ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Three shillings and sixpence

JANUARY 1959 Vol 36 *new series* No 1

For continuous use at 130°C

Teramel is BICC's new polyester enamel covering for winding wires. It combines the excellent electrical and mechanical properties called for in BS 1844/1952 with high thermal stability—*Teramel wires can safely be used at continuous temperatures of 130°C.*

They are ideal for:—

- **Armature and field windings for industrial and traction motors**
- **Air cooled windings for transformers**
- **Coils for motor starters**

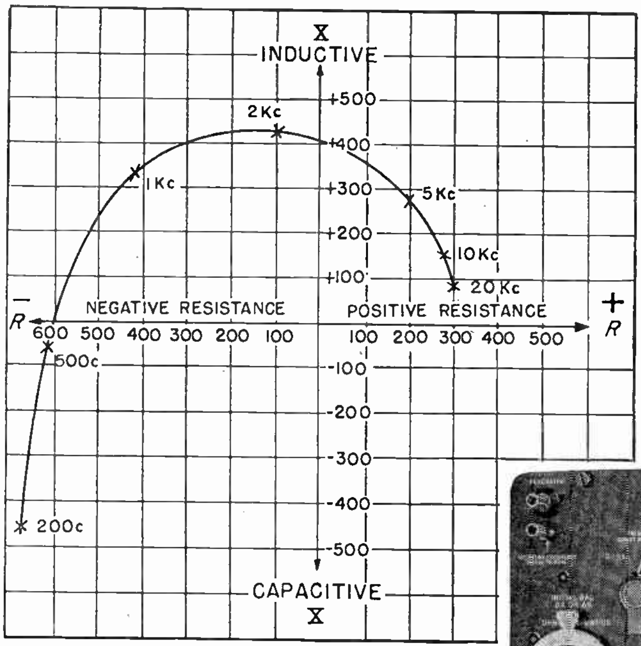
BICC TERAMEL

Winding Wires

Hard and strongly adhesive to the copper wire. Negligible thermoplastic flow.	Flexible—can be twisted, stretched or flattened without damage.	Resistant to varnish solvents, moisture and chemically contaminated atmospheres.

Further information is contained in Publication No. 391—available on request

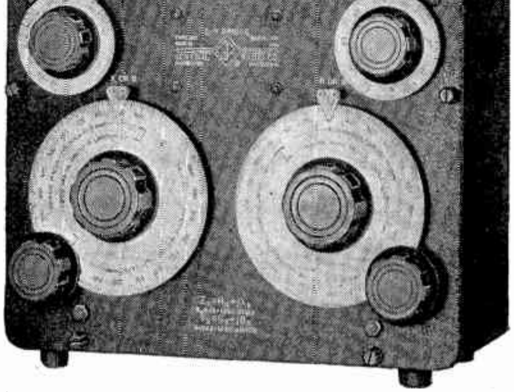
BRITISH INSULATED CALLENDER'S CABLES LIMITED. 21 Bloomsbury Street, London, W.C.1



Impedance of Feedback Circuit ... illustrates ability of the Z-Y Bridge to measure any impedance; quadrature components may be positive or negative, real or imaginary.

TRADE MARK

Type 1603-A Z-Y Bridge



Measures Any Impedance...

- From 0 to ∞ ohms
- Positive or Negative
- At Any Phase Angle
- Over 20-c/s to 20-kc Range

The General Radio Z-Y Bridge measures impedances from short circuit to open circuit, at small or large phase angle. Quadrature components, R & X or G & B, are measured directly at calibrated 100 c, 1-kc, and 10-kc bridge positions. Basic accuracy is 1% over most of this range.

The ability to measure impedances of any magnitude accurately with one instrument is an extremely valuable asset in many measurement situations. The Z-Y Bridge can be used for measuring conductivity of liquids, in dielectric cells as readily as it can be used for R-L-C component measurements in the laboratory or production-test department. It will measure open-circuit and short-circuit transformer parameters . . . impedances of batteries and electrolytic capacitors . . . characteristics of audio-transmission networks . . . impedance of electro-acoustic transducers . . . Q and resonant frequency of chokes . . . and impedances of feedback loops, since negative real parameters can be directly measured.

The Bridge also can be used to determine cable-fault locations and circular-arc plots of liquids or solids having lossy polarizations in the audio-frequency range. These are but a few of the countless applications for this unique and versatile device. *You name it — this Z-Y bridge can probably measure it.*

For complete information request a copy of the current "G.R." Catalogue "O" (258 pages), where data is given on pages 34/35.

Impedance and Admittance Range

R: ±1000 ohms	G: ±1000 μmhos
X: ±1000 ohms	B: ±1000 μmhos

Accuracy

R or G: ±(1% + (2 ohm or 2 μmho))

X or B: ±(1% + ($\frac{2f_0}{f}$ ohm or $\frac{2f}{f_0}$ μmho))

f is operating frequency, f_0 is frequency setting of panel selector switch

Impedances of less than 100Ω or (100 μmhos) can be measured on "Initial Balance" dials with considerably greater accuracy—

R or G: ±(1% + (0.2 ohm or 0.2 μmho))

X or B: ±(1% + ($\frac{0.2f_0}{f}$ ohm or $\frac{0.2f}{f_0}$ μmho))

Frequency Range—20 cycles to 20 kc

Maximum Applied Voltage

130 volts, rms on bridge;
less than 32v on unknown

Accessories Recommended

"G.R." Type 1210-B Unit R-C Oscillator and
"G.R." Type 1212-A Unit Null Detector

Accessories Supplied

2 Shielded Cables for generator and detector

Dimensions—12½" x 13¼" x 8½"

Net Weight—21½ lbs.

