidth of the filter and the rate of variation of the instantaneous frequency of the tone.

621.372.54(083.57) **366**

Design Curves for Simple Filters.—D. J. H. Maclean. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 654–660.) Charts are given for the design of simple *LC* ladder filters with either Butterworth or Tchebycheff type of response.

621.372.54(083.57) **367**

Nomographs for Designing Elliptic-Function Filters.—K. W. Henderson. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1860–1864.)

621.372.543.3 : 621-526 **368**

Simplified Design of Resonant Notch Filters for Servo Applications.—J. P. Jagy. (*Electronic Equipm. Engng*, April 1958, Vol. 6, No. 4, pp. 48–52.) Data are given in graphical form for the design of stagger-tuned constant-*k* band-stop filters. Applications to error-rate damping in servo systems are discussed.

621.372.57 **369**

The Reduction of Low-Frequency Noise in Feedback Integrators.—E. M. Dunstan & M. J. Somerville. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 532–544.) A considerable improvement in signal/noise ratio, compared with that of a conventional direct-coupled integrator, is obtained either by using an error amplifier containing a single *CR* coupling or by applying phase correction to the output from a low-accuracy direct-coupled integrator.

621.373.421.13 **370**

Thermally Compensated Crystal Oscillators.—R. A. Spears. (*J. Brit. Instn Radio Engrs*, Oct. 1958, Vol. 18, No. 10, pp. 613–620.) Frequency stability is achieved by using thermistors in a temperature-sensitive phase-shifting network incorporated in the oscillator circuit. A stability of about 1 in 10^6 is obtained over a wide frequency range without thermostats or ovens. It is suggested that even better compensation would be achieved by attaching the thermistor bead to the crystal.

621.373.421.13 : 621.3.018.41 (083.74) **371**

A Quartz Servo Oscillator.—N. Lea. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1835–1841.) A description of a 5-Mc/s quartz oscillator stable to 2 parts in 10^{10} with dual stabilization by resonant-loop balance and bridge-operated servo control.

621.373.421.13.001.4 **372**

Checking Crystal Oscillators.—D. J. Spooner. (*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 594–596.) Simple measurements are suggested to ensure that a crystal will oscillate in a specified circuit without damage, and at the required frequency.

621.373.43 **373**

A Constant-Amplitude Random-Function Generator.—G. A. Hellwarth. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 443–452.) A generator is described for producing a wave shape with constant peak-to-peak amplitude but whose path between peaks is random. Circuits for random saw-tooth, triangular, cosine and square waves operating in the a.f. range are described.

621.373.43 + 621.374.32] : 621.387 **374**

Some Novel Circuits employing Cold-Cathode Tubes.—R. S. Sidorowicz. (*Electronic Engng*, Nov. & Dec. 1958, Vol. 30, Nos. 369 & 370, pp. 624–629 & 697–701.) The circuits described are based on cold-cathode diodes and triodes, the diode being used to stabilize the anode breakdown voltage of the triodes. Details are given of a stable relaxation oscillator, a voltage discriminator, trigger circuits, a staircase-waveform generator, and decade counters.

621.373.43 : 621.396.96 **375**

Pulse - Forming Networks. — J. W. Trinkaus. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. CP-3, No. 2, pp. 63–66. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.373.44 : 535.33 **376**

The Excitation of Ionic Spectra by 100-kW High-Frequency Pulses.—L. Minnhagen & L. Stigmark. (*Ark. Fys.*, 5th Feb. 1958, Vol. 13, Part 1, pp. 27–36. In English.) Equipment previously described (1592 of 1955) has been developed and provided with a power amplifier containing a 25-kW water-cooled triode. Pulse powers up to approximately 100 kW are applied to a discharge tube with pulse duration 80 μs and repetition frequency 40/sec.

621.374.32 **377**

A Distributed-Circuit Pulse Height Analyser.—A. Boucherie & J. Mey. (*J. Phys. Radium*, Jan. 1958, Vol. 19, No. 1, pp. 98–99.) A multichannel analyser using distributed circuits and having a resolving time of 0·2 μs is briefly described.

621.374.43 **378**

Frequency Dividers using Transistors.—M. Z. Tseĭtlin. (*Elektrosvyaz'*, Sept. 1957, No. 9, pp. 33–41.) Regenerative frequency dividers using junction and point-contact transistors are described and results are given of an experimental investigation.

621.375.2.029.33 **379**

Video Amplifier Design using the PCL 84.—P. L. Mothersole. (*Mullard tech. Commun.*, July 1958, Vol. 4, No. 31, pp. 2–6.) Optimum gain is achieved by the use of anode compensation. Bias methods for this condition are discussed.

621.375.2.029.63 **380**

Ultra-High-Frequency Power Amplifiers.—J. Dain. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 513–522.) A general review of the design and construction of power amplifiers operating in the range 300–3 000 Mc/s. Travelling-wave valves designed for a band-width of one octave are limited in their mean power output by overheating of the helix. Backward-wave amplifiers based on crossed-field interaction require a variable beam voltage for wide-band operation but have a high efficiency and low operating voltage.

621.375.3 **381**

A Note on the Design of Transductors for Maximum Power Transfer.—J. C. R. Heydenrych. (*Trans. S. Afr. Inst. elect. Engrs*, Dec. 1957, Vol. 48, Part 12, pp. 370–377.) Magnetic amplifiers giving maximum power output for a given size of core are designed by known graphical and numerical methods. Theoretical and experimental results are compared.

621.375.3 **382**

A.C.-Controlled Magnetic Amplifiers.—E. W. Lehtonen & E. A. Cronauer. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 476–480.) A method is described for controlling full-wave amplifiers with a.c. signals. The response characteristic is similar to that of d.c.-controlled amplifiers and high gain is achieved with no current-limiting resistors or demodulator. The method is based on cancelling the induced e.m.f. in the control circuit.

621.375.3 (083.7) **383**

Proposed Standard Terms and Definitions for Magnetic Amplifiers.—(*Commun. & Electronics*, Sept. 1958, No. 38, pp. 429–431. Discussion, pp. 431–432.) An A.I.E.E. committee report on the terminology used. 38 terms are defined.

621.375.4.029.3 **384**

A 4·5-W Sliding-Bias Amplifier using an OC16.—J. F. Pawling & P. Tharma. (*Mullard tech. Commun.*, July 1958, Vol. 4, No. 31, pp. 19–28.) The two-stage circuit described and analysed in detail gives nearly twice the output power obtainable in conventional class-A operation, using a heat sink of the same size.

621.375.4.029.33 **385**

Video Amplifiers using Alloy Junction Transistors.—K. Holford & L. M. Newall. (*Mullard tech. Commun.*, Oct. 1958, Vol. 4, No. 34, pp. 94–105.) Grounded-emitter video amplifier stages are analysed theoretically using compensated and uncompensated circuits. Abacs and charts are given to facilitate design procedure for multistage amplifiers, together with the practical design of an amplifier with 80 dB gain and 1·5 Mc/s bandwidth.

621.375.432 **386**

Pulse Amplifier with Nonlinear Feedback.—L. H. Dulberger. (*Electronics*, 7th Nov. 1958, Vol. 31, No. 45, pp. 86–87.) The transistor amplifier described provides constant output over a 38 dB range of input signals.

621.375.432 : 621.395.625.3 **387**

Transistor Tape Preamplifier.—P. F. Ridler. (*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 572–573.) The play-back head is used as the inductance in a feedback inductance-resistance integrating circuit. A 70 dB signal/noise ratio is obtained and the frequency response is flat within ± 2 dB from 50 c/s to 15 kc/s.

621.375.9 : 538.569.4.029.6 **388**

Proposal for a Maser Amplifier System without Nonreciprocal Elements.—S. H. Autler. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1880–1881.) A system noise temperature of 30°K or less should be obtainable by using two matched masers and a lossless power-dividing network such as a hybrid T.

621.375.9 : 538.569.4.029.64 **389**

Characteristics of the Beam-Type Maser: Part 2.—K. Shimoda. (*J. phys. Soc. Japan*, Aug. 1958, Vol. 13, No. 8, pp. 939–947.) An experimental investigation of the characteristics of an ammonia maser for use as a frequency standard. Measurements of the effect of focusing voltage and cavity tuning on frequency are compared with theory, and the effect of the velocity spread of the molecules is estimated. Part 1 : 707 of 1958.

621.375.9.029.6 : 621.3.011.23 : 621.396.61 **390**

Parametric Amplifier ups Scatter Range.—(*Electronics*, 7th Nov. 1958, Vol. 31, No. 45, p. 96.) A Si-diode parametric amplifier [see 79 of January (Weber)] is used, and extends the range of a 900-Mc/s link from 250 to 350 miles. Receiver noise factor is reduced from 8 to 1 dB.

GENERAL PHYSICS

535 : 621.383 **391**

The Wider Scope of Optics.—K. M. Greenland. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 278–288.) The development and applications of new optical and optical-electronic devices are considered. 31 references.

537.226.33 **392**

The Application of Onsager's Theory to Dielectric Dispersion.—N. E. Hill. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 532–536.) A new equation for the complex dielectric constant, which includes the proper effect of the reaction field, is developed for the case of an alternating applied field. It is shown to yield results very close to those obtained with the simple Debye equation for the complex dielectric constant.

537.312.62 **393**

Experimental Evidence for an Energy Gap in Superconductors.—M. A. Biondi, A. T. Forrester, M. P. Garfunkel & C. B. Satterthwaite. (*Rev. mod. Phys.*, Oct. 1958, Vol. 30, No. 4, pp. 1109–1136.)

537.52 : 537.56 **394**

Wire–Cylinder Electric Discharges in Air in Relation to the Space-Charge Field-Emission Hypothesis.—H. Ritow. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 193–225.) Study of experimental data on wire–cylinder discharges in relation to the field-emission hypothesis yields graphical and arithmetic methods of finding the effective field at the time of flash initiation and the space-charge field at the wire. The effective field is of the order of 10^6 V/cm and is interpreted as a measure of the emission field of the cathode metal or of the ionized air. See also 2061 of 1958.

537.525.029.5 **395**

Ultra-High-Frequency Gas Breakdown between Ragowski Electrodes.—W. A. Prowse & J. L. Clark. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 625–634.) Breakdown voltage, electrode spacing, electrode size and gas pressures are observed for air, H_2, N_2 and Ne at 9·5 Mc/s.

537.533 : 538.63 **396**

The Relativistic Flow of Electrons in Parallel and Radial Straight Lines with no Externally Imposed Magnetic Field.—A. R. Lucas. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 245–250.) Analysis is given of the possibility of producing relativistic electron flows. It is shown that they could not start from a cathode surface where the electrons have zero speed. The analysis also applies to flows, normally considered nonrelativistic, in diodes where the linear dimensions of the electrode surfaces are large compared with the electrode spacing [see e.g. 3037 of 1958 (Meltzer)].

537.533 : 621.385.029.6 **397**

Electron Cooling by Heat Exchange.—R. C. Knechtli & W. Knauer. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1513–1514.) A new method for obtaining homogeneous electron streams is described. By making the electron stream part of a plasma, beams with densities of up to 10^{-3} A/cm² and currents exceeding 10^{-3} A have been achieved.

537.533.7 : 535.417 **398**

Coherence Requirements for Interferometry.—G. D. Kahl & F. D. Bennett. (*Rev. mod. Phys.*, Oct. 1958, Vol. 30, No. 4, pp. 1193–1196.) An analysis of the theory of double-beam interferometry, based on that of D. Gabor (*Rev. mod. Phys.*, July 1956, Vol. 28, No. 3, pp. 260–276). Gabor's restrictive assumptions are clarified and generalized.

537.533.71 : 621.385.833 **399**

Energy Spectrum of a 40-kV Electron Beam 'Reflected' by a Metallic Object.—F. Pradal & R. Saporte. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2880–2883.) The energy spectra of electrons reflected from different metallic targets are investigated by means of a magnetic spectrograph.

537.56 : 537.29 : 538.69 **400**

Transport Phenomena in a Completely Ionized Two-Temperature Plasma.—S. I. Braginskiĭ. (*Zh. eksp. teor. Fiz.*, Aug. 1957, Vol. 33, No. 2(8), pp. 459–472.) A theoretical investigation of particle motion and heat transfer in a plasma of electrons and positive ions under the influence of both electric and magnetic fields, the electron and ion temperatures being considered different.

537.56 : 538.12 **401**

The Amplification of a Magnetic Field by a High-Current Discharge.—R. J. Bickerton. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 618–624.) It is shown theoretically that a helical current flow discharge is set up by a longitudinal magnetic field in which the plasma pressure is balanced by electrodynamic forces. The direction of the helix is such that the initial longitudinal field is amplified. Some experimental evidence supports this theory.

537.562 **402**

Transport Phenomena in Completely Ionized Gas considering Electron-Electron Scattering.—M. S. Sodha & Y. P. Varshni. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1203–1205.) "Hall mobility and other transport properties of electrons in a completely ionized gas have been investigated when a magnetic field is applied, taking into account electron-electron scattering. Results have been presented for different mean ionic charges."

538.12 : 538.221 **403**

The Magnetic Fields of a Ferrite Ellipsoid.—R. A. Hurd. (*Canad. J. Phys.*, Aug. 1958, Vol. 36, No. 8, pp. 1072–1083.) "Approximate expressions are found for the internal and the adjacent external magnetic fields of a small ferrite ellipsoid under plane-wave excitation. Consideration is given to the variation of apparent susceptibility with the size of the ferrite."

538.3 **404**

Classical Electrodynamics as a Distribution Theory: Part 2.—J. G. Taylor. (*Proc. Camb. phil. Soc.*, April 1958, Vol. 54, Part 2, pp. 258–264.) Part 1: 2027 of 1956.

538.3 : 535.13 **405**

Application of Distributions to the Equations of Maxwell and Helmholtz.—M. Bouix. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2858–2860.) Distribution theory is applied to give a new concept of an element of current.

538.3 : 535.13 **406**

Singular Electromagnetic Induction.—Pham Mau Quan. (*C. R. Acad. Sci., Paris*, 12th May 1958, Vol. 246, No. 19, pp. 2734–2737.) Extension of the concepts discussed in 3417 of 1958.

538.3 : 535.13 **407**

Algebraic Study of the Electromagnetic Tensor in the Presence of Induction.—L. Mariot & Pham Mau Quan. (*C. R. Acad. Sci., Paris*, 28th May 1958, Vol. 246, No. 21, pp. 3018–3020.) See 406 above.

538.561 : 537.122 : 523.7 **408**

Electromagnetic Radiation from Electrons Rotating in an Ionized Medium under the Action of a Uniform Magnetic Field.—R. Q. Twiss & J. A. Roberts. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 424–446.) The radiation is shown to be predominantly in the extraordinary mode. At the harmonics of the gyrofrequency of the fast electron the power radiated in the ordinary mode is a small percentage of that in the extraordinary mode, but at the fundamental gyrofrequency it is lower by a factor $\approx 10^{-2} (v_0/c)^4$, where v_0 is the velocity of the fast electron and c is the velocity of light. The gyro theory of the sun's nonthermal radiation is discussed. This mechanism cannot explain the phenomena associated with noise bursts of Type II and III although it is conceivable that Type I bursts may be of gyro origin.

538.566 **409**

Theory of Electromagnetic Waves in a Crystal with Excitons.—S. I. Pekar. (*J. Phys. Chem. Solids*, 1958, Vol. 5, Nos. 1/2, pp. 11–22.) See 3058 of 1958.