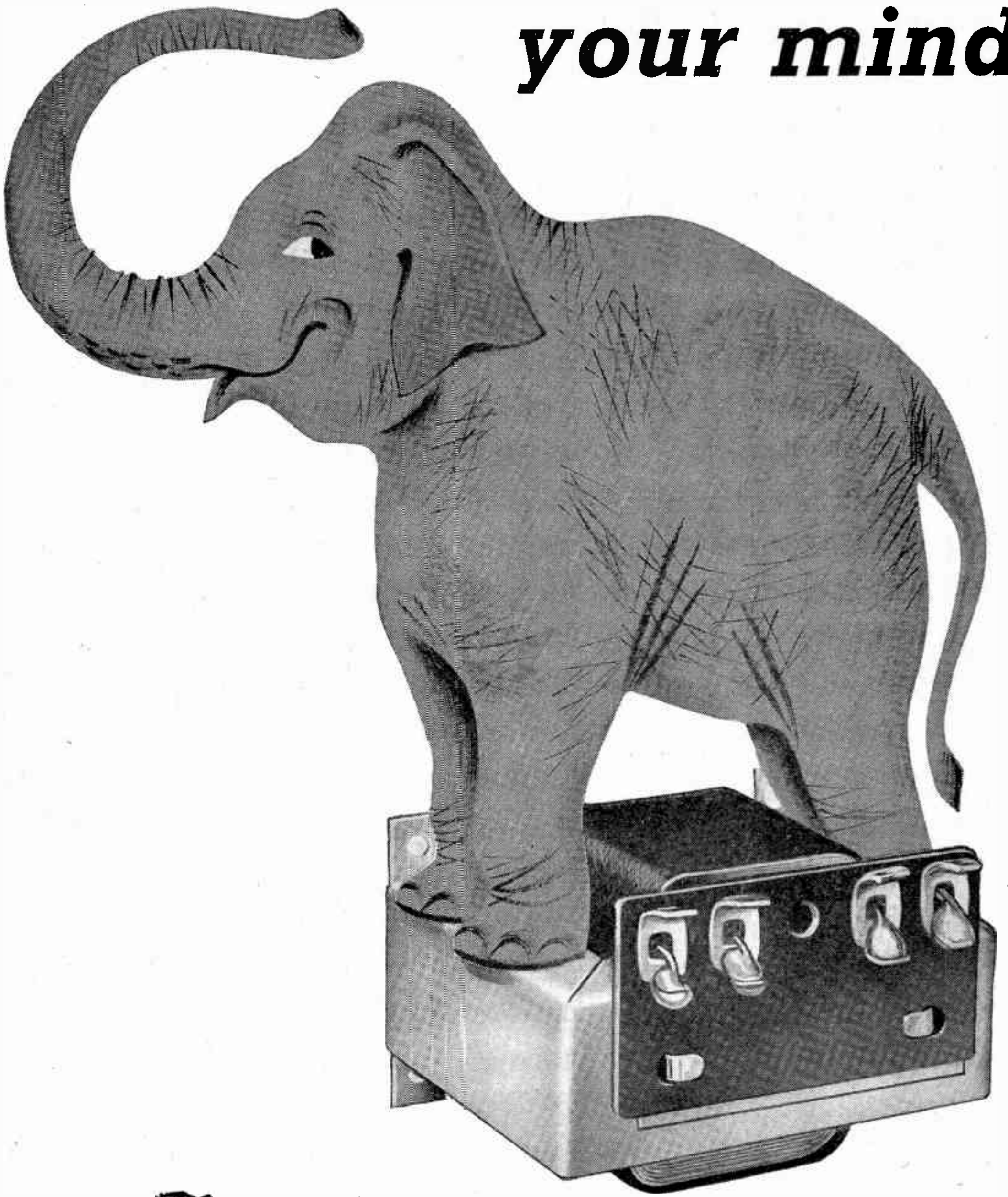
Type 1603-A Z-Y Bridge £222

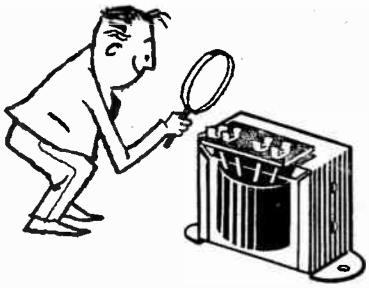
Claude Lyons Ltd.

76 Oldhall Street Liverpool 3, Lancs.
Telephone: Central 4641/2

Valley Works, Ware Road, Hoddesdon, Herts.
Telephone: HODdesdon 3007-8-9

CL.45

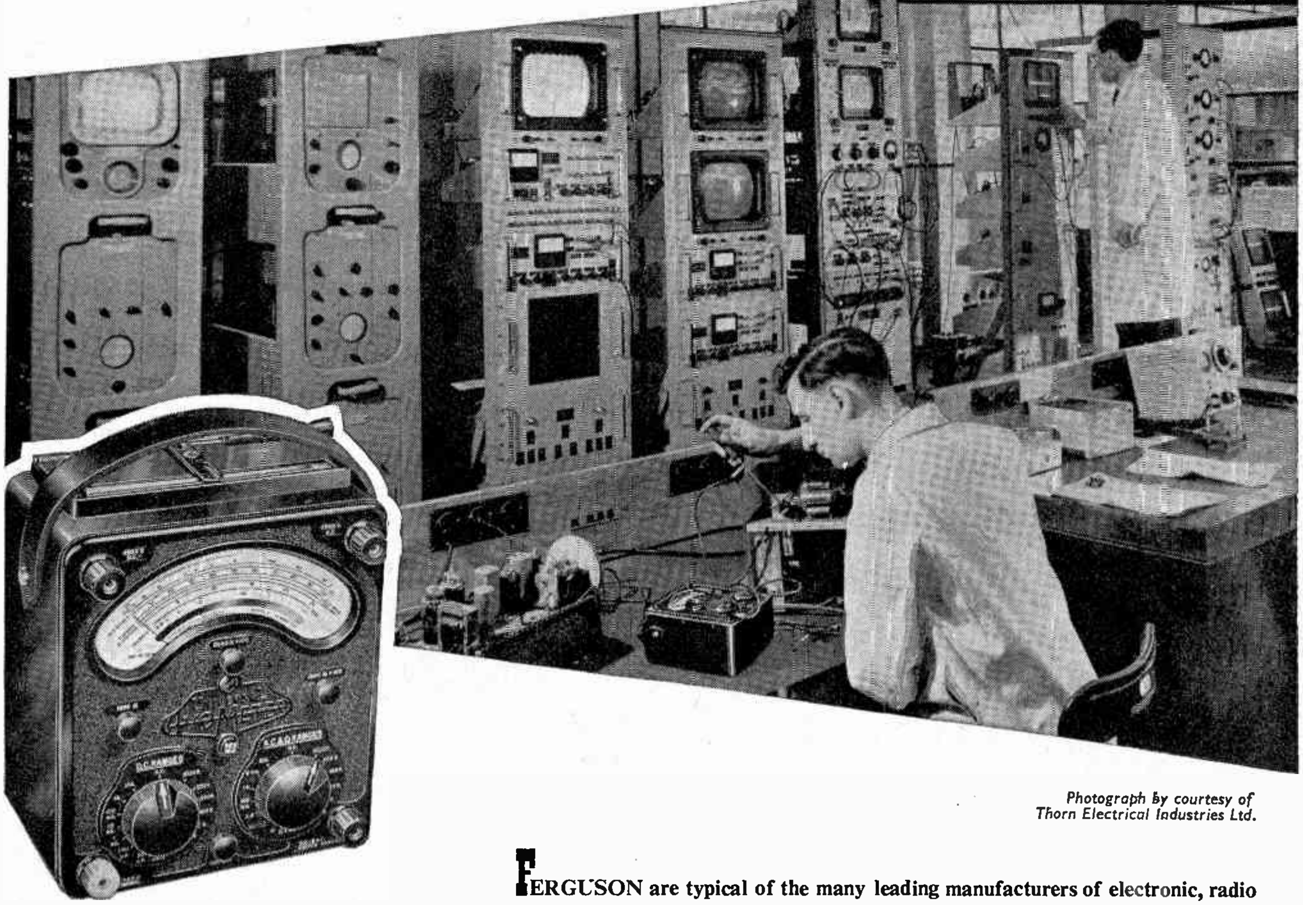




FERGUSON choose the

Model 8 AvoMeter

for their Television & Electronic Laboratory



Photograph by courtesy of
Thorn Electrical Industries Ltd.

FERGUSON are typical of the many leading manufacturers of electronic, radio and television equipment who rely on AVO instruments.

The Model 8 AvoMeter shown in use is a 30-range self-contained A.C./D.C. moving coil instrument, produced primarily for the electronic, radio and television engineer. The upper photograph shows a mounted pivot under examination. This is only one of many operations carried out in a special air-conditioned dust-free zone in the AVO factory to ensure the highest possible standards of accuracy and reliability.

A.C. VOLTAGE: 25 mV. to 2,500 volts.
A.C. CURRENT: 1 mA. to 10 amps.
D.C. VOLTAGE: 25 mV. to 2,500 volts.
D.C. CURRENT: 0.5 microamps to 10 amps.
SENSITIVITY:
20,000 ohms per volt on all D.C. ranges.
1,000 ohms per volt on A.C. ranges
from 100 volts upwards.
RESISTANCE:
0 to 20 megohms (using internal batteries).
0 to 200 megohms (using external D.C. supply).
DECIBELS: —15dB to +15dB.

Various accessories are available for extending the above ranges.

Size: $8\frac{1}{4} \times 7\frac{1}{4} \times 4\frac{1}{2}$ inches.
Weight: $6\frac{1}{2}$ lbs. (approx.)
including leads.

List Price: **£23 : 10s.**

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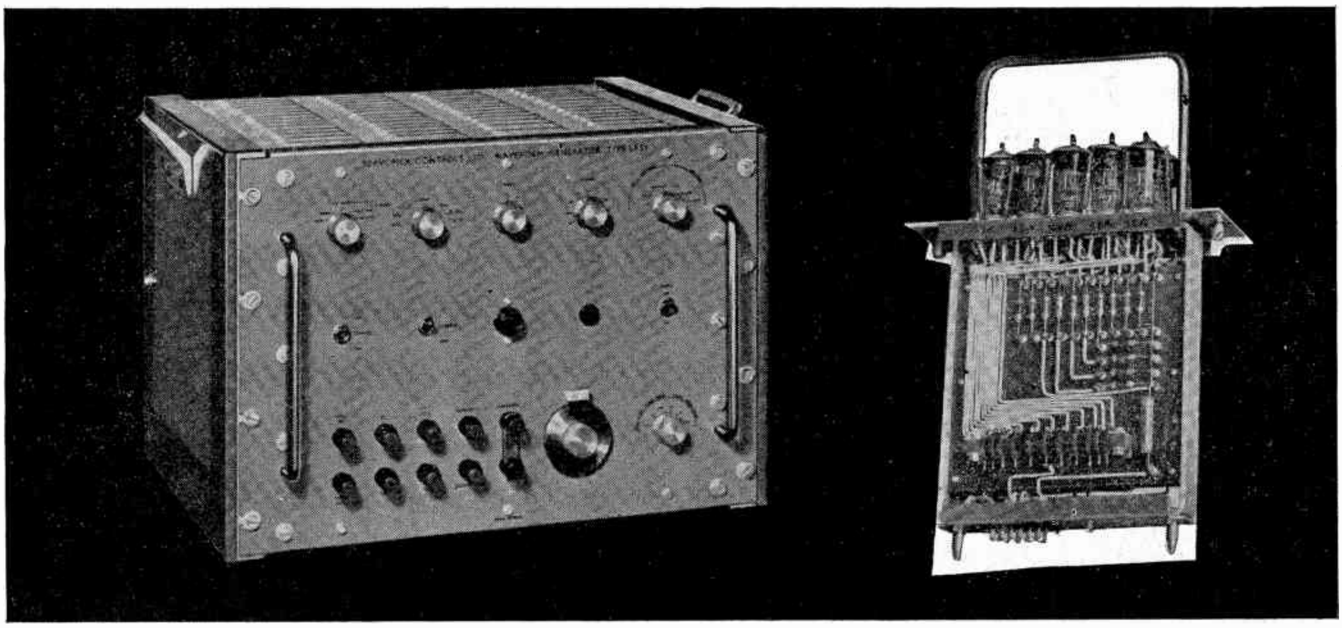
Please send fully illustrated Brochure describing the Model 8 AvoMeter. ERE

Unchallenged

SERVOMEX LF.51 Still the world's best low frequency waveform generator!



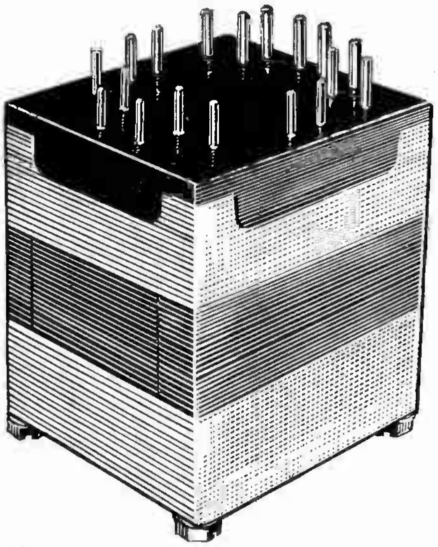
Still the most flexible supply of test signals for testing servo mechanisms and automatic controllers

This unique instrument is being used by discriminating control engineers all over the world for measuring the dynamic response of automatic control systems.
With it, one can carry out measurements of the frequency response to sinusoidal input; the step-function response; the response to ramp functions; the response to sine-squared pulses; and many other special tests.

- ★ *Sine waves from 500 cycles/sec. down to one cycle every 2,000 secs. (33 minutes)*
- ★ *Pulses and square waves with rise time of 5 microseconds*
- ★ *Ramp functions with a linear rise variable from 1 millisecond to 1,000 seconds*
- ★ *Over 30 different waveforms available*
- ★ *Voltage variable between 100 microvolts and 150 volts peak-peak*
- ★ *Load current up to 5 milliamps peak*

1. Selector switch at "CONTINUOUS"	
2. Selector switch at "SINGLE" - a step with continuously variable slope.	
3. Selector switch at "DOUBLE" - a single isolated pulse in TEN different forms	

SERVOMEX CONTROLS LIMITED, CROWBOROUGH SUSSEX. CROWBOROUGH 1247



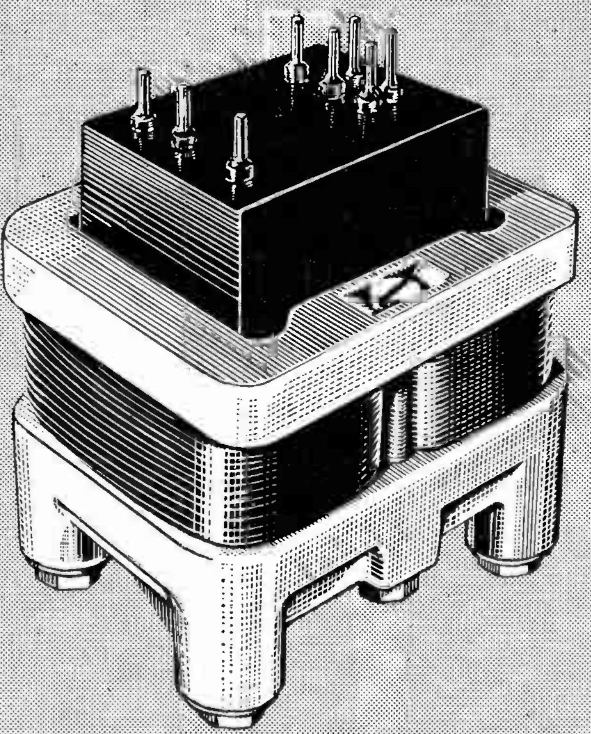
R120 and R130 laminated series.

FORTH SERIES

TRANSFORMERS AND CHOKES

smaller in size – lighter in weight

R200 'C' Core series.



FERRANTI

The new range of Ferranti Resin Cast Transformers and Chokes has been named after this famous Scottish landmark which represented a remarkable advance in engineering design when it was constructed over 60 years ago. To-day, the new techniques in manufacture and construction of 'C' Core Transformers have enabled Ferranti Ltd. to make a significant contribution to Electronic Engineering.

The Forth series components will have particular appeal to designers of airborne equipment since savings in weight and volume of up to one-third can be achieved over the resin cast and oil-filled units now available. Moreover, the quality requirements of the Joint Service Specification RCS.214 are met in every respect. Please write for a catalogue which gives full rating information.

FERRANTI

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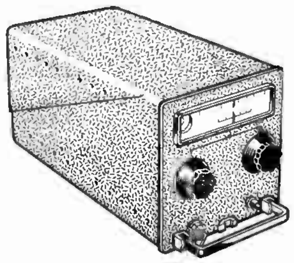
ES/T41

A.W.A ELECTRONICS

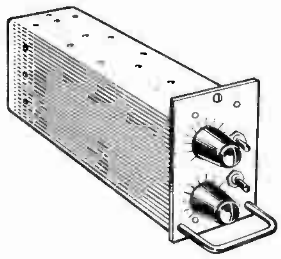
New concepts in electronics have been developed at AWA, as a result of experience with missile systems. Now they have a wider application. Here are some of the new AWA devices now available to industry.



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.*

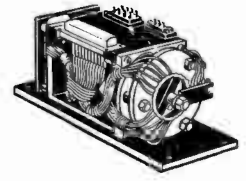
TRANSISTOR GALVANOMETER AMPLIFIER



This Amplifier has been designed to drive viscous damped recording galvanometers which normally have a resistance of 50 ohms and a working range of D.C. 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: 10 Microvolts at either input; Input impedance: 40,000 ohms on range 5, 110,000 ohms all other ranges; Gain: Maximum 5 Milliamps/Millivolt, minimum 0.04 Milliamps/Volt; Power requirements: ± 6 Volts D.C. 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:

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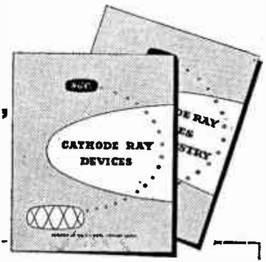
Baginton, Coventry.