538.566 : 535.312 **410**

The Characteristics of an Electromagnetic Wave Reflected from a Moving Object.—C. F. Cole, Jr. (*J. Franklin Inst.*, June 1958, Vol. 265, No. 6, pp. 463–471.)

538.566 : 535.42 **411**

Diffraction by a Wide Slit and Complementary Strip.—R. F. Millar. (*Proc. Camb. phil. Soc.*, 17th Oct. 1958, Vol. 54, Part 4, pp. 479–511.) The diffraction of E- and H-polarized waves by an infinite slit is considered. Induced current densities, aperture and far fields, and the transmission coefficient, are calculated in the form of infinite series in inverse powers of the slit-width/wavelength ratio. The solution for diffraction by a strip is also obtained. A comparison is made with previous results; this method appears to provide accurate information when the slit width is greater than a wavelength.

538.566 : 535.43 **412**

Variational Principles in High-Frequency Scattering.—R. D. Kodis. (*Proc. Camb. phil. Soc.*, 17th Oct. 1958, Vol. 54, Part 4, pp. 512–529.) Two variational principles are formulated for two-dimensional scattering by obstacles. The more successful of these treats the obstacle like an aperture coupling two half-spaces. The zero-order calculation for the cross-section of a circle is found to have the correct $(ka)^{-2/3}$ frequency dependence.

538.566 : 535.43 **413**

Scattering from a Small Anisotropic Ellipsoid.—R. A. Hurd. (*Canad. J. Phys.*, Aug. 1958, Vol. 36, No. 8, pp. 1058–1071.) "Scattering of an electromagnetic wave by a small ellipsoid having tensor permeability and dielectric properties is dealt with by expanding the fields as power series in λ^{-1}. Consideration has been given to the first three terms of the expansion."

538.566.2 **414**

Contribution to the Theory of Electromagnetic Wave Propagation in Media with Random Heterogeneities of the Refractive Index.—V. V. Merkulov. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 1051–1055.) A correlation function is derived which can be used in connection with e.m. wave diffraction problems.

538.566.2 : 535.43] + 534.26 **415**

Scattering of Plane Waves by Locally Homogeneous Dielectric Noise.—R. A. Silverman. (*Proc. Camb. phil. Soc.*, 17th Oct. 1958, Vol. 54, Part 4, pp. 530–537.) When plane waves are scattered by locally homogeneous dielectric noise (random refractive-index fluctuations) and observed in the Fraunhofer region, it is found that the local structure of the noise determines the average scattered power received at a fixed point, whereas its overall structure determines the space correlations of the radiation received at two different points.

538.569.4.029.6 : 535.33.08 **416**

Criteria determining the Design and Performance of a Source-Modulated Microwave Cavity Spectrometer.—R. W. R. Hoisington, L. Kellner & M. J. Pentz. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 537–544.) "An analysis is given of the radio- and audio-frequency modulation method employed in microwave spectroscopy. The results of the theory are compared with measurements taken on a 8-mm microwave spectrometer and are found to be in close agreement. The calculations are extended to include the case where an absorbing gas is enclosed in a resonant cavity."

538.569.4.029.64 : 539.2 **417**

Direct Measurement of Electron Spin-Lattice Relaxation Times.—C. F. Davis, Jr, M. W. P. Strandberg & R. L. Kyhl. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1268–1272.) A discussion of the experimental problems encountered in making spin-lattice relaxation measurements in electron paramagnetic systems at low temperatures. Gadolinium and chrome ion spin-lattice relaxation times are given. The relation of these spin-lattice relaxation times to relaxation times measured in the frequency domain by observing a saturation parameter is discussed.

539.2 **418**

The Electron Structure of Transition Metals and Alloys and Heavy Metals.—J. Friedel. (*J. Phys. Radium*, June 1958, Vol. 19, No. 6, pp. 573–581.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 **419**

Energy Spectrum of Particles Bombarding the Earth.—B. J. O'Brien. (*Nature, Lond.*, 23rd Aug. 1958, Vol. 182, No. 4634, p. 521.) The estimated fluxes of interstellar and auroral particles appear to fit the straight-line portion of the cosmic-ray integral energy spectrum extrapolated to lower energies.

523.164 **420**

Radio Astronomy.—S. Khaikin. (*Radio, Mosk.*, Nov. 1957, No. 11, pp. 25–27.) A brief description of the fields of exploration opened by developments in radio astronomy.

523.164 **421**

On the Radio Emission of Hydrogen Nebulae.—C. M. Wade. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 388–399.) Random variations in electron density and electron temperature through the nebulae are shown to alter the optical depth. Radio emission of Strömgren spheres is also discussed and an empirical method for determining Strömgren's constant is described.

523.164 **422**

An Investigation of the Strong Radio Sources in Centaurus, Fornax, and Puppis.—K. V. Sheridan. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 400–408.)

523.164 **423**

A Catalogue of Radio Sources between Declinations +10° and −20°.—B. Y. Mills, O. B. Slee & E. R. Hill. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 360–387.)

523.164.32 **424**

Investigations of Persistent Solar Sources at Centimetre Wavelengths.—M. R. Kundu. (*C. R. Acad. Sci., Paris*, 12th May 1958, Vol. 246, No. 19, pp. 2740–2743.) Measurements of the brightness distribution and apparent size of solar r.f. sources made at 3·2 cm λ with an interferometer [2733 of 1957 (Alon et al.)] show that small apparent diameters are associated with periods of eruptive solar activity.

523.164.32 **425**

The Dimensions of Sources of Bursts of Solar Radiation at Centimetre Wavelengths.—M. R. Kundu. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2852–2855.) Sources are classified, the growth and decay in their apparent size are observed and their equivalent temperature estimated on the basis of interferometer measurements at 3 cm λ.

523.164.4 : 523.755 **426**

Outer Corona of the Sun.—V. V. Vitkevich. (*Priroda, Mosk.*, Dec. 1957, No. 12, pp. 15–20.) Radio emissions from the Crab nebula passing through the corona are observed by an interferometric method. The results indicate that the outer corona extends to a distance of 20 sun radii.

523.164.4 : 551.510.535 **427**

Amplitude Scintillation of Extra-terrestrial Radio Waves at Ultra High Frequency.—H. C. Ko. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1872–1873.) Measurements are described which show that at latitude 40°N, ionospheric scintillation effects are still significant at 915 Mc/s when the radio star is near the northern horizon.

523.7 : 538.561 : 537.122 **428**

Electromagnetic Radiation from Electrons Rotating in an Ionized Medium under the Action of a Uniform Magnetic Field.—Twiss & Roberts. (See 408.)

523.72 : 621.396.822 **429**

Ionizing Radiation associated with Solar Radio Noise Storm.—K. A. Anderson. (*Phys. Rev. Lett.*, 1st Nov. 1958, Vol. 1, No. 9, pp. 335–337.) Comparison of records obtained during a storm on 22nd August 1958 from three balloon-borne detectors, a single counter, counter telescope and ion chamber, indicates the appearance of protons with kinetic energy of 170 MeV.

523.72 : 621.396.822 **430**

Solar Brightness Distribution at a Wavelength of 60 Centimetres : Part 2—Localized Radio Bright Regions.—G. Swarup & R. Parthasarathy. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 338–349.) Observations were taken with a 32-element interferometer with a beam width of 8·7 min of arc, from July 1954 to March 1955. Sources of radio brightness were found to be closely correlated with sunspot areas. Their estimated size lay between 3 and 6 min of arc. Sometimes their slowly varying component showed marked changes in intensity over periods of half an hour. The largest radio-brightness temperatures measured were about 10^7°K. Part 1 : 1707 of 1956.

523.75 : 523.164.32 **431**

Flare-Puffs as a Cause of Type III Radio Bursts.—R. G. Giovanelli. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 350–352.) Type III bursts occur within ± 2 min of two-thirds of the flare-puffs. Most puffs are followed by surges and this suggests two ejections of differing velocities : one, at about 1/5 the velocity of light, causing the burst, and the other, at 100 km/sec causing the surge. See also 1715 of 1958 (Loughhead et al.).

523.75 : 523.164.32 **432**

Optical Observations of the Solar Disturbances causing Type II Radio Bursts.—R. G. Giovanelli & J. A. Roberts. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 353–359.) Type II radio bursts have been identified with ejections having velocities exceeding that of sound in the corona for events near the limb, and with very bright flares with dark-filament activity when the event is on the disk.

523.75 : 550.385.4 **433**

On the Great Solar Flare which Started at 21 h 09 m, February 9th, 1958, as the Likely Source of Geomagnetic Storm, February 11th.—K. Sinno. (*Rep. Ionosphere Res. Japan*, March 1958, Vol. 12, No. 1, pp. 6–9.) It is shown that there is a high probability that the flare caused the magnetic storm. Experimental evidence is given supporting a connection between the early part of a 200-Mc/s solar noise burst and a short-wave fade-out, and between the late part and magnetic-storm occurrence.

550.372 (47) **434**

Radio Wave Propagation and Soil Conductivity.—V. Kashprovskiĭ. (*Radio, Mosk.*, July 1958, No. 7, pp. 19–21.) Description of a scheme for mapping the soil conductivity throughout the Soviet Union by radio techniques.

550.38 **435**

The External Magnetic Field of the Earth.—A. Beiser. (*Nuovo Cim.*, 1st April 1958, Vol. 8, No. 1, pp. 160–162. In English.) The discrepancy between the equivalent geomagnetic dipole based on cosmic-ray observations and that derived from surface observations is investigated. See also 3721 of 1956 (Simpson et al.).

550.38 : 538.3 **436**

Reversals of the Earth's Magnetic Field.—D. W. Allan. (*Nature, Lond.*, 16th Aug. 1958, Vol. 182, No. 4633, pp. 469–470.) Calculations made by Rikitake (3074 of 1958) have been extended by the use of a digital computer.

550.389.2 : 523.165 : 629.132.1 **437**

Balloon Gear monitors Cosmic Radiation.—L. E. Peterson, R. L. Howard & J. E. Winckler. (*Electronics*, 7th Nov. 1958, Vol. 31, No. 45, pp. 76–79.) The balloon with a 60 lb load can be flown at 100 000 ft altitude for 22 hours. Equipment carried includes an omnidirectional Geiger counter, a spherical integrating ionization chamber and telemetry equipment.

550.389.2 : 551.510.535 **438**

Electron-Density Profiles in the Ionosphere during the I.G.Y.—R. L. Smith-Rose. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, p. 1874.) Note on a program organized by the Radio Research Station, Slough, England, of using electronic computers in the preparation of $N(h)$ profiles from ionograms obtained at four observatories.

550.389.2 : [629.19+629.136.3 **439**

Investigation of Upper Layers of the Atmosphere by means of Rockets and Artificial Earth Satellites.—E. K. Fedorov. (*Priroda, Mosk.*, Sept. 1957, No. 9, pp. 3–12.) Description of possible methods of investigation and the nature of the instrumentation required.

550.389.2 : 629.19 **440**

Scientific Investigations by means of Artificial Earth Satellites.—G. A. Skuridin & L. V. Kurnosova. (*Priroda, Mosk.*, Dec. 1957, No. 12, pp. 7–14.) A description of Sputnik II and the scientific equipment carried by it.

550.389.2 : 629.19 **441**

The Determination of the Trajectory of Artificial Satellites.—N. Carrara, P. F. Checcacci & L. Ronchi. (*Ricerca sci.*, July 1958, Vol. 28, No. 7, pp. 1341–1355.) Methods and the arrangement of ground equipment are described.

550.389.2 : 629.19 **442**

Exact Determination of the Velocity of an Artificial Satellite.—S. Khaĭkin. (*Radio, Mosk.*, Dec. 1957, No. 12, pp. 5–7.) Doppler measurements enable the velocity of the satellite to be measured to an accuracy within 1 part in 10^4.

550.389.2 : 629.19 **443**

Doppler Measurements on Soviet Satellites.—A. H. Allan & J. E. Drummond. (*N.Z. J. Sci.*, June 1958, Vol. 1, No. 2, pp. 143–153.) The first two Soviet satellites were successfully tracked by means of Doppler measurements alone. The apparatus and the method of analysis used are described.

550.389.2 : 629.19 **444**

Observations of Radio Signals from the First Man-Made Earth Satellite.—R. R. Long & G. H. Munro. (*Proc. Instn Radio Engrs, Aust.*, May 1958, Vol. 19, No. 5, pp. 201–206.)

550.389.2 : 629.19 **445**

Radio Observations on the First Russian Artificial Earth Satellite.—(*Trans. S. Afr. Inst. elect. Engrs*, Dec. 1957, Vol. 48, Part 12, pp. 363–369.) Doppler measurements made at Johannesburg at 40 Mc/s are reported. Orbit calculations are made neglecting ionospheric effects. See also 2087 of 1958 (Fejer).

550.389.2 : 629.19 **446**

Radio Observation of the Earth Satellite 1957α.—K. Miya, Y. Taguchi & S. Tabuchi. (*Rep. Ionosphere Res. Japan*, March 1958, Vol. 12, No. 1, pp. 16–27.) Describes observations of field strength, bearing and Doppler shift of the 20 005-kc/s signal. It is considered that to explain anomalous field strengths and Doppler shifts, propagation involving ground scattering followed by ionospheric reflection must be considered. Anomalous Doppler effects include rapidly varying shift, and apparent recession when the satellite is approaching the receiver.

550.389.2 : 629.19 **447**

Last Minutes of Satellite 1957β (*Sputnik II*).—D. G. King-Hele & D. M. C. Walker. (*Nature, Lond.*, 16th Aug. 1958, Vol. 182, No. 4633, pp. 426–427.)

550.389.2 : 629.19 : 523.165 **448**

Cosmic Rays Observed by Satellite 1958α.—Y. Aono & K. Kawakami. (*Rep. Ionosphere Res. Japan*, March 1958, Vol. 12, No. 1, pp. 28–36.) An analysis of the telemetered cosmic-ray information received in Japan. Except during magnetic-storm conditions the number of cosmic rays decreases exponentially with height. Diurnal and storm variations are discussed.

550.389.2 : 629.19 : 523.75 **449**

Effect of Solar Flares on Earth Satellite 1957β.—T. Nonweiler. (*Nature, Lond.*, 16th Aug. 1958, Vol. 182, No. 4633, pp. 468–469.) Fluctuations in the rate of decrease of the period of the satellite are apparently connected with variations in the total intensity of solar flares.

550.389.2 : 629.19 : 551.510.535 **450**

Faraday Fading of Earth-Satellite Signals.—F. B. Daniels & S. J. Bauer. (*Nature, Lond.*, 30th Aug. 1958, Vol. 182, No. 4635, p. 599.) A correction to the existing method of estimation of the integrated electron content of the ionosphere from earth-satellite signals is given.

551.510.5 : 621.396.96 : 621.396.11 **451**

Incoherent Scattering of Radio Waves by Free Electrons with Applications to Space Exploration by Radar.—W. E. Gordon. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1824–1829.) A powerful radar can detect the incoherent backscatter from free electrons in and above the earth's atmosphere and the received signal is spread in frequency by the Doppler shifts associated with the thermal motion of the electrons. Many practical applications are discussed including measurements of electron density, electron temperature, auroral ionization, and radar echoes from the sun, Venus and Mars.

551.510.535 **452**

Main Results of Meteorological Research done in Hungary during the Years 1954–1956.—B. Béll. (*Acta tech. Acad. Sci. hungarica*, 1957, Vol. 18, Nos. 1/2, pp. 133–160.) Work on the ionosphere carried out by the Central Meteorological Institute is noted.