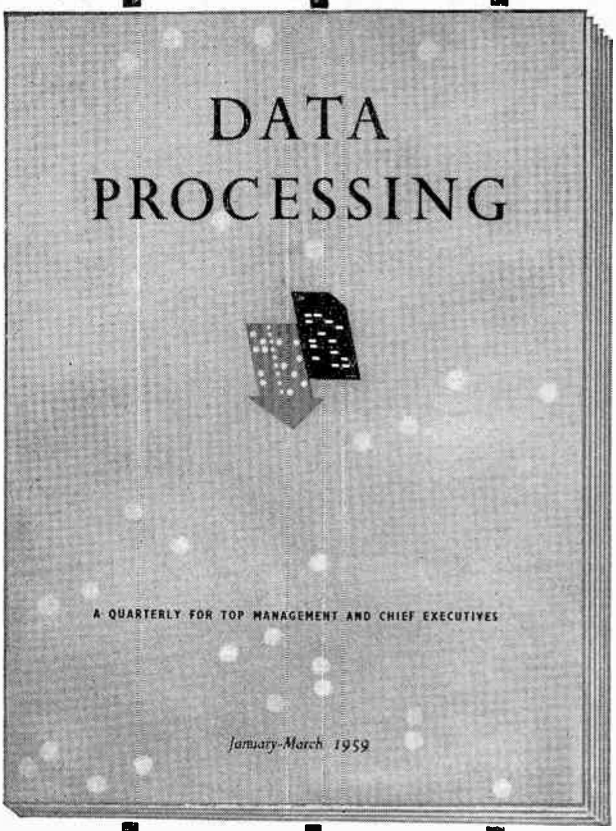
MEMBER OF THE HAWKER SIDDELEY GROUP


CATHODE RAY DEVICES

1 **This new journal provides today the information which management must have for higher efficiency tomorrow**

2 AUTOMATIC AIDS TO CONTROL, ADMINISTRATION AND ROUTINE WORK IN OFFICE AND FACTORY CAN BRING RICH REWARDS TO THOSE WHO FIRST APPLY THEM

3 *Here is the top executive's guide to the choice of equipment and its correct application*



With new methods of business and industrial control, important decisions, on which profits and progress depend, can be based on a complete and accurate assessment of all relevant data. With these methods, too, routine administrative efficiency can be raised and its cost substantially reduced.

DATA PROCESSING, a new quarterly journal in the Associated Iliffe group, will describe in each issue the means by which this can be achieved, surveying the whole range of automatic aids to good management, commercial and industrial. Computers, punched card machinery and peripheral equipment will be examined and the best of the current operational practices presented in a form readily applicable to particular problems. Completion of the form below is the first move in ensuring that your organisation is early among those who benefit from these new methods.

FIRST ISSUE *just published!*

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World Radio History

These photocells give you the simplest photo-electric control possible

Photo-electric control with the Mullard ORP11 and ORP90 cadmium sulphide cells is the simplest possible because a photocell and relay form the complete circuit.

The unusual combination of high current capacity and extreme sensitivity of these Mullard cells enables robust relays to be operated direct—amplifiers are unnecessary.

Both cells can be operated from either a.c. or d.c. supplies, they are inherently rugged and have a wide range of applications in industry.

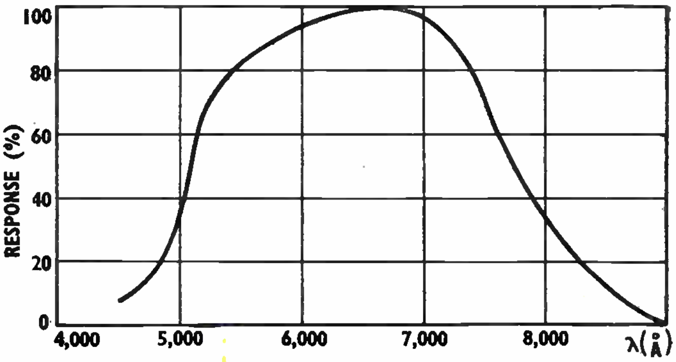
The usable response extends through the entire visible spectrum to the near infra-red.

The ORP11 differs from the ORP90 chiefly in being "end-viewing" and having a somewhat smaller photocathode area. This type of photocell is made available to simplify mounting problems encountered in certain applications—particularly in flame failure detectors in oil fired furnaces.

Data sheets giving further information are readily available from the address below.

ABRIDGED DATA

	ORP11	**ORP90**
Required direction of incident light	End-on	Side-on
Area of photo-element	1.25 sq. cm.	2.9 sq. cm.
Average cell current at 10V d.c., 5 foot candles and lamp colour temperature 2,700°K.	6mA	6mA
Maximum ultimate dark current at 100V d.c.	5 µA	< 2.5 µA
Maximum cell dissipation at 25°C.	200 mW	600 mW
Spectral response	Same for both cells—see curve.	



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Telephone: Langham 6633

Mullard
GOVERNMENT AND
INDUSTRIAL VALVE DIVISION

MVT 360

Striking success

Electrical tools for breaking concrete call upon the remarkable mechanical as well as electrical properties of Araldite epoxy resins. They play no small part in the success of this Kango H Type Breaker which must combine toughness with durability under the most arduous conditions of use.

Araldite B is the form of the resin used for potting the armature of this breaker, and a special system of winding was evolved to allow maximum penetration and so produce an armature able to withstand vibration, high temperatures and moisture. In this, as in other applications, Araldite offers the advantages of outstanding adhesion to metals, negligible shrinkage on setting, excellent electrical properties and absolute stability.

May we send you full descriptive literature on Araldite epoxy resins and adhesives?

Araldite epoxy resins are used

- **for bonding metals, ceramics etc.**
- **for casting high grade solid insulation**
- **for impregnating, potting or sealing electrical windings and components**
- **for producing glass fibre laminates**
- **for making patterns, models, jigs and tools**
- **as fillers for sheet metal work**
- **as protective coatings for metals, wood and ceramic surfaces**

Araldite *epoxy resins*

Araldite is a registered trade name

CIBA (A.R.L.) LTD Duxford, Cambridge. Telephone: Sawston 2121

AP338

Semiconductors

COMPUTER TRANSISTORS

The Semiconductors range of Computer Transistors, designed and tested to the special requirements of computer engineers, is the key to a new order of computer speed and reliability. Overall reliability is further increased by making possible a substantial reduction in the number of associated components.

The two types of Silicon Alloy Transistors shortly going into production will make it possible to extend this high-speed computer performance into **ambient** temperatures well above 100°C. **Samples are available now.**

	TYPE	DESCRIPTION	RISE TIME millimicroseconds	Vc max	Ic max
HIGH-SPEED LOW-LEVEL SWITCHING GERMANIUM	SB 344 SB 345	General purpose transistors for conventional logic circuits.	50	5v	5mA
	SB 240	Designed for directly coupled circuits. Controlled input, saturation and hole storage characteristics.	30	6v	15mA
	MA 393	High gain transistor for high-speed driving of parallel circuits.	30	6v	50mA
	2N 501	Ultra-high speed transistor with controlled input and saturation characteristics.	10	12v	50mA
HIGH-SPEED LOW-LEVEL SWITCHING SILICON	SA 495	General purpose 10Mc/s transistor for conventional logic circuits.	100	25v	50mA
	SA 496	15Mc/s transistor for directly coupled circuits. Saturation resistance typically 10 ohms. Controlled input and hole storage characteristics.	80	10v	50mA
CORE DRIVING GERMANIUM	2 N 597	min f_α 3Mc/s — 250 mW high frequency alloy transistors with high gain and low saturation resistance	400 *	20v	400mA
	2 N 598	min f_α 5Mc/s	250 *	20v	400mA
	2 N 599	min f_α 12Mc/s	100 *	20v	400mA
	2 N 600	min f_α 5Mc/s — 750 mW versions of 2 N 598 and 2 N 599. Peak current 3 amps.	250 *	20v	400mA
	2 N 601	min f_α 12Mc/s	100 *	20v	400mA