551.510.535 **453**

Prevailing Wind in the Ionosphere and Geomagnetic S_q Variations.—S. Kato. (*J. Geomag. Geoelect.*, 1957, Vol. 9, No. 4, pp. 215–217.) It is shown theoretically that the prevailing ionospheric wind makes no contribution to the geomagnetic S_q current system despite the diurnal variation of ionospheric conductivity. See also 2406 of 1958.

551.510.535 **454**

A Study of 'Spread-F' Ionospheric Echoes at Night at Brisbane: Part 4—Range Spreading.—H. C. Webster. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 322–337.) From an examination of the variation of the amount of range spreading produced by sweeping the gain of a fixed-frequency ionospheric recorder, it is possible to gauge the degree of roughness of ionospheric layers. The effective roughness is a function of the separation of transmitter and receiver, being less the greater the distance between them. The intensity of Z-ray echoes recorded in Brisbane is consistent with Ellis's theory (2195 of 1956). Part 3: 121 of 1958 (Singleton).

551.510.535 : 550.385.4 : 621.396.11 **455**

On the Short-Wave Transmission Disturbance of 11th February, 1958.—Hakura & Takenoshita. (See 573.)

551.594.1 + 551.594.21 **456**

Measurement of the Size and Electrification of Droplets in Cumuliform Clouds.—B. B. Phillips & G. D. Kinzer. (*J. Met.*, Aug. 1958, Vol. 15, No. 4, pp. 369–374.) Charge distributions in clouds with fair-weather electric fields, at a mountain site in the United States, were Gaussian with symmetry about zero charge. Thundercloud droplets were highly electrified and in a given volume could be all negatively or all positively charged or a mixture of the two.

551.594.5 **457**

Auroral Echoes in the Ionograms Obtained in the Minauroral Region.—Y. Nakata. (*Rep. Ionosphere Res. Japan*, March 1958, Vol. 12, No. 1, pp. 1–5.) The observation of auroral echoes on a frequency-sweep ionosonde at Kokabunji in Japan is reported. Echoes were seen on three magnetically disturbed days, on one of which visual aurora was observed in Japan. The echo range corresponds to normal-incidence reflection from scattering centres at F-layer heights.

551.594.6 **458**

Correlation of Whistlers and Lightning Flashes by Direct and Visual Observation.—M. G. Morgan. (*Nature, Lond.*, 2nd Aug. 1958, Vol. 182, No. 4631, pp. 332–333.) Lightning was observed simultaneously with whistlers at Hanover, N.H., on 27th May 1957, but it is concluded that most lightning flashes do not generate whistlers.

551.594.6 **459**

Waveforms of Atmospherics.—B. A. P. Tantry & R. S. Srivastava. (*Proc. nat. Inst. Sci. India*, Part A, 26th May 1958, Vol. 24, No. 3, pp. 217–225.) A classification and interpretation of observed waveforms is given. See also 2416 of 1958 (Tantry et al.) and for a description of the equipment 3824 of 1958 (Tantry).

LOCATION AND AIDS TO NAVIGATION

621.396.933 **460**

The Flight Testing of Radio Facilities.—M. Cassidy. (*Proc. Instn Radio Engrs, Aust.*, June 1958, Vol. 19, No. 6, pp. 253–260.)

621.396.933.1 **461**

The Tacan Air Navigational System.—L. G. Thomas. (*Proc. Instn Radio Engrs, Aust.*, June 1958, Vol. 19, No. 6, pp. 247–252.) A general description is given of the main features of the system.

621.396.933.1 **462**

Air Trials of the Decca Navigator System.—H. Keeling. (*J. Inst. Nav.*, Oct. 1958, Vol. 11, No. 4, pp. 385–395.) A report is given of trials held in 1957 and 1958 to determine the operational suitability of the Mark 10 receiver and to compare its performance with that of the Mark 7.

621.396.933.1 **463**

The Evaluation and Use of the Dectra Navigation System.—E. W. Hare. (*J. Inst. Nav.*, Oct. 1958, Vol. 11, No. 4, pp. 377–384.) An interim report is given of field trials held by the British government since May 1957. The system appears to be capable of providing highly accurate position information over the North Atlantic Ocean.

621.396.933.23 **464**

'No Hands' Blind Landing.—(*Wireless World*, Dec. 1958, Vol. 64, No. 12, p. 579.) An automatic landing device, suitable for aircraft within 300 ft of the ground is described. Rate of fall is controlled by a radio altimeter. Centre-line guidance is provided by a system using the magnetic fields generated by two cables running parallel to the runway.

621.396.96 : 621.396.82 **465**

Radar Interference and its Reduction.—D. B. Brick & J. Galejs. (*Sylvania Technologist*, July 1958, Vol. 11, No. 3, pp. 96–108.) A survey of methods which can be used for the suppression of r.f. radar interference.

621.396.969.001.362 **466**

Marine Radar Simulation. — (*Brit. Commun. Electronics*, July 1958, Vol. 5, No. 7, pp. 508–509.) Block diagrams and brief descriptions are given of a navigation trainer and a radar simulator.

621.396.969.34 **467**

3-D Tactical Air-Position Radar in *H.M.S. Victorious*.—(*Brit. Commun. Electronics*, July 1958, Vol. 5, No. 7, pp. 510–511.) In addition to notes on special features of the system, an outline is given of a method for displaying information on the face of a c.r. tube using combinations of l.f. waveforms to produce the characters.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : [546.482.31 + 546.482.41 **468**

Some Photoelectric Properties of CdSe and CdTe Single Crystals.—S. V. Svechnikov & V. T. Aleksandrov. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 919–920.)

535.215 : 546.482.31 **469**

Special Features of the Photoconductive Properties of Cadmium Selenide.—S. V. Svechnikov. (*Zh. eksp. teor. Fiz.*, March 1958, Vol. 34, No. 3, pp. 548–554.) A two-stage excitation process is suggested for the explanation of observed anomalies in the photoconductivity of single crystals.

535.215 : 546.817.23 **470**

Photoconductivity in Lead Selenide.—D. H. Roberts. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 256–269.) Results are given of experiments with PbSe in the form of chemically deposited films, solid filaments and evaporated films to study the shape of the spectral response, the importance of potential barriers, the nature of the recombination mechanism and the role of oxygen in the sensitizing process.

535.243 : 546.482.21 **471**

Line Spectra of the Fundamental Absorption Edge of Cadmium Sulphide Crystals.—E. F. Gross, B. S. Razbirin & M. A. Yakobson. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 1149–1151.) An investigation of the spectral lines of CdS single crystals at 4·2°K in the range 4 853–4 889 Å. See also 3493 of 1957.

535.37 : 546.472.21 **472**

Notes on the Cathodoluminescence Efficiency of Zinc-Sulphide-Type Phosphors.—G. Gergely. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 270–272.)

535.37 : 546.472.21 **473**

Control of Luminescence by Charge Extraction.—P. J. Daniel, R. F. Schwarz, M. E. Lasser & L. W. Hershinger. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1240–1244.) The application of a potential of a few volts to phosphors of the ZnS group can quench fluorescence. This effect is investigated; it is a fundamental property of phosphors with large differences in hole and electron mobilities or capture cross-section. A simple mathematical theory is proposed to account for the observed effects.

535.37 : 546.482.21 **474**

Anisotropy of Edge Luminescence in Cadmium Sulphide.—D. Dutton. (*J. Phys. Chem. Solids*, July 1958, Vol. 6, No. 1, pp. 101–102.)

537.226/.228 : 546.431.824-31 **475**

The Internal Friction of Barium Titanate Ceramics.—T. Ikeda. (*J. phys. Soc. Japan*, Aug. 1958, Vol. 13, No. 8, pp. 809–818.) Heat dissipation in $BaTiO_3$ ceramics used in transducers is attributed to internal friction which is dependent on temperature, biasing field and vibration, but independent of frequency and porosity. The friction appears to originate as dielectric loss in the individual clamped domain crystals in the presence of piezoelectric coupling.

537.226/.227 : 546.431.824-31 **476**

Interpretation of Electron Paramagnetic Resonance in $BaTiO_3$.—A. W. Hornig, R. C. Rempel & H. E. Weaver. (*Phys. Rev. Lett.*, 15th Oct. 1958, Vol. 1, No. 8, pp. 284–286.) Experimental results differ considerably from those obtained by Low & Shaltiel (3490 of 1958). It is concluded that the resonance observed is due to an impurity of ferric ions at Ti sites.

537.226/.227 : 546.431.824-31 **477**

Electron Paramagnetic Resonance in $BaTiO_3$.—W. Low & D. Shaltiel. (*Phys. Rev. Lett.*, 15th Oct. 1958, Vol. 1, No. 8, p. 286.) Describes further investigations which have revealed a number of points in agreement with those reported by Hornig et al. (476 above), and a few in disagreement.

537.226/.227 : 546.431.824-31 **478**

Ferroelectric Switching Time of $BaTiO_3$ Crystals at High Voltages.—H. L. Stadler. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1485–1487.) Experimental results imply that ferroelectric switching in the voltage range 100–1 300 V does not involve the movement of elastic waves from one side of the crystal to the other.

537.226/.227 : 546.431.824-31 **479**

Contribution to the Theory of the Ferroelectric Properties of Polarized Barium Titanate Ceramics.—L. P. Kholodenko & M. Ya. Shirobokov. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 929–935.) The properties are examined for all crystal phases. Tensors for the dielectric constant and the piezoelectric moduli of polarized $BaTiO_3$ are calculated. See also 1787 of 1957 (Kholodenko).

537.226 **480**

Relaxation Polarization and Losses in Nonferroelectric Dielectrics Possessing Very High Dielectric Constants.—G. I. Skanavi, Ya. M. Ksendzov, V. A. Trigubenko & V. G. Prokhvatilov. (*Zh. eksp. teor. Fiz.*, Aug. 1957, Vol. 33, No. 2(8), pp. 320–334.) See also 801 of 1957 (Nomura).

537.226 : 546.431.824-31 **481**

Investigation of the Influence of Unilateral Compression on the Dielectric Permittivity of $BaTiO_3$ Ceramics in Strong Fields.—I. A. Izhak. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 953–961.) The permittivity decreases in the direction of the compression and increases in directions perpendicular to this. The variation of permittivity with compression also depends on the intensity of the electric field and temperature.

537.226 : 621.396.67 **482**

Anomalous Dispersion in Artificial Dielectrics.—A. F. Wickersham, Jr. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1537–1542.) The dependence of dispersion on array and scattering-element geometry has been investigated experimentally using planar arrays of thin metallic rectangles and varying the lengths and planar distributions of the rectangles. An attempt has been made to control dispersion by changing the array configuration. Applications are mentioned.

537.227 **483**

Stability of Ferroelectric Crystals.—V. Kh. Kozlovskiĭ. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1395–1397.)

537.227 : 547.476.3 **484**

Theory of the Ferroelectric Effect in Rochelle Salt.—T. Mitsui. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1259–1267.) A local-field theory of the clamped crystal is developed. The conditions for spontaneous polarization are investigated and the extent to which the theory can explain the properties of the clamped crystal is discussed.

537.311.1 **485**

Electrical Conduction in Solids.—(*Proc. roy. Soc. A*, 22nd July 1958, Vol. 246, No. 1244, pp. 1–31.)

Part 1—Influence of the Passage of Current on the Contact between Solids.—F. P. Bowden & J. B. P. Williamson (pp. 1–12).

Part 2—Theory of Temperature-Dependent Conductors.—J. A. Greenwood & J. B. P. Williamson (pp. 13–31).

537.311.31 + 537.311.33 **486**

High-Purity Metals and Semiconductors.—N. N. Murach. (*Priroda, Mosk.*, Dec. 1957, No. 12, pp. 21–26.)

537.311.31 : 537.323 : 539.23 **487**

Influence of Thickness on the Resistivity and Thermoelectric Power of Thin Films of Cobalt.—F. Savornin. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2866–2869.)

537.311.31 : 539.23 **488**

The Influence of a Layer of Selenium on the Electrical Conductivity of Very Thin Gold Films.—S. Minn & H. Damany. (*J. Phys. Radium*, June 1958, Vol. 19, No. 6, p. 612.) A note on the reduced surface resistivity of a gold film deposited on a thin film of Se. See also 2847 of 1957 (Minn & Offret).

537.311.33 **489**

Present and Future of Semiconductors.—A. F. Ioffe. (*Priroda, Mosk.*, Nov. 1957, No. 11, pp. 43–48.) A short survey of the development of semiconductors in the last 30 years is given and future applications are outlined.

537.311.33 **490**

Some Problems Concerning the Further Development of the Theory of Semiconductors.—A. F. Ioffe. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1153–1160.) An examination of the existing theory of semiconductors shows that revision is needed. Concepts applicable in the electron theory of metals are shown to be less useful in the study of semiconductors.

537.311.33 **491**

Metallic Contacts to Germanium and Silicon.—L. W. Davies & D. K. Milne. (*J. sci. Instrum.*, Nov. 1958, Vol. 35, No. 11, p. 423.) Details are given of the preparation of contacts which have high mechanical strength, good electrical properties and a readily controllable shape.

537.311.33 **492**

Effects of Electron-Electron Scattering on the Electrical Properties of Semiconductors.—R. W. Keyes. (*J. Phys. Chem. Solids*, July 1958, Vol. 6, No. 1, pp. 1–5.) The effects of electron-electron scattering, usually neglected in semiconductor theory, are investigated by solving the Boltzmann equation modified by the addition of an extra term. Results are given for the spherical and Ge band structures. In the latter case some effects are produced which are similar to those observed in the impurity-scattering region.

537.311.33 **493**

Generation-Recombination Noise in a Two-Level Impurity Semiconductor.—S. Teitler. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1585–1587.) "A general expression for the generation-recombination noise in a two-level impurity semiconductor is derived. Application is then made to zinc-doped germanium in the dark from 20°K to 100°K. The total white noise in this case exhibits a maximum and a minimum as the temperature is increased and the contributions to the noise which can be associated with the individual levels vary."

537.311.33 **494**

Investigation of the Temperature Dependence of the Work Function of some Semiconductors.—G. N. Lekhtinen, M. A. Rzaev & L. S. Stil'bans. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1221–1228.) Measurements on PbS, PbSe and PbTe show that at 150°C the work function varies differently for *n*-type and *p*-type semiconductors.

537.311.33 **495**

The Role of Surface Properties of Semiconductors in Adhesion Phenomena.—V. P. Smilga & B. V. Deryagin. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Oct. 1958, Vol. 122, No. 6, pp. 1049–1052.) Investigation of the development of adhesive forces in a metal/semiconductor contact on application of very-high-voltage electric fields.

537.311.33 **496**

Dependence of Emission Capacity of a *p-n* Junction upon its Structure and Condition of Operation.—K. B. Tolpygo. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 884–898.) Barrier-layer phenomena controlling the injection efficiency of a *p-n* junction are discussed and the influence of acceptors on the lifetime of minority carriers is considered. See also 472 of 1957.

537.311.33 : 061.3 (493) **497**

1958 Brussels Semiconductor Convention.—(*Brit. Commun. Electronics*, Aug. 1958, Vol. 5, No. 8, pp. 612–614.) A brief report is given of some of the papers read at the 'International Congress on Solid-State Physics and their Applications to Electronics and Telecommunications'.

537.311.33 : 535.215 **498**

A Note on Surface Recombination Velocity and Photoconductive Decays.—A. C. Sim. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 251–255.) The range of validity of correction formulae generally applied in photoconductive decay experiments for the measurement of the lifetime of excess carriers in semiconductors is shown to be restricted and a further correction is offered for the remaining range.