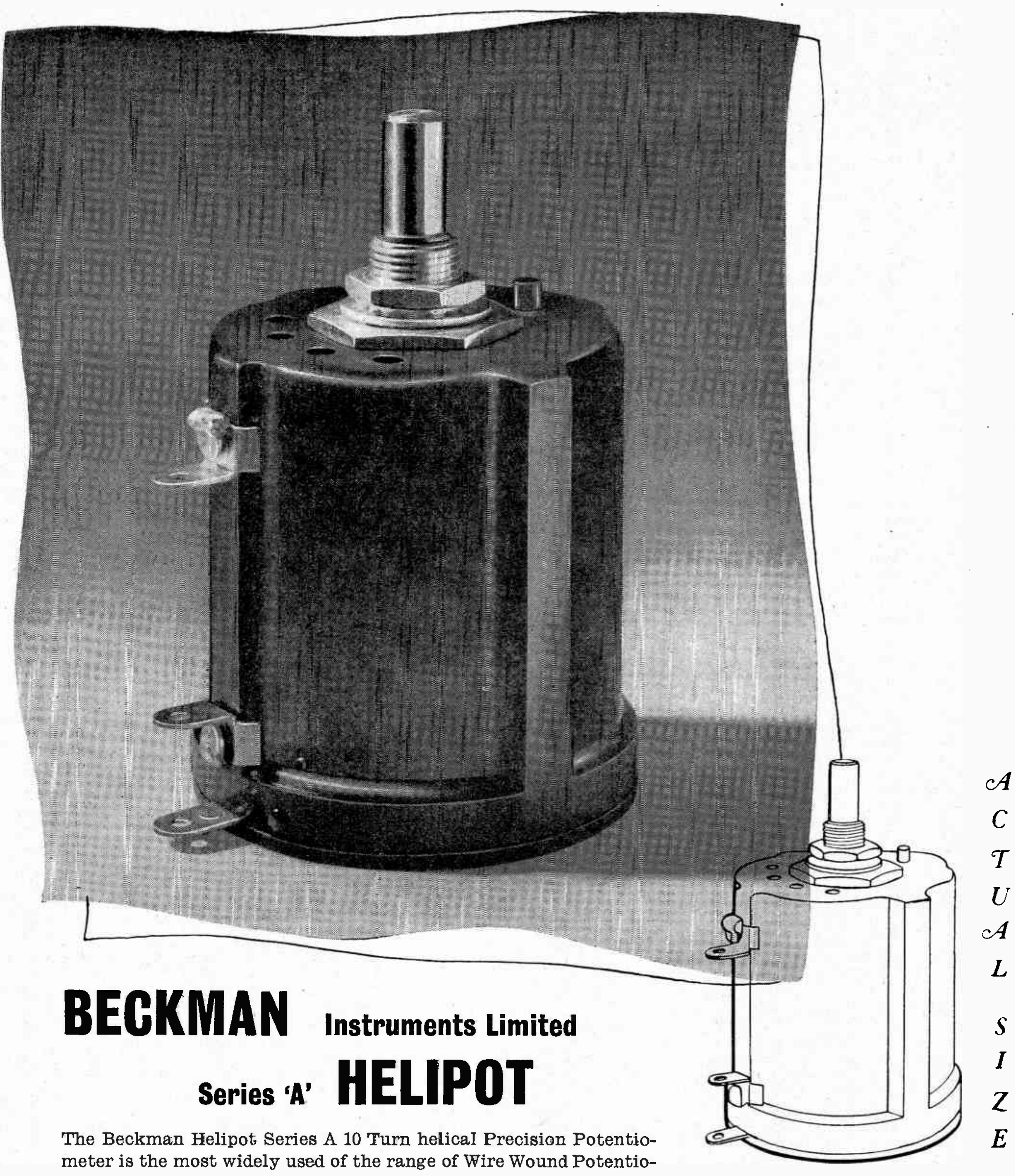
* *rise time to 400mA*

Semiconductors Limited

**Full technical details
and applications assistance
available on request.**

CHENEY MANOR
SWINDON · WILTS
TELEPHONE: SWINDON 6421/7

SC10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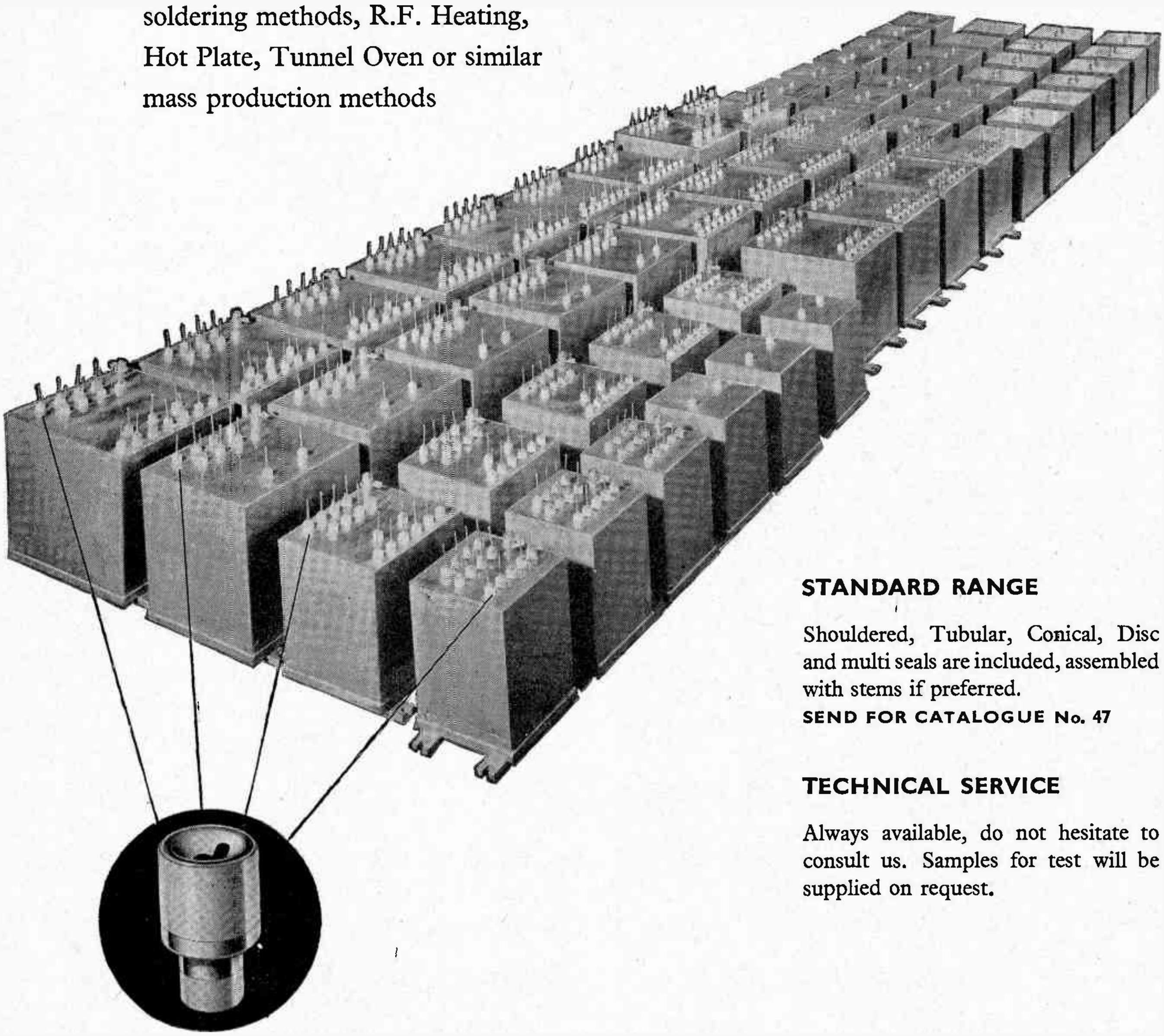

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Shouldered, Tubular, Conical, Disc and multi seals are included, assembled with stems if preferred.