537.311.33 : 538.21 **499**

Magnetic Susceptibility of Semiconductors with an Impurity Zone in a Strong Magnetic Field.—M. I. Klinger. (*Zh. eksp. teor. Fiz.*, Aug. 1957, Vol. 33, No. 2(8), pp. 379–386.) See also 2803 of 1957 (Klinger et al.).

537.311.33 : 538.63 **500**

Method of Determination of Surface Recombination Velocity by Changing the Resistance of Semiconductors in a Magnetic Field.—V. P. Zhuze, G. E. Pikus & O. V. Sorokin. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1167–1173.) A description of a new experimental procedure: results are in good agreement with theory.

537.311.33 : 538.63 **501**

Theory of the Effect of a Magnetic Field on the Absorption Edge in Semiconductors.—R. J. Elliott, T. P. McLean & G. G. Macfarlane. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 553–565.) The electron energy bands in a solid can be split into sub-bands with a magnetic field. The absorption edge which arises from transitions between these bands shows a fine structure due to sub-band transitions. The shape of the transition structure is evaluated for spherical, spheroidal and degenerate bands. Often a series of peaks is formed from which the effective masses of the bands may be determined.

537.311.33 : 538.632 **502**

Hall and Holes.—(*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 601–605.) A simple explanation of the Hall effect and its applications is given.

537.311.33 : 538.632 : 621.317.3 **503**

Equipment for Hall-Effect Measurements in Semiconductors.—V. N. Bogomolov & V. A. Myasnikov. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1209–1214.) The equipment is particularly suitable for measurements on materials having low carrier mobility and having either very small or very large conductivity.

537.311.33 : 546.26-1 **504**

Rectification, Photoconductivity, and Photovoltaic Effect in Semiconducting Diamond.—M. D. Bell & W. J. Leivo. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1227–1231.) The potential barrier formed between a metal point and a *p*-type semiconducting diamond is due to the establishment of equilibrium between charges in surface and interior states. The semiconducting diamonds are photoconducting in the ultraviolet and visible regions.

537.311.33 : [546.28 + 546.289 **505**

Surface Mobility in Germanium and Silicon.—M. F. Millea & T. C. Hall. (*Phys. Rev. Lett.*, 15th Oct. 1958, Vol. 1, No. 8, pp. 276–278.) Experimental field effect data are presented suggesting that complete diffused surface scattering is incorrect, better agreement between experiment and theory being obtained by assuming partially diffused surface scattering.

537.311.33 : [546.28 + 546.289 **506**

Optical Properties of Semiconductors under Hydrostatic Pressure.—W. Paul & D. M. Warschauer. (*J. Phys. Chem. Solids*, 1958, Vol. 5, No. 1/2, pp. 89–106 & July 1958, Vol. 6, No. 1, pp. 6–15.)

Part 1—Germanium (pp. 89–101).

Part 2—Silicon (pp. 102–106).

Part 3—Germanium-Silicon Alloys (pp. 6–15).

537.311.33 : [546.28 + 546.289 **507**

Observation by Cyclotron Resonance of the Effect of Strain on Germanium and Silicon.—A. C. Rose-Innes. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 514–522.) Effects in microwave cyclotron resonance spectra at low temperatures are used to observe changes in the band structure of Ge and Si caused by non-isotropic elastic strain. Results are consistent with conclusions derived from piezoresistance measurements.

537.311.33 : 546.28 **508**

Valence-Band Structure of Silicon.—L. Huldt & T. Staflin. (*Phys. Rev. Lett.*, 1st Nov. 1958, Vol. 1, No. 9, pp. 313–315.) Using the technique of photogeneration of free carriers, an absorption spectrum, probably arising from theoretically predicted valence interband transitions, has been excited and observed in Si. See also 176 of January.

537.311.33 : 546.28 **509**

Fine Structure in the Absorption-Edge Spectrum of Si.—G. G. Macfarlane, T. P. McLean, J. E. Quarrington & V. Roberts. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1245–1254.) Measurements of the absorption spectrum of Si, made with high resolution near the main absorption edge at various temperatures between 4·2°K and 415°K, have revealed fine structure in the absorption on the long-wavelength side of this edge. This structure is analysed in detail and interpreted in terms of indirect transitions. See also 1463 of 1958.

537.311.33 : 546.28 **510**

The Temperature Variation of the Concentration of Impurity Carriers in Silicon.—E. H. Putley. (*Proc. phys. Soc.*, 1st Nov. 1958, Vol. 72, No. 467, pp. 917–920.) Discussion of this variation is frequently based on an expression which includes simplifying assumptions. A more general expression, which takes excited states of the impurity centre into account, is derived. It is shown that the results of analyses of carrier concentration data based on the simplified expression may be considerably modified when detailed knowledge of the various impurity centres becomes available.

537.311.33 : 546.28 **511**

Oxygen Impurity in Silicon Single Crystals.—A. Smakula & J. Kalnajs. (*J. Phys. Chem. Solids*, July 1958, Vol. 6, No. 1, pp. 46–50.)

537.311.33 : 546.28 **512**

Electron Spin-Lattice Relaxation in Phosphorus-Doped Silicon.—H. Honig & E. Stupp. (*Phys. Rev. Lett.*, 15th Oct. 1958, Vol. 1, No. 8, pp. 275–276.) The dependence of relaxation probability on magnetic field has been obtained under conditions of at least partial elimination of background photon flux, thereby isolating one of the phonon mechanisms involved in the relaxation process.

537.311.33 : 546.28 **513**

Diffusion of Gallium in Silicon.—A. D. Kurtz & C. L. Gravel. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1456–1459.) An open-tube vapour-solid diffusion technique at atmospheric pressure and temperatures between 1 130°C and 1 358°C gave lower diffusitivities and a higher activation energy than have been previously reported [3095 of 1956 (Fuller & Ditzenberger)]. Differences are discussed.

537.311.33 : 546.28 514

On the Delineation of *p-n* Junctions in Silicon.—P. A. Iles & P. J. Coppen. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, p. 1514.)

537.311.33 : 546.28 : 535.215 515

Phase-Shift Method of Carrier Lifetime Measurements in Semiconductors.—E. Harnik, A. Many & N. B. Grover. (*Rev. Sci. Instrum.*, Oct. 1958, Vol. 29, No. 10, pp. 889–891.) The phase difference between a sinusoidal modulation of carrier injection and the resulting modulation of the semiconductor conductance is measured by an *RC* compensating network. The conditions for direct proportionality between the phase difference and the effective lifetime are discussed. See also 3908 of 1957 (van der Pauw).

537.311.33 : 546.289 516

Recombination Centres in Germanium.—J. Okada. (*J. phys. Soc. Japan*, Aug. 1958, Vol. 13, No. 8, pp. 793–800.) The dependence of carrier lifetime in pure Ge upon injection level has been studied using a photoconductivity decay method. The results show that at least two recombination levels exist in pure Ge; one is active in *n*-type and the other in *p*-type Ge.

537.311.33 : 546.289 517

The Vibrational Spectrum and Specific Heat of Germanium.—F. A. Johnson & J. M. Lock. (*Proc. phys. Soc.*, 1st Nov. 1958, Vol. 72, No. 467, pp. 914–917.)

537.311.33 : 546.289 518

Experimental Determination of Electron Temperature in High Electric Fields Applied to Germanium.—E. G. S. Paige. (*Proc. phys. Soc.*, 1st Nov. 1958, Vol. 72, No. 467, pp. 921–923.) By observing the field dependence of drift velocity for an *n*-type Ge specimen in a strained and unstrained state, the electron temperature T_e can be deduced. For fields in the range 100–2 000 V/cm, values of T_e between 150° and 700°K were obtained with experimental errors not exceeding ±20 %.

537.311.33 : 546.289 519

Resistivities and Hole Mobilities in Very Heavily Doped Germanium.—F. A. Trumbore & A. A. Tartaglia. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, p. 1511.) Results are given of resistivity and Hall-effect measurements at 300°K on crystals of Ge containing up to 5×10^{20} acceptor atoms per cm³.

537.311.33 : 546.289 520

Diffusion and Electric State of Thermal Acceptors in Germanium.—V. A. Zhidkov & V. E. Lashkarev. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 877–883.) The temperature dependence of the diffusion coefficient of thermal acceptors is derived. See also 2796 of 1957.

537.311.33 : 546.289 521

Enhanced Cu Concentration in Ge containing Ni at 500°C.—A. G. Tweet & W. W. Tyler. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1578–1580.)

537.311.33 : 546.289 522

Evidence of Vacancy Clusters in Dislocation-Free Ge.—A. G. Tweet. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1520–1522.) Evidence of the existence of vacancy aggregates in Ge crystals is reported. The crystal etches much more rapidly when dislocations are absent over volumes of the order of cubic centimetres; this enhanced etching behaviour is eliminated by appropriate heat treatment.

537.311.33 : 546.289 : 535.376 523

Radiative Surface Effect in Germanium.—J. I. Pankove. (*J. Phys. Chem. Solids*, July 1958, Vol. 6, No. 1, pp. 100–101.) Radiation from the surface of a Ge crystal, into which holes were injected, was observed over a wide band with a peak at about 4·6 μ. It is attributed to an interband transition, involving the excitation of light holes in the strong electric field of the surface inversion layer.

537.311.33 : 546.289 : 537.32 524

Measurements of the Bulk Thermo-e.m.f. in Germanium.—P. I. Baranskiĭ & V. E. Lashkarev. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1161–1166.) An improved method of measuring the thermo-e.m.f. in *n*- and *p*-type Ge is described. Results obtained on polished or etched specimens are tabulated.

537.311.33 : 546.289 : 538.615 525

Zeeman Splitting of Donor States in Germanium.—R. R. Haering. (*Canad. J. Phys.*, Sept. 1958, Vol. 36, No. 9, pp. 1161–1167.) The linear Zeeman effect of the $2p\ m = \pm 1$ states of donor impurities is calculated using the approximation of effective mass theory of impurity states.

537.311.33 : 546.289 : 538.63 526

Resistivity and Hall Coefficient of Antimony-Doped Germanium at Low Temperatures.—H. Fritzsche. (*J. Phys. Chem. Solids*, July 1958, Vol. 6, No. 1, pp. 69–80.) The Hall coefficient *R* and resistivity ρ of Ge single crystals containing between 5×10^{14} and 10^{18} Sb atoms per cm³ were re-investigated at temperatures *T* between 1·3 and 300°K. The low-temperature anomalies—a steep maximum in the log *R* versus $1/T$ curves and a change of slope of the log ρ versus $1/T$ curves—are discussed on the basis of impurity conduction.

537.311.33 : 546.289 : 541.135 527

Germanium Electrode with a *p-n* Junction.—E. A. Efimov & I. G. Erusalimchik. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Oct. 1958, Vol. 122, No. 4, pp. 632–634.) A report of measurements made in a 0·1N solution of HCl using an electrode of *n*-type Ge 250 μ thick containing a *p-n* junction formed by diffusion of In.

537.311.33 : 546.3-1'289'28 528

Thermal Conductivity and Thermoelectric Power of Germanium-Silicon Alloys.—M. C. Steele & F. D. Rosi. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1517–1520.) Measurements on a series of alloys are reported and show that solid-solution alloying in the carrier concentration range where the carrier mobility is limited by impurity scattering, can significantly increase the figure of merit of thermoelectric materials.

537.311.33 : 546.681.19 529

Electron Mobilities in Gallium Arsenide.—L. R. Weisberg, J. R. Woolston & M. Glicksman. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1514–1515.)

537.311.33 : 546.682.86 530

Electrical Conductivity in *n*-Type InSb under Strong Electric Field.—Y. Kanai. (*J. phys. Soc. Japan*, Aug. 1958, Vol. 13, No. 8, pp. 967–968.) Conductivity was measured for fields sufficient to give deviations from Ohm's law. The deviation occurred at 2×10^2 V/cm and is considered to be caused by carrier ionization processes from the filled band.

537.311.33 : 546.873.241 531

Chemical Bonding in Bismuth Telluride.—J. R. Drabble & C. H. L. Goodman. (*J. Phys. Chem. Solids*, 1958, Vol. 5, Nos. 1/2, pp. 142–144.) The model proposed for Bi_2Te_3 disposes of some earlier difficulties; it explains some of the properties of Bi_2Te_3 and its alloys with Bi_2Se_3.

537.311.33 : 546.873.241 532

The Electrical Properties of Bismuth Telluride.—R. Mansfield & W. Williams. (*Proc. phys. Soc.*, 1st Nov. 1958, Vol. 72, No. 467, pp. 733–741.) Measurements were made of the electrical conductivity, Hall coefficient, thermoelectric power and Nernst coefficient on specimens cut from zone-melted Bi_2Te_3 and on a single crystal. The temperature range was 100°–600°K and specimens with a wide range of impurity content were examined.

537.311.33 : 546.873.241 533

The Optical Properties of Bismuth Telluride.—I. G. Austin. (*Proc. phys. Soc.*, 1st Oct. 1958, Vol. 72, No. 466, pp. 545–552.) The shape of the absorption edge is studied and is of the form expected for indirect transitions. The energy gap is found to be ≈0·13 eV at room temperature and the refractive index, determined from interference fringes, is 9·2 at 8–14 μ.

537.311.33 : 621.314.63 534

Reverse Breakdown in In-Ge Alloy Junctions.—D. R. Muss & R. F. Greene. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1534–1537.) Experiments show that in abrupt In-Ge alloy p^+n junctions breakdown occurs by the Zener mechanism in narrow junctions, by avalanche in broad junctions, and that both effects occur in intermediate-width junctions.

538.221 535

Instability of Bloch Walls due to Interstitial Atoms in a Ferromagnetic Material with Body-Centred Cubic Structure.—G. Biorci, A. Ferro & G. Montalenti. (*R. C. Accad. naz. Lincei*, May 1958, Vol. 24, No. 5, pp. 542–547.)

538.221 536

The Fluctuating-Field Ferromagnet at Low Temperatures.—F. D. Stacey. (*Aust. J. Phys.*, Sept. 1958, Vol. 11, No. 3, pp. 310–317.) The theoretical and experimental values of the constant in the $T^{3/2}$ law for Ni agree if an ordered state is produced by the mutual attraction of parallel elementary magnets each consisting of a coupled pair of spins.

538.221 **537**

The Antiferromagnetic Orientation of Magnetic Moments in the Alloy Ni_3Fe.—M. V. Dekhtyar. (*Zh. eksp. teor. Fiz.*, March 1958, Vol. 34, No. 3, pp. 772–773.) A description of measurements in the temperature range 0–600°C.

538.221 : 538.569.4 **538**

On the Thermodynamical Theory of Resonance and Relaxation Phenomena in Ferromagnetics.—G. V. Skrotskiĭ & V. T. Shmatov. (*Zh. eksp. teor. Fiz.*, March 1958, Vol. 34, No. 3, pp. 740–745.) The role of spin-lattice relaxation in the ferromagnetic resonance phenomenon is discussed. The equations obtained are compared with the Landau-Lifshitz and Bloch equations.

538.221 : 539.23 : 53.087.63 **539**

Magnetic Writing with an Electron Beam.—L. Mayer. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1454–1456.) Curie-point writing (*ibid.*, June 1958, Vol. 29, No. 6, p. 1003) permits local reversal of the direction of magnetization in suitable premagnetized magnetic films by using the dissipation energy of a focused electron beam to elevate the temperature temporarily above the Curie point. Well-defined traces of reversed magnetization which can be erased magnetically were recorded on MnBi films. Writing speeds corresponding to 3×10^4 bits/s and information densities corresponding to 10^5 bits/cm² were achieved. Electronic read-out of magnetically stored information is possible. See also 845 of 1958 (Williams et al.).

538.221 : 621.318.124 **540**

Cation Substitutions in $BaFe_{12}O_{19}$.—A. H. Mones & E. Banks. (*J. Phys. Chem. Solids*, 1958, Vol. 4, No. 3, pp. 217–222.) An experimental study of the variation in magnetic intensity of $BaFe_{12}O_{19}$ as a function of the substitution of ions such as Al^{III}, Ga^{III} Cr^{III} and Zn^{II} for Fe^{III}.

538.221 : 621.318.124 **541**

Investigation of the Substitution of Fe by Al, Ga and Cr in Barium Hexaferrite, $BaO.6Fe_2O_3$.—F. Bertaut, A. Deschamps & R. Pauthenet. (*C. R. Acad. Sci., Paris*, 5th May 1958, Vol. 246, No. 18, pp. 2594–2597.)

538.221 : 621.318.134 **542**

Magnetic Properties of $TiFe_2O_4$-Fe_3O_4 System and their Change with Oxidation.—S. Akimoto, T. Katsura & M. Yoshida. (*J. Geomag. Geoelect.*, 1957, Vol. 9, No. 4, pp. 165–178.)

538.221 : 621.318.134 **543**

Ferrimagnetic Resonance in $NiMnO_3$.—H. S. Jarrett & R. K. Waring. (*Phys. Rev.*, 1st Sept. 1958, Vol. 111, No. 5, pp. 1223–1226.) Ferrimagnetic resonance shows that the magnetic anisotropy is axial and the easy direction of magnetization lies in the basal plane. The magnetic anisotropy field equals $5 \cdot 2 \times 10^4$ G, corresponding to an anisotropy energy of $2 \cdot 6 \times 10^6$ ergs/cm³.

538.221 : 621.318.134 **544**

Grain Growth in Nickel Ferrites.—P. Levesque, L. Gerlach & J. E. Zneimer. (*J. Amer. ceram. Soc.*, 1st Aug. 1958, Vol. 41, No. 8, pp. 300–303.)

538.221 : 621.318.134 : 621.372.413 **545**

Resonant-Cavity Methods of Measuring Ferrite Properties.—R. A. Waldron. (*Brit. J. appl. Phys.*, Nov. 1958, Vol. 9, No. 11, pp. 439–442.) Formulae are given for the frequency shift on introducing a ferrite into a resonant cavity. Various sample shapes and positions are considered; it is concluded that a spherical shape is best, particularly because dielectric constant and permeability can be measured on it without change of sample, cavity, or mode. See also 2658 of 1956.

538.23 : 538.221 **546**

Coupling between Elementary Ferromagnetic Domains: Seesaw Effect.—L. Néel. (*C. R. Acad. Sci., Paris*, 28th May 1958, Vol. 246, No. 21, pp. 2963–2968.) Coupling between elementary domains, other than interacting ferromagnetic grains (203 of January) is considered, and it is shown that while the discrepancy between successive hysteresis cycles rapidly diminishes it does not necessarily vanish after the first alternation.

538.23 : 538.221 **547**

Creep of Asymmetric Hysteresis Cycles as a Function of the Amplitude of Asymmetry.—Nguyen Van Dang. (*C. R. Acad. Sci., Paris*, 28th May 1958, Vol. 246, No. 21, pp. 3034–3037.) The experiments described earlier (204 of January) were continued for fixed *n*. Maximum creep was found for cycles in which the maximum value of the field was about $1 \cdot 1$ times the coercive field.

MATHEMATICS

517.41 : 621.372 **548**

Functional Characteristics of a Node Determinant.—R. E. Bonner, L. H. Kosowsky & P. F. Ordung. (*J. Franklin Inst.*, May 1958, Vol. 265, No. 5, pp. 395–406.) A modification of the Laplace expansion is developed for use in network analysis, which removes initially all the negative terms.

517.7 **549**

The Numerical Evaluation of Expressions involving Complete Elliptic Integrals.—F. W. Grover. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 496–502.)

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) : 529.786 **550**

Primary Frequency Standard using Resonant Caesium.—W. A. Mainberger. (*Electronics*, 7th Nov. 1958, Vol. 31, No. 45, pp. 80–85.) Description of the 'atomichron' equipment. See also 212 of January (Essen et al.).

621.317.088.6 **551**

A Method of Correcting for the Response Time Delays of Measuring Equipment.—J. A. Sirs. (*J. sci. Instrum.*, Nov. 1958, Vol. 35, No. 11, pp. 419–422.) The error due to the delay is obtained by considering the output response to a unit step input impulse and applying Laplace transform analysis. Correction formulae are derived and their application illustrated.

621.317.2 : 621.373.42 **552**

Low-Frequency Sine-Wave Generators.—(*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 459–467.) A review of modern commercial-type l.f. oscillators with details of some of their circuitry.

621.317.3 : 621.316.722.078.3 **553**

Rapid Testing of Electronic Direct-Voltage Stabilizers.—Perrier & d'Ast. (See 608.)

621.317.33 **554**

A Novel, High-Accuracy Circuit for the Measurement of Impedance in the A.F., R.F. and V.H.F. Ranges.—D. Karo. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 505–510.) The circuit consists of two branches, one of which contains the unknown impedance. These branches are fed in phase opposition from the secondaries of two mutual inductors or two transformers. Between 100 c/s and 50 Mc/s the error limit varies, according to experimental conditions, from $\pm 0 \cdot 001$ % to $\pm 0 \cdot 01$ %.

621.317.332 : 539.23 **555**

Measurement of Very Slight Variations of Resistance. Applications to the Magnetoresistance of Thin Films.—A. Colombani, P. Huet & C. Vautier. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2869–2872.) Description of a differential method of measurement with sensitivity $\Delta R/R$ of 10^{-6}, in which a compensating l.f. voltage of opposite phase is derived by means of a resistance in series with the sample.

621.317.382 : 538.632 : 537.311.33 **556**

Use of the Hall Effect in Semiconductors for Electric Power Measurements.—L. S. Berman. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1192–1196.) Two circuits for power measurements in the frequency range 400–500 kc/s have been investigated: one using *n*-type Ge and the other InSb. Results show that performance is linear and independent of frequency.

621.317.39 **557**

The Differential Transformer as a Sensitive Measuring Device.—J. H. Heath. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 630–633.) A differential transducer is used as a sensing head with the two secondary windings connected in series addition. The linear range may be subdivided into a series of sensitive sections.

621.317.616 : 621.373.4 **558**

Broad-Band Generator has Wide and Narrow Sweeps.—C. C. Cooley, Jr. (*Electronics*, 7th Nov. 1958, Vol. 31, No. 45, pp. 88–91.) The frequency-sweep generator described covers sweep widths from 100 kc/s to 300 Mc/s in the centre-frequency range 200 kc/s–1 000 Mc/s.

621.317.7 : 621.387 : 621.396.822.029.63 **559**

Application of Gas-Discharge Tubes as Noise Sources in the 1 700–2 300 Mc/s Band.—M. Kollanyi. (*J. Brit. Instn Radio Engrs*, Sept. 1958, Vol. 18, No. 9, pp. 541–548.) The design considerations and the performance of a gas-discharge helix-coupled noise source are given. Noise-figure measurements can be made with an accuracy of 0·2 dB.

621.317.725.027.3 **560**

Compensation Electron-Beam High-Voltage Voltmeter.—G. I. Shal'nikov. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1371–1378.) A new instrument for the accurate measurement of direct voltages up to 30 000 and theoretically applicable to alternating voltages at frequencies up to 5 Mc/s, and also to short pulses.

621.317.733 **561**

A Precision, Guarded Resistance Measuring Facility.—F. H. Wyeth, J. B. Higley & W. H. Shirk, Jr. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 471–475. Discussion, pp. 475–476.)

621.317.737 **562**

A Simple 3-cm *Q*-Meter.—A. E. Barrington & J. R. Rees. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 511–512.) A simple reflectometer method for measuring *Q*-factors up to about 4 000, with an error limit of approximately 10 %.

621.317.763.029.6 : 535.417 **563**

The Optical Approach in Microwave Measurement Technique.—J. I. Caicoya. (*Brit. Commun. Electronics*, July 1958, Vol. 5, No. 7, pp. 500–507.) A survey is made of interferometer and grating - spectrometer techniques which can be applied to microwave measurements.

621.317.794.029.6 : 621.316.825 **564**

Experimental Wide-Band Thermistor Mounts.—J. Swift. (*Proc. Instn Radio Engrs, Aust.*, June 1958, Vol. 19, No. 6, pp. 261–264.) Several simple mounts are described which consist of a coaxial line terminated by two thermistors placed across an untuned cavity. One type covers the band 450–5 000 Mc/s with a maximum voltage s.w.r. of 1·3.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.787 : 621.372.413 **565**

Microwave Manometer.—A. G. Kramer & P. M. Platzman. (*Rev. sci. Instrum.*, Oct. 1958, Vol. 29, No. 10, pp. 897–898.) A differential pressure indicator using a cavity resonator at 8 650 Mc/s is described. The sensitivity was 2·4 Mc/s per mm Hg pressure difference.

535.376 : 621.397.62 **566**

Problems in Electroluminescent Television Display.—R. M. Bowie. (*Sylvania Technologist*, July 1958, Vol. 11, No. 3, pp. 82–85.) Technical and economic problems which have to be overcome before an electroluminescent device can compete with a c.r. tube are outlined with particular reference to the Sylvatron [see 244 of January (Butler & Koury)].

621.385.833 **567**

Image of a Surface obtained with Negative Ions.—R. Bernard & R. Goutte. (*C. R. Acad. Sci., Paris*, 5th May 1958, Vol. 246, No. 18, pp. 2597–2599.)

654.171 : 535.376 **568**

Transfluxor-Controlled Electroluminescent Display Panels.—J. A. Rajchman, G. R. Briggs & A. W. Lo. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1808–1824.) A detailed description of a display system using electroluminescence magnetically controlled by an electrical input signal. The 1 200 elements of the array are arranged in 30 rows and are each associated with a transfluxor [3509 of 1955 (Rajchman & Lo)]. Advantages and disadvantages of the system are discussed.

PROPAGATION OF WAVES

621.396.11 : 551.510.5 : 621.396.96 **569**

Incoherent Scattering of Radio Waves by Free Electrons with Applications to Space Exploration by Radar.—Gordon. (See 451.)

621.396.11 : 551.510.52 **570**

Some Generalized Scattering Relationships in Transhorizon Propagation.—A. T. Waterman, Jr. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1842–1848.) A discussion of the consequences which follow from the assumption that the physical mechanism is a single-scattering process distributed systematically throughout the atmosphere. Expressions are derived for the variation of received power with distance, for various scattering angles and beam widths.

621.396.11 : 551.510.52 **571**

Geometric Characteristics of the Scattering of Radio Waves at Turbulent Inhomogeneities of the Troposphere.—D. M. Vysokovskiĭ. (*Elektrosvyaz'*, Sept. 1957, No. 9, pp. 12–39.) Exact and approximate formulae are derived for determining the dimensions of the scattering region and the angle of scattering. An expression for the scattered power is given in the form of an integral over the scattering region. On the basis of an investigation of the extremum of this integral, the dimensions of the effective scattering region are determined for the case of wide polar diagrams, and the choice of aerials for communication based on scatter propagation is discussed. The main geometric characteristics of the scattering region are given for the case of narrow polar diagrams.

621.396.11 : 551.510.535 **572**

The Magnetoionic Theory and its Results.—D. Lépéchinsky. (*Ann. Télécommun.*, Feb. & March 1957, Vol. 12, Nos. 2 & 3, pp. 60–70 & 74–91.) A practical method of calculating propagation parameters is derived from the general Appleton-Hartree equation. Propagation in the Q.L., Q.T., and limiting Q.L.-Q.T. regions is examined and applications of the method are considered.

621.396.11 : 551.510.535 : 550.385.4 **573**

On the Short-Wave Transmission Disturbance of 11th February, 1958.—Y. Hakura & Y. Takenoshita. (*Rep. Ionosphere Res. Japan*, March 1958, Vol. 12, No. 1, pp. 10–15.) Reports observations in Japan of signal strength and fading rates on three h.f. circuits during an ionospheric storm. Flutter-fading began on a trans-polar route at the time of the sudden commencement, and moved south to the lower-latitude paths during the course of the storm. High night-time field strengths and fading rates were observed at a time when auroral echoes were detected by ionospheric soundings.

621.396.11.029.6 **574**

Sporadic-E Skip on 200 Mc/s?—R. B. Cooper, Jr. (*QST*, Nov. 1958, Vol. 42, No. 11, pp. 33–35 . . 162.) A summary is given of reports of long-range reception of television signals on frequencies between 60 and 204 Mc/s.

621.396.11.029.62 **575**

Role of Turbulent Scattering in Long-Distance Radio Propagation at Metre Wavelengths.—F. A. Kitchen & M. A. Johnson. (*Nature, Lond.*, 2nd Aug. 1958, Vol. 182, No. 4631, pp. 302–304.) Field-strength measurements were made in November and December 1957 of propagation at 203·5 Mc/s over sea in the English Channel area for distances up to 350 miles. Results support the theory that turbulence and scattering are almost always present at all levels in the troposphere.

621.396.11.029.62 : 523.5 **576**

The Forward-Scattered Radio Signal from an Overdense Meteor Trail.—P. A. Forsyth. (*Canad. J. Phys.*, Aug. 1958, Vol. 36, No. 8, pp. 1112–1124.) A recently presented expression [883 of 1958 (Hines & Forsyth)] for the forward-scattered signal from an overdense meteor trail was tested in a particular observed meteor trail. The electron line density is calculated by three different methods, of which two are based on the new expression. The resulting agreement is within the experimental error.

621.396.11.029.62 : 621.397.81 **577**

More on the 'Plymouth Effect'.—J. P. Grant. (*Wireless World*, Dec. 1958, Vol. 64, No. 12, pp. 587–590.) Back-scatter from the sea and reflection from aerial arrays on Guernsey are suggested as possible causes of anomalous television reception at Plymouth of the B.B.C. Devon television transmitter. See also 3612 of 1958 (Sofaer).

621.396.11.029.64 **578**

Influence of the Semi-permanent Low-Level Ocean Duct on Centimetre-Wave Scatter Propagation Beyond the Horizon.—F. A. Kitchen, W. R. R. Joy & E. G. Richards. (*Nature, Lond.*, 9th Aug. 1958, Vol. 182, No. 4632, pp. 385–386.) Experiments made over sea in the English Channel area in 1957 and 1958 for various heights of transmitter and receiver show that when the site of the receiving aerial is relatively high, e.g. 400 ft, the surface-guided component of the signal beyond the horizon is not directly observable.

621.396.11.029.64 **579**

Statistical Data for Microwave Propagation Measurements on Two Oversea Paths in Denmark.—P. Gudmandsen & B. F. Larsen. (*Acta polyt., Stockholm*, 1957, No. 213, 37 pp.) Measurements were made using vertically spaced aerials for wavelengths of 17 and 6·4 cm over two E-W paths of lengths 54 km and 82 km. Fading conditions have been studied using diversity systems and single receivers.