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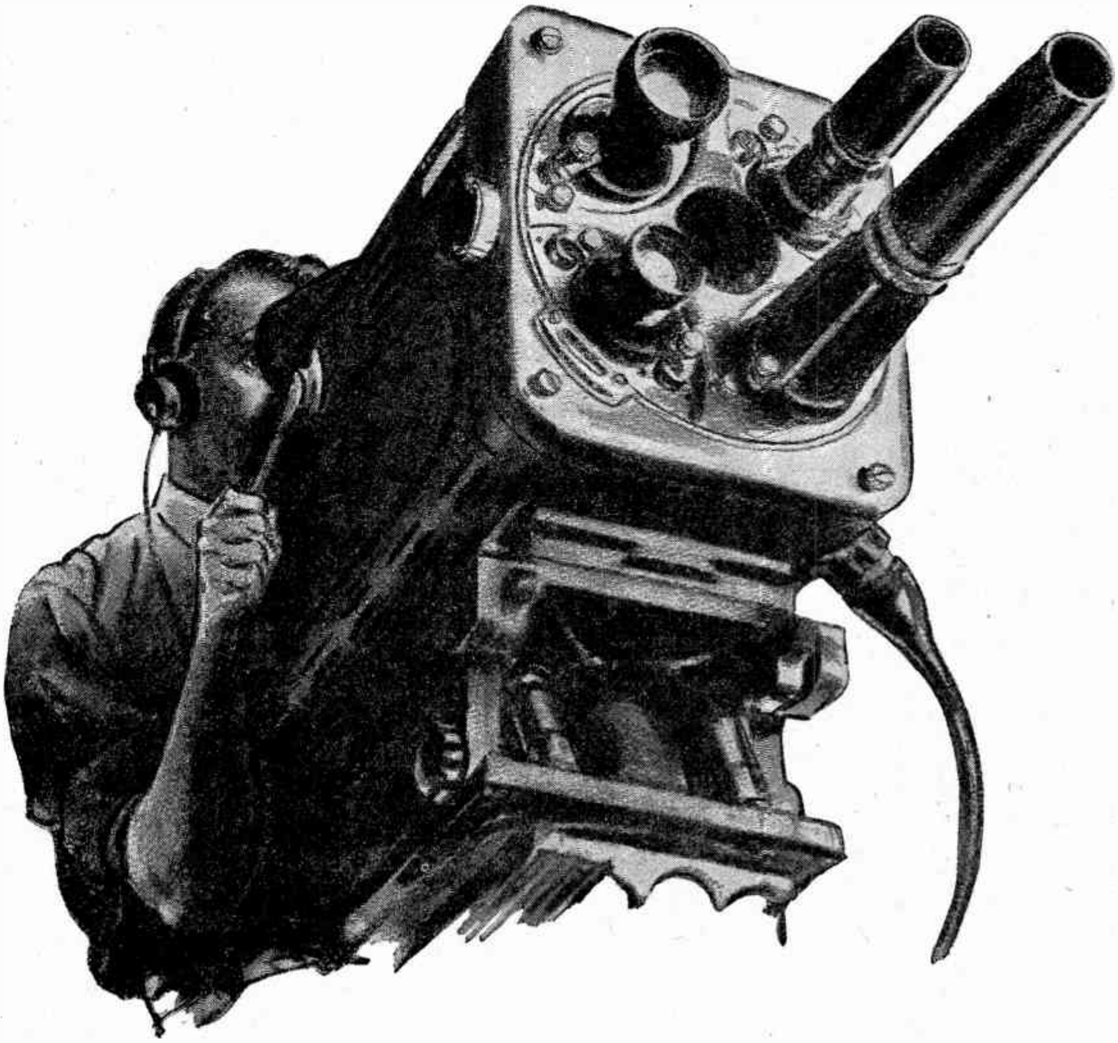
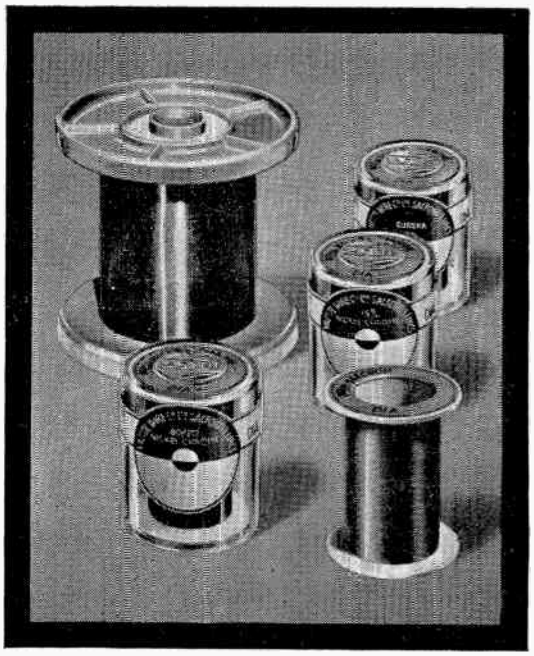
SP100





bare and insulated

resistance wires

The Electronics Industry calls for high precision and exceptional properties in the production of fine and superfine wires. To meet the special requirements, much development work has been carried out in the field of stress relieved and other qualities of the 'VACROM' and 'EUREKA' range of resistance wires. These can be supplied either bare or with standard coverings of cotton, silk, rayon, enamel and glass. Please write for further details and technical information.

'Vacrom' and 'Eureka' are normally supplied in accordance with B.S.S. 115/1954 but can also be supplied to customer's own specification.

VACROM (Regd.)

A nickel chrome alloy in either 80/20 or 15%.

EUREKA (Regd.)

A cupro nickel alloy, with a low temperature co-efficient.

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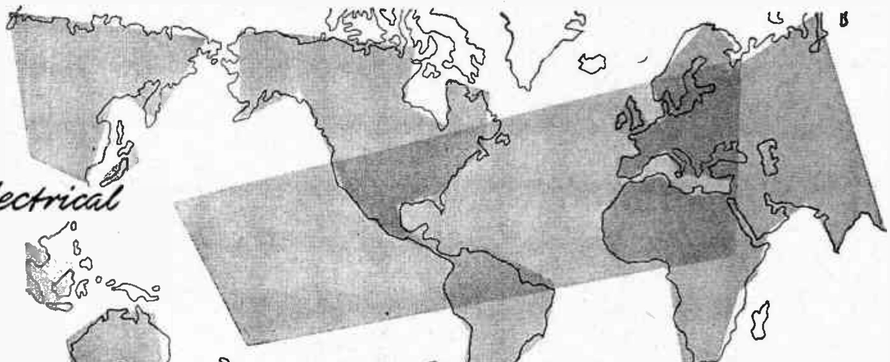
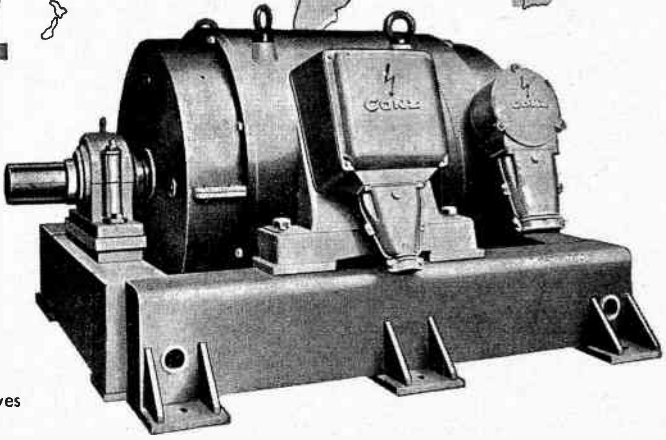
guarantee efficient service in all parts of the globe





Manufacturing Programme

Threephase Squirrelcage Motors up to 2500 kW
Threephase Slipring Motors up to 2500 kW
Threephase Generators up to 3600 kVA
D. C. Motors and Generators up to 2000 kW
Converters and Motor-Generator Sets of all kinds
Ward-Leonard Converters
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World Radio History