RECEPTION

621.376.232.2 **580**

Design of Detector Stages for Signals with Symmetrical or Asymmetrical Sidebands.—A. van Weel. (*J. Brit. Instn Radio Engrs*, Sept. 1958, Vol. 18, No. 9, pp. 525–538. Correction, *ibid.*, Oct. 1958, Vol. 18, No. 10, p. 581.) See also 249 of January.

621.376.33 : 621.396.82 **581**

Alternative Detection of Co-channel F.M. Signals.—H. W. Farris. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1876–1877.) A weaker signal can be separated by correlating the sum of weaker and stronger signals with the stronger signal at the i.f. of the receiver.

621.396.62.029.62 **582**

Further Notes on the ARR 3 Sonobuoy Receiver.—(*Wireless World*, Dec. 1958, Vol. 64, No. 12, p. 590.) Additional precautions against the possibility of radiation at television channel-1 frequencies. See also 254 of January (Taylor).

621.396.81 **583**

Simultaneous Variation of Amplitude and Phase of Gaussian Noise, with Applications to Ionospheric Forward-Scatter Signals.—T. Hagfors & B. Landmark. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 555–559.) The scatter signal is shown to possess amplitude and phase characteristics similar to those of Gaussian noise. Spaced-aerial observations indicate that the angular spectrum of received waves is randomly phased.

621.396.81 : 621.396.65 **584**

The Analysis of Field Strength Records for Radio Link Assessment.—M. W. Gough. (*Point to Point Telecommun.*, June 1958, Vol. 2, No. 3, pp. 28–47.) Recording and analytical techniques are outlined and propagation effects represented in chart form are discussed.

621.396.812.3 **585**

Fading of Radio Waves.—P. Venkateswarlu & R. Satyanarayana. (*Curr. Sci.*, Aug. 1958, Vol. 27, No. 8, p. 296.) Theoretical amplitude distribution curves for medium frequencies agree closely with experimental curves for distances of 110 and 320 km but not for 1 700 km, where the experimental curve shows two maxima.

621.396.812.3.029.6 : 621.396.65 **586**

On the Fading of Ultra Short Waves in Radio Links.—V. N. Troitskiĭ. (*Elektrosvyaz'*, Oct. 1957, No. 10, pp. 32–39.) An analysis is given of the possible types of fading in radio links. The use of an effective gradient of permittivity in calculations of field intensity is discussed and experimental values for central U.S.S.R. are given. The effect of horizontal inhomogeneities is considered in an appendix.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2 **587**

Single-Sideband Modulation.—B. Rassadin. (*Radio, Mosk.*, June 1958, No. 6, pp. 25–27.) Description of a system operating in the range 7·0–7·1 Mc/s.

621.376.2 **588**

Phase - Compensation Methods of Shaping a Single-Sideband Signal.—A. Semenov & V. Verzunov. (*Radio, Mosk.*, June 1958, No. 6, pp. 27–29.)

621.376.2 : 621.396.41 **589**

Suppression of the Unwanted Sideband in Single-Band Multiphase Radio Systems.—I. V. Lobanov. (*Elektrosvyaz'*, Sept. 1957, No. 9, pp. 3–11.) Formulae are derived for determining the degree of suppression of the sideband in three- and four-phase systems, depending on the magnitude of amplitude and phase errors of voltages feeding the system. From these formulae graphs are plotted showing the possibility of realizing these systems under various specific conditions.

621.376.23 : 621.396.41 **590**

Step Detection.—A. R. Billings. (*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 453–455.) The attenuation distortion produced by step detection is small, and this method when applied to time-division multiplex systems considerably reduces adjacent-channel crosstalk.

621.391 **591**

Binary Symmetric Decision Feedback Systems.—B. Harris & K. C. Morgan. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 436–443.) Schemes are considered in which the decision to accept or reject a symbol is based on word groups as well as on a digit-by-digit basis. Both the information rate and error probability are improved and general expressions are given from which they may be calculated.

621.391 **592**

Channels with Side Information at the Transmitter.—C. E. Shannon. (*IBM J. Res. Develop.*, Oct. 1958, Vol. 2, No. 4, pp. 289–293.) In communication systems where information is to be transmitted from one point to another, additional side information is available at the transmitting point, which relates to the state of the transmission channel and can be used to aid in the coding and transmission of information. A type of channel with side information is studied and its capacity determined.

621.394.14 **593**

Relative Speeds of Telegraphic Codes.—D. A. Bell & T. C. Duggan. (*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 476–480.) A comparison of code speeds taking account of the frequency of occurrence of different letters shows that, for English, the advantage of a statistically weighted code would not be commensurate with the complexity of the decoding apparatus required.

621.395.5 : 621.314.7 **594**

Potential Uses for Transistors in Line Communications.—J. R. Tillman. (*Brit. Commun. Electronics*, Aug. 1958, Vol. 5, No. 8, pp. 594–600.) The existing systems are reviewed and advantages and disadvantages of the replacement of valves by transistors are considered.

621.396.2 : 551.510.52 **595**

Tropospheric Scatter System Evaluation.—M. Telford. (*J. Brit. Instn Radio Engrs*, Sept. 1958, Vol. 18, No. 9, pp. 511–523.) A chart is presented to enable performance and/or equipment parameters to be determined for a wide range of conditions. Particular reference is made to the requirements of f.m. multichannel telephony systems. The economics, present engineering limitations, and possible future trends in such systems are discussed.

621.396.2 : 621.394.3 **596**

A Communication Technique for Multipath Channels.—G. D. Hulst. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, p. 1882.) Note on 1873 of 1958 (Price & Green).

621.396.3 : 621.396.43 : 523.5 **597**

On the Choice of Frequencies for Meteor-Burst Communication.—M. L. Meeks & J. C. James. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1871.)

621.396.4 : 621.376.4 **598**

Radio-Frequency Powers and Noise Levels in Multichannel Radiotelephone Systems using Angular Modulation.—J. D. Thomson. (*Proc. Instn Radio Engrs, Aust.*, May 1958, Vol. 19, No. 5, pp. 211–220. Discussion.) Formulae and curves are derived for the calculation of the required receiver input to ensure a specified noise standard. The method is applied to the design of a five-channel phase-modulated system employing 16 repeater sections.

621.396.41 : 621.376.5 **599**
Methods for Investigating Transients in Phase-Correcting Systems when Receiving Code Combinations of Telegraph Pulses. — L. N. Shchelovanov. (*Elektrosvyaz'*, Sept. 1957, No. 9, pp. 42–49.) Methods applicable to open and closed circuits with a variable sequence period of pulses are discussed. The process of regulation in a system for correcting the phase of the tuning fork in a multiplex telegraph apparatus for p.c.m. is examined.

621.396.65 **600**
Factors affecting the Use of Over-the-Horizon Links in Telecommunication Networks.—C. A. Parry. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 485–496.) The use of multichannel scatter links for national communications is considered. The overall system is considered including strategic, environmental and commercial aspects.

SUBSIDIARY APPARATUS

621.311.62 : 621.314.7 **601**
Transistor Stabilized Power Supply for 5–9 V, 800 mA.—H. Hahn & M. Sauzade. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2875–2878.) Circuit details of a unit with output resistance $0 \cdot 003\ \Omega$, and output voltage variation $3 \cdot 6$ mV or less for 15 % change in input voltage.

621.311.62 : 621.314.7 : 621.397.6 **602**
A Transistor Regulated Power Supply for Video Circuits.—R. H. Packard & M. G. Schorr. (*Trans. Inst. Radio Engrs*, Dec. 1957, No. PGBTS-9, pp. 32–38. Abstract, *Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, p. 672.)

621.311.69 : 621.314.63 : 533.215 **603**
Solar Battery.—V. Shchekin. (*Radio, Mosk.*, Aug. 1958, No. 8, pp. 29–30.) The battery consists of silicon plates about 1 mm thick covered by thin boron films, forming *p-n* junction photo-elements. The efficiency of the battery is approximately 12 % and a possible improvement up to 22 % is indicated.

621.314.63 : 546.28 **604**
Some Basic Physical Properties of Silicon and How they Relate to Rectifier Design and Application.—G. Finn & R. Parsons. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CP-3, No. 3, pp. 110–113. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 573.)

621.316.721/.722 **605**
A Constant-Voltage/Constant-Current Stabilizer.—D. P. C. Thackeray. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 646–647.)

621.316.721 : 621.314.6 **606**
Current-Balancing Reactors for Semiconductor Rectifiers.—I. K. Dortort. (*Commun. & Electronics*, Sept. 1958, No. 38, pp. 452–456. Discussion.) In both semiconductor and mercury arc rectifiers of high current capacity where many diodes are connected in parallel, the currents in the separate units must be balanced. Balancing arrangements for semiconductor rectifiers are described, including the use of punched laminations as strip-type reactors.

621.316.722.078.3 **607**
The Analysis and Design of Constant-Voltage Regulators.—I. B. Friedman. (*Trans. Inst. Radio Engrs*, March 1956, Vol. CP-3, No. 1, pp. 11–14. Abstract, *Proc. Inst. Radio Engrs*, June 1956, Vol. 44, No. 6, Part 1, pp. 831–832.)

621.316.722.078.3 : 621.317.3 **608**
Rapid Testing of Electronic Direct-Voltage Stabilizers.—F. Perrier & L. d'Ast. (*C. R. Acad. Sci., Paris*, 19th May 1958, Vol. 246, No. 20, pp. 2878–2880.) Routine tests of voltage stabilizers at an electron-optics laboratory in Toulouse are described.

621.316.79 : 537.311.33 : 537.32 **609**
Semiconductor Thermostat for Self-Oscillators.—E. K. Iordanishvili & L. G. Tkalich. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1215–1220.) The thermostat provides control for an ambient temperature range of $-60°$ to $+60°$C.

621.318.56 **610**
How to Improve Relay Reliability.—L. B. Kleiger. (*Electronic Equipm. Engng*, April 1958, Vol. 6, No. 4, pp. 37–40.) Practical advice is given on the choice and use of electromechanical relays.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 : 621.396.82 **611**
Fluctuating Interference in the Trunk Television Channel of a Coaxial Cable.—A. K. Oksman. (*Elektrosvyaz'*, Oct. 1957, No. 10, pp. 3–10.) Typical spectral distributions of fluctuating interference are considered and the corresponding requirements with respect to the signal/interference ratio are discussed. Results are given of an experimental investigation.

621.397.5 : 535.623 **612**
Electronic Composites in Modern Television.—R. C. Kennedy & F. J. Gaskins. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1798–1807.) A review of various electronic techniques used in television to simulate optical effects used in motion-picture photography is followed by a description of a new process called 'chroma-key'. This utilizes a highly saturated colour background for the inset subject and has some advantages compared with a monochrome inset.

621.397.6.001.4 : 535.623 **613**
Video Transmission Testing Techniques for Monochrome and Colour.—J. R. Popkin-Clurman. (*Trans. Inst. Radio Engrs*, June 1957, No. PGBTS-8, pp. 14–24.) A description of the window-signal, the multifrequency-burst, the modulated-stair-step-signal and the sine-squared-wave methods of testing is given together with a list of possible defects in television transmission.

621.397.61 : 535.623 **614**
The Correction of Differential Phase Distortion in Colour Television Transmitters.—V. J. Cooper. (*Trans. Inst. Radio Engrs*, June 1957, No. PGBTS-8, pp. 1–5.) Two distinct methods of correcting differential phase distortion without affecting the amplitude linearity characteristic, and two types of test are explained.

621.397.611.2 **615**
Reduction of Image Retention in Image-Orthicon Cameras.—S. L. Bendell & K. Sadashige. (*Trans. Inst. Radio Engrs*, Dec. 1957, No. PGBTS-9, pp. 52–58. Abstract, *Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, p. 672.)

621.397.611.2 **616**
Recent Developments in TV Camera Tubes.—F. S. Veith. (*Trans Inst. Radio Engrs*, Dec. 1957, No. PGBTS-9, pp. 21–31. Abstract, *Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, p. 672.)

621.397.62 **617**
A Television Receiver Circuit for the 625-line C.C.I.R. Standard.—(*Mullard tech. Commun.*, Sept. 1958, Vol. 4, No. 33, pp. 46–92; Correction, *ibid.*, Nov. 1958, Vol. 4, No. 35, p. 62.) A group of papers including a note on the C.C.I.R. specifications and giving a detailed description of the design and construction of a 19-valve experimental receiver Type CNU 10. Overall sensitivity on C.C.I.R. channel 4 is 10 μV for 1 V at the video detector in a 3 dB bandwidth of $4 \cdot 5$ Mc/s at the i.f. of $38 \cdot 9$ Mc/s. Intercarrier f.m. sound is used with an i.f. of $33 \cdot 4$ Mc/s, limiter, ratio detector and 2-stage audio with 50 μs de-emphasis and feedback. The d.c. component is fully maintained through the video amplifier, with a.g.c. operating as a black-level clamp suitably noise-cancelled and delayed. Double-clipping is used in the sync-separator with integration and clipping for the frame pulse and a flywheel circuit for the line timebase. Particular attention is paid to linearity and freedom from ringing in the latter. The 90° picture tube operates at 16 kV.

621.397.62 : 535.376 **618**
Problems in Electroluminescent Television Display.—Bowie. (See 566.)

621.397.62 : 535.623 : 621.385.832 **619**
A New Cathode-Ray Tube for Monochrome and Colour Television.—D. Gabor, P. R. Stuart & P. G. Kalman. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 581–606. Discussion, pp. 604–606.) A flat c.r. tube is described (see 588 of 1957) and details are given of its novel features which have been tested singly and partly in combination. These include a reversing lens which rotates the plane formed by a fan of rays through 180°

and increases the angle of divergence by a factor of 4. Methods of manufacture are suggested and details of the electron-optical calculations are given.

621.397.62 : 535.88 **620**

The Eidophor System is Successful.—E. Gretener. (*Elektron, Linz*, 1958, No. 9, pp. 222–226.) The operating principles of this method of television projection are described with details of a recently developed projector. See also 2350 of 1952 (Baumann) and back references, in particular 296 of 1948 (Thiemann).

621.397.621 : 535.623 **621**

Novel Colour-Television Display System.—R. W. Wells. (*Brit. Commun. Electronics*, July 1958, Vol. 5, No. 7, pp. 520–522.) The experimental device described uses a projection tube in conjunction with a Faraday cell controlled by colour switching waveforms, and a fixed composite 'cellophane' filter, the layers of which have their molecular orientation offset by 6°. Other types of cells are considered and sequential and simultaneous display systems using this principle are outlined.

621.397.7 **622**

The Maintenance of Television Studio Equipment.—V. G. Perry. (*Brit. Commun. Electronics*, Aug. 1958, Vol. 5, No. 8, pp. 586–591.)

TRANSMISSION

621.396.61 **623**

Combined Operation of Broadcast Transmitters.—W. N. Black. (*A.W.A. tech. Rev.*, 1958, Vol. 10, No. 3, pp. 110–139.) A description is given of the complete system for combining, with a bridged-T network, the outputs of two 10 kW transmitters.

621.396.61 : 621.375.2 **624**

Amplitude-Modulated Transmitter Class-C Output Stage.—C. G. Mayo & H. Page. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 523–531.) The output stage in which the load impedance varies over the working frequency band is discussed in detail. An anode impedance which is symmetrical with respect to the carrier frequency permits the radiation of an undistorted output envelope.

VALVES AND THERMIONICS

621.314.63 **625**

Measurement of Voltage/Current Characteristics of Junction Diodes at High Forward Bias.—A. K. Jonscher. (*J. Electronics Control*, Sept. 1958, Vol. 5, No. 3, pp. 226–244.) The theoretical voltage/current relation $I^{\frac{1}{2}} = S(V - V_0)$ obtained previously (4003 of 1958) is confirmed experimentally for a wide range of planar *p-n* diode structures up to current densities $> 10^3$ A/cm².

621.314.63 : 537.311.33 **626**

Diode Hole Storage and 'Turn-On' and 'Turn-Off' Time.—G. Grimsdell. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 645–646.) The conditions of measurement must be considered in each case when comparing semiconductor diodes by their published hole-storage times.

621.314.63 : 546.289 **627**

The Temperature Dependence of Noise Temperature Ratio in Germanium Diodes.—A. Hendry. (*Brit. J. appl. Phys.*, Nov. 1958, Vol. 9, No. 11, pp. 458–460.) The 30 Mc/s noise temperature ratio of a d.c. biased Ge mixer diode is observed to increase as its temperature is lowered, indicating the presence of noise which is in excess of thermal and shot noise and increases as the temperature is lowered.

621.314.7 **628**

On the Origin of the Fluctuation of Crystal Triode Parameters.—(*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1197–1208.)

Part 1—*P-N-P*-Type Triodes.—A. P. Vyatkin.

Part 2—*N-P-N*-Type Triodes.—A. P. Vyatkin & V. A. El'chin.

Investigation of the influence of temperature, fusion time and impurity concentration on the depth of penetration into Ge.

621.314.7 **629**

Present-Day Limits of Transistor Characteristics.—E. R. Hauri. (*Bull. schweiz. elektrotech. Ver.*, 16th Aug. 1958, Vol. 49, No. 17, pp. 809–810..833.) A review with over 30 references.

621.314.7 **630**

The Effect of a Magnetic Field on Point-Contact Transistors.—K. K. Bose. (*Electronic Engng*, Nov. 1958, Vol. 30, No. 369, pp. 639–641.) Experiments to determine the changes in frequency response, output and amplification characteristics are described. All the changes can be attributed to the disturbance of the flow of injected carriers by the magnetic field.

621.314.7 **631**

The Current Amplification of a Junction Transistor as a Function of Emitter Current and Junction Temperature.—W. W. Gärtner, R. Hanel, R. Stampfl & F. Caruso. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1875–1876.) An approximate expression is derived, on the basis of existing theories, for obtaining α as a function of emitter current.

621.314.7 **632**

Effective Collector Capacitance in Transistors.—R. Zuleeg. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1878–1879.)

621.314.7 **633**

A Method of Studying Surface Barrier Height Changes on Transistors.—J. R. A. Beale, D. E. Thomas & T. B. Watkins. (*Proc. phys. Soc.*, 1st Nov. 1958, Vol. 72, No. 467, pp. 910–914.) A *p-n-p* alloy-junction transistor was connected in the grounded-emitter configuration, the base current being fed from a high-impedance source. A probe was placed perpendicular to the base adjacent to the emitter pellet. The transistor was used in the Bardeen-Brattain ambient cycle and the variations of the collector-to-base current gain measured.

621.314.7 **634**

Transition Frequency and Phase Characteristics of a Transistor with Common Emitter.—E. I. Adirovich & K. V. Temko. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1174–1181.) A theoretical treatment of the problem.

621.314.7 **635**

New Transistor Design—the 'Mesa'!—C. H. Knowles. (*Electronic Ind.*, Aug. 1958, Vol. 17, No. 8, pp. 55–60.) Constructional details are given of a new class of miniature transistor suitable for the 10–20 000 Mc/s range, having great reliability and stability. The base-collector junction is formed by vapour diffusion and the emitter-base junction by high-vacuum evaporation alloying. No alloys are involved in the formation of the collector junction thereby reducing the possibility of thermal runaway.

621.314.7-71 **636**

Increased Cooling for Power Transistors.—C. Booher. (*Electronic Ind.*, Aug. 1958, Vol. 17, No. 8, pp. 66–68.) The most effective cooling was obtained using an assembly of metal fims. Temperature-rise characteristics are given for various configurations.

621.314.7 : 546.289 **637**

Germanium Diffused Minicrystals and their Use in Transistors.—I. A. Lesk & R. E. Coffman. (*J. appl. Phys.*, Oct. 1958, Vol. 29, No. 10, pp. 1493–1494.) The process yields a Ge *p-n-p* bar-type structure with fewer practical limitations on emitter, base, and collector resistivities and base width than other processes. Application of developmental units at v.h.f. has been limited by base lead overlap capacitance.

621.383.5 **638**

Determination of the Parameters of Silver Sulphide Barrier-Layer Photocells.—S. V. Svechnikov. (*Zh. tekh. Fiz.*, May 1957, Vol. 27, No. 5, pp. 914–918.)

621.383.5 : 546.289 : 621.396.822 **639**

The Flicker Effect in *p-n* Junction Photovoltaic Diodes.—M. Teboul & N. Nifontoff. (*C. R. Acad. Sci., Paris*, 5th May 1958, Vol. 246, No. 18, pp. 2591–2594.) Report of measurements of the flicker effect in Ge photocells as a function of illumination and applied voltage.

621.383.8 : 546.28 **640**

New Developments in Silicon Photovoltaic Devices.—M. B. Prince & M.

Wolf. (*J. Brit. Instn Radio Engrs*, Oct. 1958, Vol. 18, No. 10, pp. 583–594. Discussion, pp. 594–595.) A discussion and analysis of the performance of three types of *p-n* junction devices prepared by solid-state diffusion methods, (*a*) a solar cell, suitable for moderately low to high light levels, (*b*) a low-level cell, and (*c*) a photodiode for low to high levels. Spectral response, transient response, and temperature dependence are considered.

621.385.029.6 **641**

Development of Electronic Devices for Extremely High Frequencies.—N. D. Devyatkov. (*Izv. Ak. Nauk S.S.S.R., Otd. tekh. Nauk*, Feb. 1958, No. 2, pp. 104–113.) A review of the development during the past 20 years of various kinds of valve oscillator for the metre-, decimetre- and centimetre-wave bands.

621.385.029.6 **642**

New Developments in Wide-Band Microwave Tubes.—D. A. Dunn. (*Electronic Ind.*, Aug. 1958, Vol. 17, No. 8, pp. 72–78.) New methods of beam focusing and new circuits for high-power wide-band amplifiers are discussed. Valve types available in the U.S.A. in May 1957 are tabulated. 28 references.

621.385.029.6 **643**

Design of Broad-Band Ceramic Coaxial Output Windows for Microwave Power Tubes.—R. R. Moats. (*Sylvania Technologist*, July 1958, Vol. 11, No. 3, pp. 86–90.) An analysis is made of a design for broad-band matching by undercutting the centre conductor much less than is required for constant Z_0, and extending the undercut a significant distance each side of the ceramic window.

621.385.029.6 **644**

Current Distribution in Modulated Magnetically Focused Electron Beams.—M. Chodorow, H. J. Shaw & D. K. Winslow. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1525–1533.) Detailed measurements have been made of the d.c. and r.f. current distribution in a modulated, magnetically focused electron beam having normalized parameters in the range of values appropriate for practical medium- and high-power klystrons. The ratio of the total r.f. current to the total direct current in the beam as a function of drift distance was determined experimentally, the experimental values being compared with the results predicted theoretically.

621.385.029.6 **645**

Contribution to the Diffusion Theory of the Magnetron (Static Condition).—L. E. Pargamanik & M. Ya. Mints. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1301–1305.) The theory accounts for the rapid increase in temperature of the electron gas with increasing magnetic field and shows good agreement with experimental observations.

621.385.029.6 **646**

Contribution to the Theory of the Magnetron with a Single Anode.—M. Ya. Mints. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1306–1312.) An application of diffusion theory to the case of small oscillations with particular reference to impedance evaluations. See also 645 above.

621.385.029.6 **647**

Contribution to the Theory of the Magnetron with a Split Anode.—M. Ya. Mints. (*Zh. tekh. Fiz.*, June 1957, Vol. 27, No. 6, pp. 1313–1318.) An extension of the work described in 646 above to the case of the split-anode magnetron.

621.385.029.6 **648**

Pulser Component Design for Proper Magnetron Operation.—P. R. Gillette & K. Oshima. (*Trans. Inst. Radio Engrs*, March 1956, Vol. CP-3, No. 1, pp. 26–31. Abstract, *Proc. Inst. Radio Engrs*, June 1956, Vol. 44, No. 6, Part 1, p. 832.)

621.385.029.6 **649**

Helices for Travelling-Wave Valves: Effect of Supports; Attenuation; Parasitic Modes.—P. Lapostolle. (*Ann. Télécommun.*, Feb. 1957, Vol. 12, No. 2, pp. 34–59.) Charts are derived to facilitate the design of travelling-wave amplifiers. Anomalies in operation are also discussed. A table is given of equivalent notations used by American authors.

621.385.029.6 : 621.372.8 **650**

Propagation Characteristics of Slow-Wave Structures Derived from Coupled Resonators.—E. Belohoubek. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 283–310.) A general method is given for finding qualitatively the ω-β_0 diagram for slow-wave structures of the coupled-resonator type. The application of different coupling systems to slow-wave structures is discussed. The qualitative considerations are compared with some measurements made on a circular waveguide with differently shaped partition walls. See also 306 of 1955 (Nalos).

621.385.032.213 **651**

A Gas-Evolution Controlled Servo System for the Processing of Oxide-Coated Cathodes.—R. P. Misra & W. H. Moll. (*Le Vide*, March/April 1957, Vol. 12, No. 68, pp. 167–175. In French & English.) Gas outbursts during breakdown are controlled by a d.c. error voltage proportional to the increase in pressure. This voltage, derived from an ionization gauge within the vacuum system, controls the heater voltage of the valve being processed.

621.385.032.263 **652**

The Annular-Geometry Electron Gun.—J. W. Schwartz. (*Proc. Inst. Radio Engrs*, Nov. 1958, Vol. 46, No. 11, pp. 1864–1870.) A new type of kinescope electron gun of high resolution is described. High modulation sensitivity, inverted modulation characteristics, internal electronic video signal amplification and automatic 'white noise' inversion are features of this system.

621.385.032.269.1 **653**

The Theory of the Pierce-Type Electron Gun.—P. T. Kirstein. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 163–164.) Comment on 1929 of 1958 (Radley).

621.385.1-71 **654**

A New Method of Cooling High-Power Valves by Vaporization of Water.—P. E. Cane & W. E. Taylor. (*J. Brit. Instn Radio Engrs*, Oct. 1958, Vol. 18, No. 10, pp. 621–626.) The vapotron technique is described [see also 2640 of 1957 (Beurtheret)]. Such systems have operated satisfactorily for several years on high-power valves at frequencies between 500 kc/s and 200 Mc/s.

621.385.3 **655**

The PCC88 High-Frequency Double Triode.—(*Electronic Applic. Bull.*, Jan. 1958, Vol. 18, No. 1, pp. 27–36.) Constructional details, characteristics and applications of a high-slope, low-noise-factor valve, suitable for cascode circuits.

621.385.832 : 621.397.62 : 535.623 **656**

A New Cathode-Ray Tube for Monochrome and Colour Television.—Gabor, Stuart & Kalman. (See 619.)

621.387 : 621.396.822.029.63 : 621.317.7 **657**

Application of Gas-Discharge Tubes as Noise Sources in the 1 700–2 300-Mc/s Band.—Kollanyi. (See 559.)

621.387 : 621.396.822.029.64 **658**

Measurements on Gas-Discharge Noise Sources at Centimetre Wavelengths.—A. C. Gordon-Smith & J. A. Lane. (*Proc. Instn elect. Engrs*, Part B, Nov. 1958, Vol. 105, No. 24, pp. 545–547.) Measurements using a thermal noise source and a c.w. signal give values of 10 590 $\pm$ 500°K and 11 050 $\pm$ 1 350°K respectively for the effective noise temperature of the Type CV 1881 argon discharge tube at 3 cm λ.

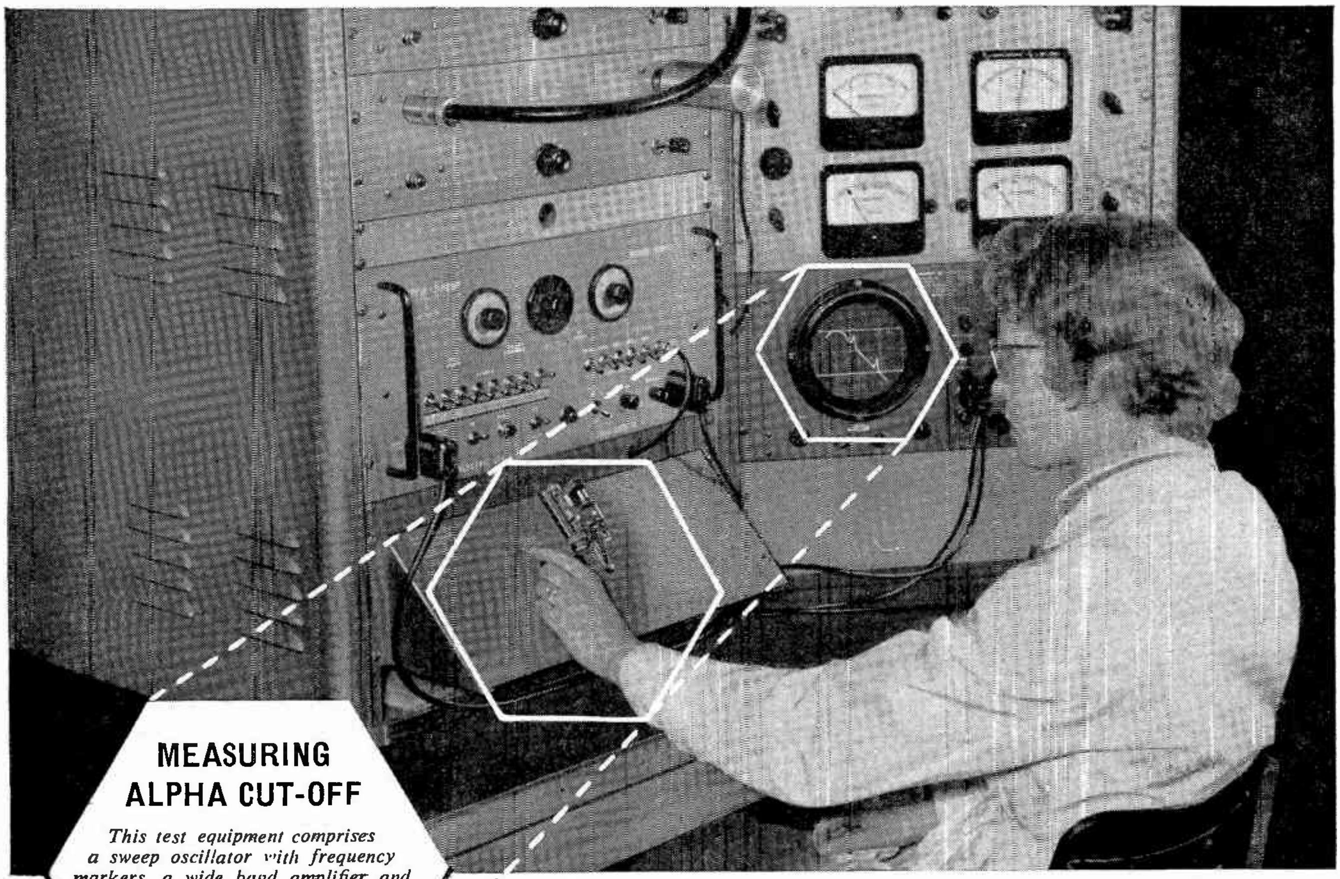
MISCELLANEOUS

621.3.002.3 : 519.27 **659**

Selection of Matched Components from Random Samples.—D. P. C. Thackeray. (*Electronic Radio Engr*, Dec. 1958, Vol. 35, No. 12, pp. 473–476.) Special reference is made to the selection of transistors from random samples, the selection of such samples from stocks and the stocking of quantities which are adequate for such procedures.