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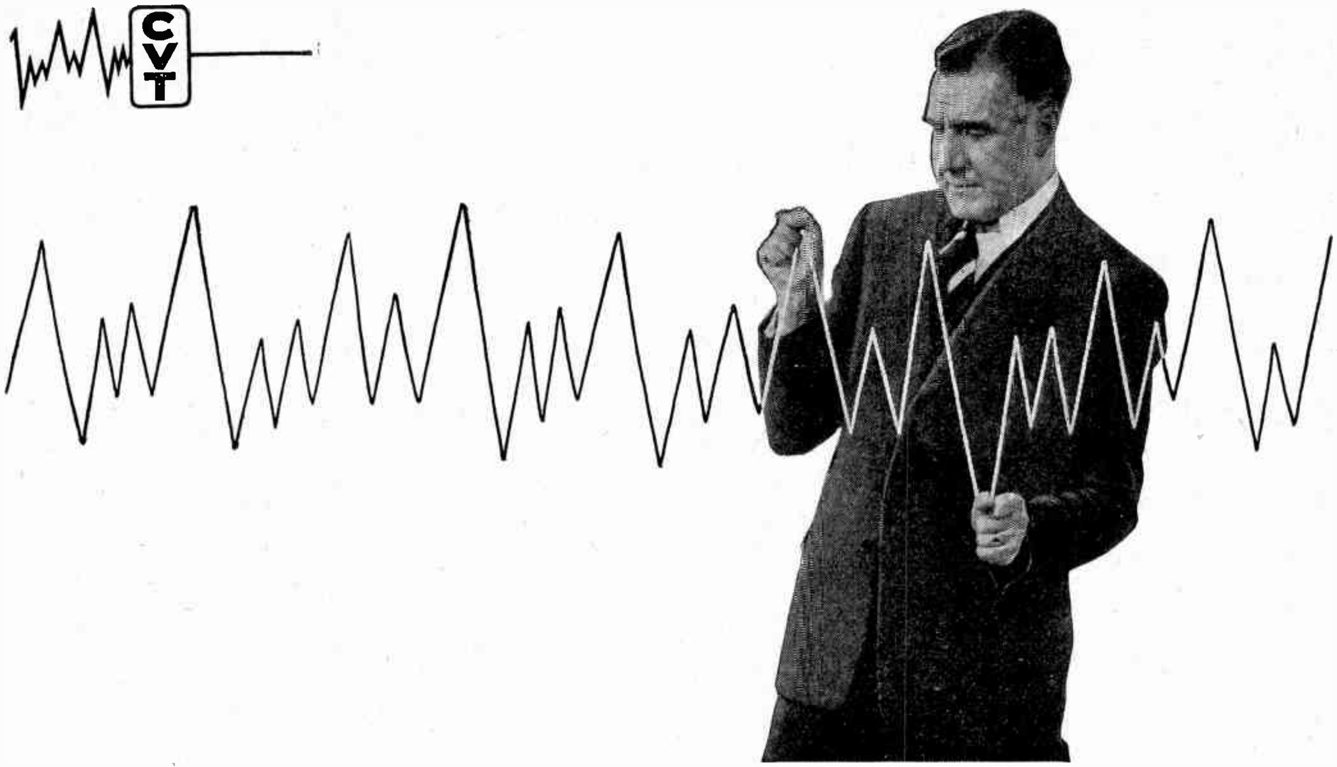

Why struggle with Mains Voltage Fluctuation?

If you have <u>*any*</u> *problem involving a.c. voltage regulation, the solution is to call in 'Advance' — the C.V.T. specialists.*

Investigation of your problem may prove that a standard type Constant Voltage Transformer will meet the case; or maybe, a special design is called for. In either event, the wealth of experience gained by 'Advance' over many years in probing every aspect of mains stabilization provides the surest, quickest, and certainly the most economical, solution to your difficulties.

Advance

CONSTANT VOLTAGE TRANSFORMERS

"VOLTAGE STABILIZATION"

This 'Advance' Booklet gives authoritative information on 'Advance' Constant Voltage Transformers, and the service available to deal with your particular voltage fluctuation problems. Send for a copy.

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GD 10

THIS IS THE LEWIS VALVE RETAINER

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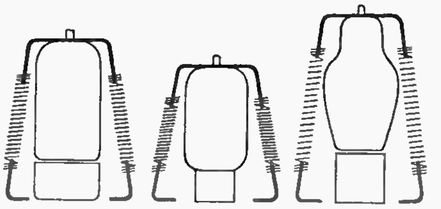

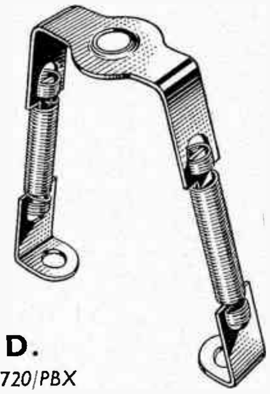

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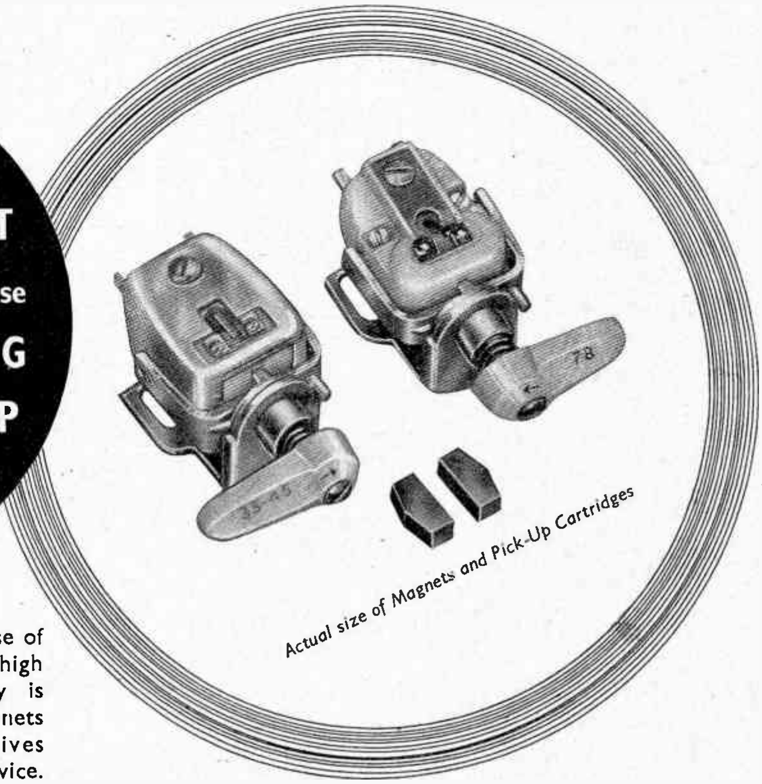

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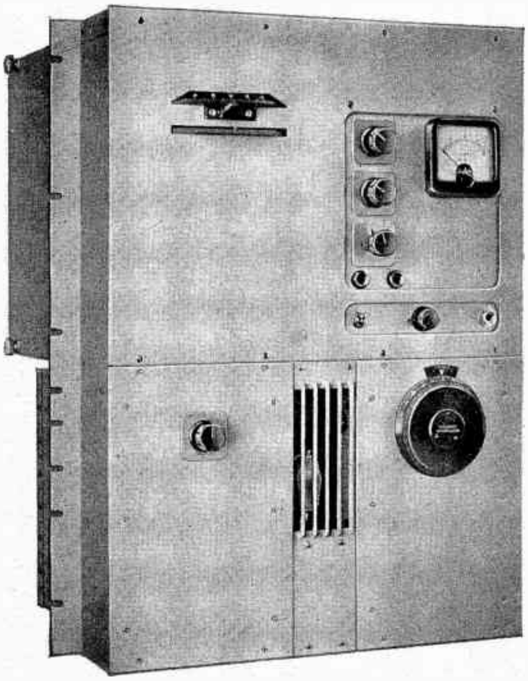

FREQUENCY RANGE 0–16,000 c/s

The frequency calibrated scale of this oscillator has been designed to follow the internationally adopted C.C.I. law:

Linear between 0 and 100 cycles per second.
Logarithmic between 100 c/s and 10 kc/s.
Linear between 10 kc/s and 16 kc/s.