413.164 = 82 = 20 **660**

Russian-English Electronics and Physics Glossary. [Book Notice]—Publishers: Consultants Bureau, New York, $10. (*J. Electronics Control*, July 1958, Vol. 5, No. 1, p. 88.) Part 3 of eight interim glossaries on specialized fields of physics. A ten-page appendix covers U.S.–Soviet valve and unit equivalents, circuit components, notations and abbreviations.



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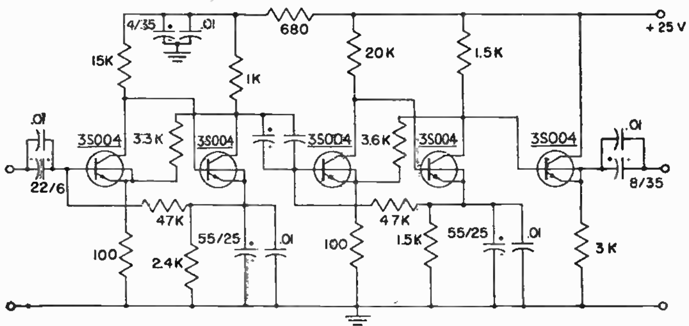
Collector dissipation at 100°C.	50 mW
Collector Breakdown voltage	30 volts
Collector current	10 mA

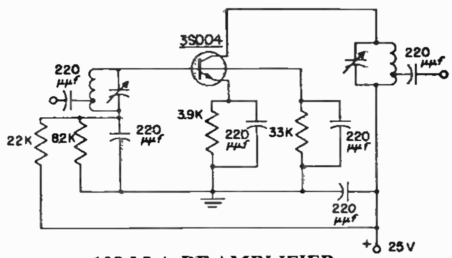
SPECIFICATION

Type No.	*Power Gain*
3S002 (3N34)	16 db (min) at 30 Mc/s.
3S004 (3N35)	18 db (min) at 70 Mc/s.

TWO CIRCUIT EXAMPLES



10 Mc/s VIDEO AMPLIFIER
Measured Power Gain, 53 db. Bandwidth 10 Mc/s. Load Used, 470 Ohms.



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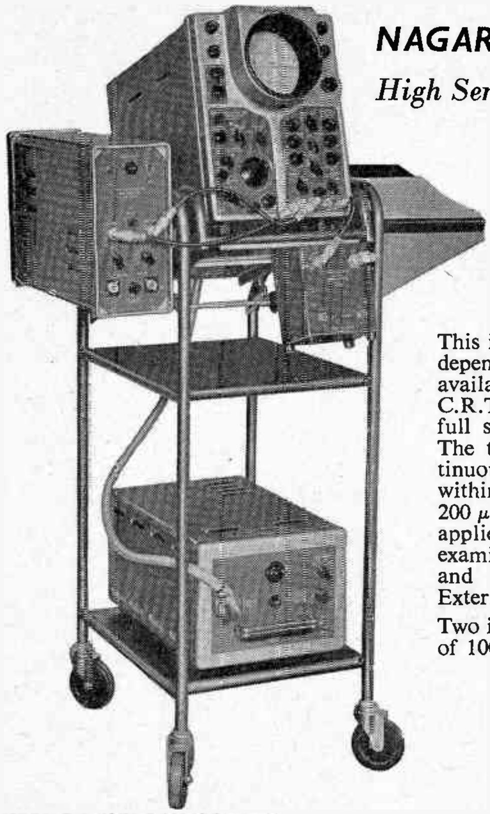

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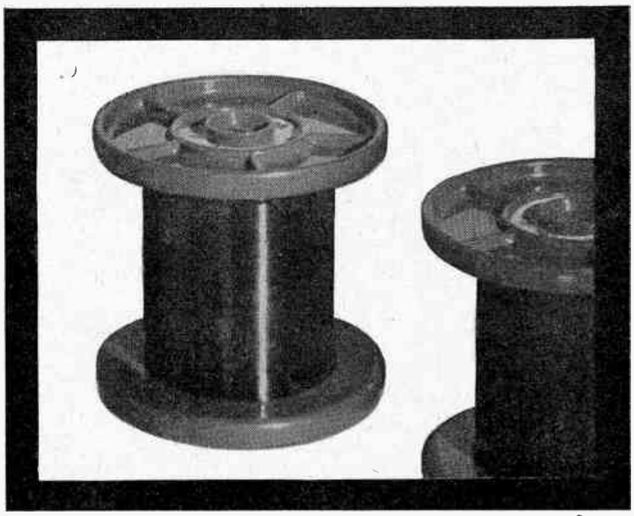

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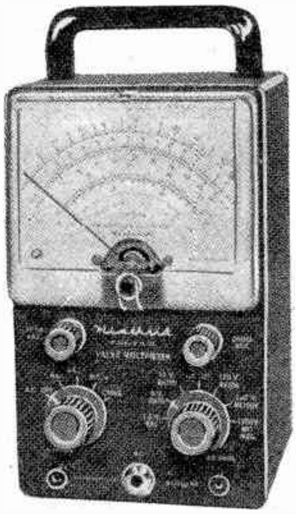

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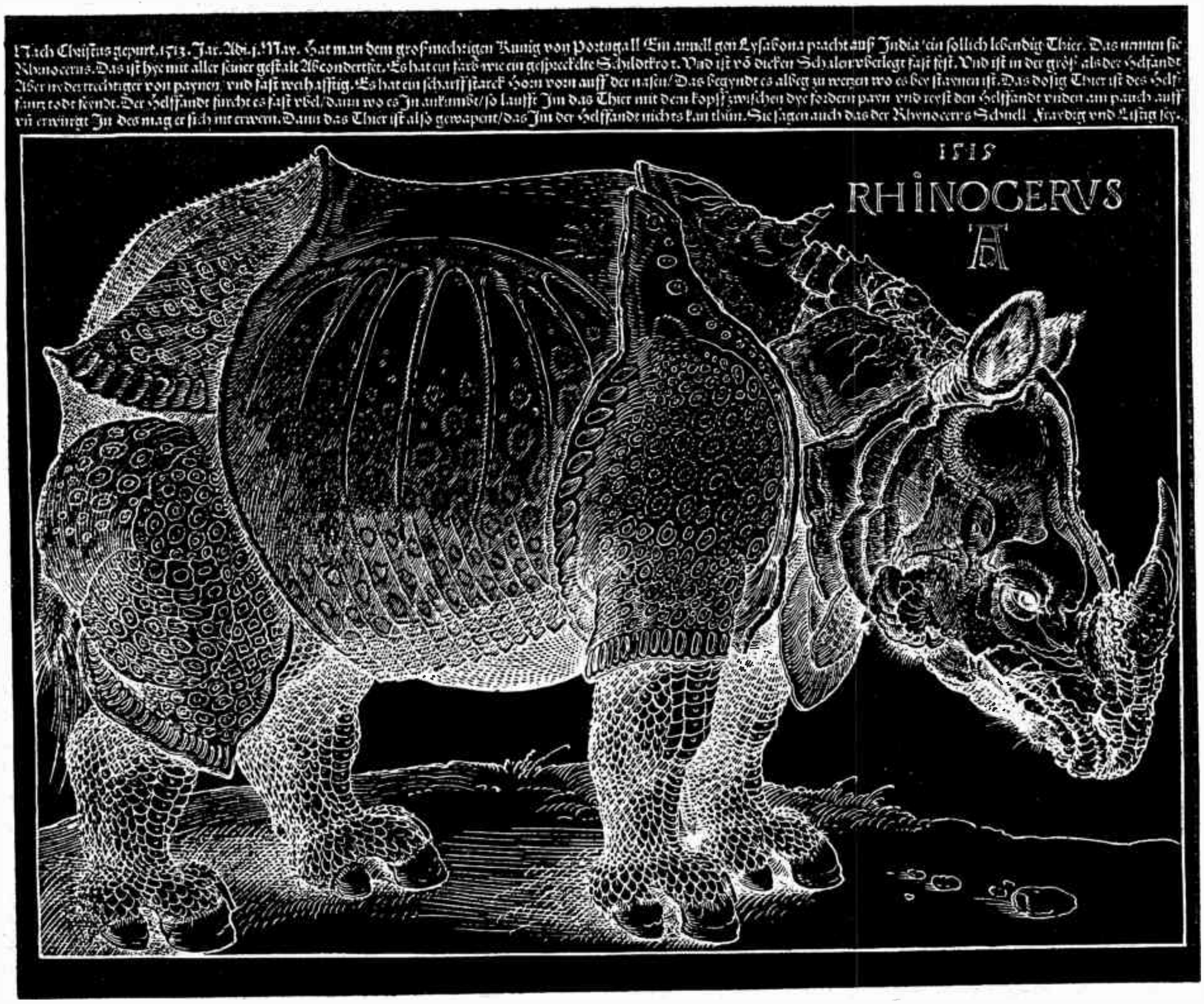
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Press date for the March 1959 issue is first post 20th February 1959

SITUATIONS VACANT

THE COLLEGE OF AERONAUTICS

THE Department of Aircraft Electrical Engineering is planning considerable developments in the teaching and research aspects of Feedback Control Systems, Simulation Techniques and Process Control. Existing courses include: (i) two-year Diploma Course in Aircraft Electrical Engineering; (ii) one-year Advanced Course in Guided Weapon Control Systems and (iii) Guided Weapon Guidance Systems. Preparation is also being made for a new course in Process Control.

APPLICATIONS are invited for the following appointments:

DEMONSTRATOR with preferably some experience in one of the above fields or alternatively with a general electronic background. Responsibilities will eventually include the everyday running of the simulation and control laboratories, development of new equipment, supervision of students' practical work and assistance with research activities. The appointment should appeal to a young man who wishes to consolidate his theoretical knowledge and participate in research. Candidates should hold a university degree or equivalent professional qualification. Salary scale £600 x £25 to £1,000 p.a. with superannuation under F.S.S.U. and family allowance at £25 per annum for each dependent child. Consideration would be given at a later date to possible promotion to Lecturer.

RESEARCH Engineer preferably with experience in one of the following subjects: aircraft or G.W. control systems; servo-mechanisms, analogue computers; instrumentation; communications or radar. Successful candidate would collaborate with the academic staff in the design and development of equipment associated with teaching and research activities. Preference will be given to applicants holding a university degree or equivalent professional qualification. Salary on various scales up to £1,200 p.a. with Local Government Superannuation; family allowance depending upon grade of appointment.

RESEARCH Assistant to assist in the development of interesting and novel applications of simulation techniques. Successful candidate, who should have a good practical knowledge of electronics, would be a member of a small team engaged on this project. Preference will be given to candidates with H.N.C. in electrical engineering or equivalent qualification. Salary up to £900 p.a. The appointed candidate would be considered for eventual promotion.

IN all cases the commencing salary within the prescribed scale will depend upon age, qualifications and experience. Consideration will be given to candidates' housing requirements.

APPLICATIONS giving full particulars and the names and addresses of three referees should be forwarded to the Recorder, The College of Aeronautics, Cranfield, Bletchley, Bucks. [1291

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THE Department of Aircraft Electrical Engineering is proposing to expand its activities in the field of E.M. Machines and Electrical Power Systems. This new section will be responsible for more extensive lectures in the 2nd year Diploma Course in Aircraft Electrical Engineering and for the preparation of an advanced one-year course in Aircraft Electrical Power Systems. The section will also be responsible for the installation and running of a large Electrical Power System which will be used for teaching, research, and development purposes.

APPLICATIONS are invited for the following appointments:

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DEMONSTRATOR whose responsibilities will include the everyday running of the new Electrical Power System Laboratory, development of new equipment, supervision of students' practical work, and assistance with research. Candidates should have a university degree or equivalent professional qualification and should preferably have had experience in a suitable field. Salary scale £600 x £25 to £1,000 p.a. with superannuation under F.S.S.U. and family allowance at £25 per annum for each dependent child. Consideration would be given at a later date to possible promotion to Lecturer.

RESEARCH Engineer preferably with experience in one of the following subjects: electrical machines; aircraft power systems; guided weapon electrical systems. Successful candidate would collaborate with the academic staff in the design and development of equipment associated with teaching and research activities. Preference will be given to candidates with a university degree or equivalent professional qualification. Salary on various scales up to £1,200 p.a. with Local Government Superannuation; family allowance depending upon grade of appointment.

IN all cases the commencing salary within the prescribed scale will depend upon age, qualifications and experience. Consideration will be given to candidates' housing requirements.

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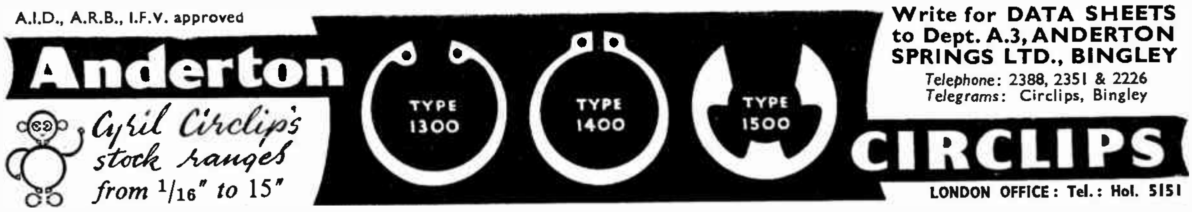

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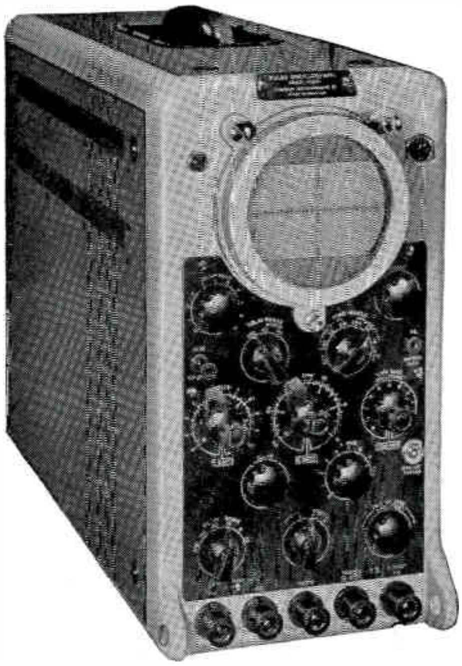

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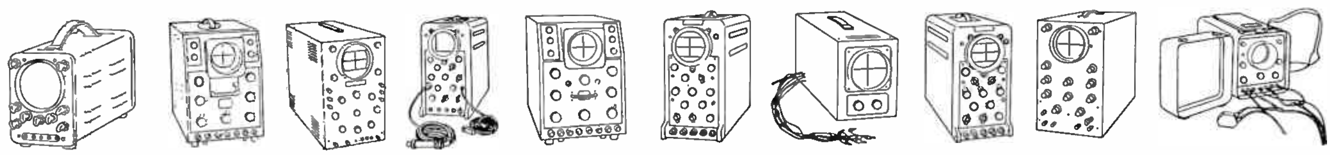

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