FREQUENCY INTERPOLATING SCALE

Provision is made to vary the frequency of the fixed oscillator ± 50 c/s by means of a separate control calibrated every cycle per sec.

ALTERNATIVE OUTPUTS ARE PROVIDED:

OUTPUT NO. 1

This should be terminated with a resistive load of not less than 10^4 ohms. When thus loaded an output voltage of approximately 300 mV will be obtained with a harmonic content of less than − 60 dB over a frequency range of 30 c/s to 16 kc/s.

OUTPUT NO. 2

This is the main output from the oscillator and is, in effect, a constant voltage source with respect to load, the internal impedance being approximately 1.2 ohms. An adjustable output up to 12 volts with a 15-ohm load is obtainable.

OUTPUT LEVEL

Constant to ± 0.1 dB over full frequency range 20 c/s to 16 kc/s at any given constant load. This feature is common to both Outputs Nos. 1 and 2.

HARMONIC CONTENT

Output No. 1: − 60 dB total harmonic over whole frequency range.

Output No. 2: The harmonic content will vary somewhat according to the load, e.g. with a load resistance of 50 ohms or greater, and the output adjusted to 8 volts, the harmonic content will be less than − 50 dB over a frequency range 100 c/s–10 kc/s.

MAINS HUM (50 c/s and 100 c/s)	− 70 dB for all output voltages above 4 V. 700 μV with output controls at zero.
NOISE	− 70 dB for all output voltages above 4 V. 700 μV with output controls at zero.
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Permanent Magnets

Design Advisory Service

No. 10

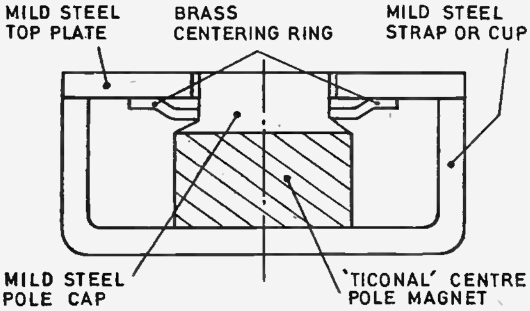
Moving Coil Applications –1

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.

The use of modern permanent magnets has resulted in the efficiency of the magnetic systems for moving coil applications being substantially increased during recent years.

The magnet system employed for domestic loudspeakers, which is one of the best known moving coil applications, can be divided into two general types—'centre pole' and 'ring' designs.

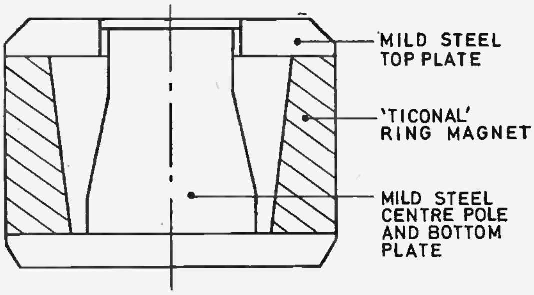
The centre pole design uses a small cylindrical magnet which forms the central pillar of the assembly and is surrounded by a mild steel strap or cup as illustrated. In this design, performance is limited by the size of the magnet which can be arranged within the assembly. If the diameter of the magnet necessitates a collector cap much larger than the one illustrated, the resultant excessive magnetic fringing or leakage will substantially reduce the efficiency of the assembly.



Section of a Typical Centre Pole Moving Coil Strap or Cup Type Assembly. Approximately ½ scale.

An example of the economical use of 'Ticonal' G, the foremost commercial magnet alloy available, is that a magnet with a diameter of 1.14" and height .765", weighing only 3.16 oz., will give a total gap flux of 31,200 maxwells. This is equivalent to an average flux density in a .033" radial air gap of 10,500 gauss, with a mild steel pole diameter of .75" and a top-plate thickness of .187". This represents a magnetic efficiency of 50%.

When higher flux densities in the range of 12,000–17,000 gauss are required, the ring type of assembly is utilised. This type of assembly, which is illustrated, can seldom be designed with an efficiency higher than 30% and is, therefore, not normally used in the design of loudspeakers, where cost is an important factor.



Section of a Typical High Performance Ring Type Moving Coil Assembly. Approximately ½ scale.

For High Fidelity loudspeakers, where the performance is the most important consideration, and when flux densities in the region of 20,000 gauss are required, the ring type of construction is the only possible solution. However, the following table of leakage factors indicates how rapidly leakage losses increase with increasing gap flux density.

1"	diameter	speech	coil	·045"	gap	up	to	12,000 gauss	x 3
1"	,,	,,	,,	·045"	,,	,,	,,	16,000 gauss	x 6
1½"	,,	,,	,,	·050"	,,	,,	,,	16,000 gauss	x 6
2"	,,	,,	,,	·050"	,,	,,	,,	20,000 gauss	x 8

Whilst the designs mentioned above are referred to as loudspeaker systems, they are also applicable to many other applications using the dynamic moving coil principle, such as vibrators, punch card pulse actuators, public address pressure units, moving coil microphones, etc.

If you wish to receive reprints of this advertisement and others in this series write to the address below.

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VOLUME 36 NUMBER 1

ELECTRONIC & RADIO ENGINEER

JANUARY 1959 *incorporating* *WIRELESS ENGINEER*

Mobility

THE literature on semiconductors contains a good deal about the mobility of charge carriers; by this is meant the freedom of movement of electrons or holes in the material. The unit in which mobility is measured has no name, but is usually expressed in terms of cm^2/volt-sec.

This is very puzzling to the thoughtful newcomer to semiconductor theory for, if he is of an enquiring turn of mind, he asks himself what it means. His thoughts naturally follow the lines, "cm^2 is an area; a volt-second is the same as a weber, which is the unit of magnetic flux. Therefore, cm^2/volt-sec is area/magnetic-flux, which is the reciprocal of magnetic flux density; that is, mobility is dimensionally equivalent to $1/B$". Having got so far, he can get no further by following this line of thought.

The difficulty arises over the way in which the unit is expressed. The real meaning of mobility is velocity/electric field strength, which is (cm/sec)/(volts/cm). Put this way, its meaning is obvious; put as cm^2/volt-sec, to which it is formally equivalent, its meaning is not at all clear and starts a misleading train of thought.

The form (cm/sec)/(volts/cm) is rarely used, presumably because it seems a little clumsy. We would like to suggest that a name should be given to the unit. It might be called, for instance, the mobile, for the other meaning of this word is unlikely to confuse anyone! One mobile would be defined as (one metre per second)/(one volt per metre) to make it an m.k.s. unit.

Practical values of mobilities are of the order of 400–4,000 cm^2/volt-sec, or 0·04–0·4 mobile. The 'mobile' is thus of a convenient order of magnitude and it is unnecessary at present to use millimobiles or kilomobiles, although this can always be done if common orders of magnitude change sufficiently in the future.

Hall Effect in Semiconductor Compounds

MODERN APPLICATIONS USING INDIUM ARSENIDE AND INDIUM ANTIMONIDE

By M. J. O. Strutt, D.techn.Sc., D.Eng.(Lond.), F.I.R.E.*

The Hall effect is more pronounced in indium antimonide (InSb) and indium arsenide (InAs) semiconductor compounds than in, for example, the semiconductor compounds of germanium and silicon[1]. Assuming a probe shaped as shown in Fig. 1, with a magnetic field of flux density B (Wb/m^2), perpendicular to the upper flat surface of the probe, we obtain an open-circuit Hall voltage between the points 3 and 4 of

$$V_{ho} = \frac{k_h}{h} F I B \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

In this formula, and referring to Fig. 1, k_h is the Hall coefficient, h is the thickness of the probe, F is a function of the ratio of length l to width b of the probe, and I is the current in amperes through the contacts 1 and 2. The function F is zero when l is zero but attains a value of unity when the ratio l/b is about 2·5. For larger values of this ratio, F is still unity[2].

With the indium antimonide probes, k_h is about $0{\cdot}25 \times 10^{-3}$ m^3/coulomb but, with the indium arsenide probes, k_h is about half this value. We shall show that these values stay constant if the current I, or the magnetic flux, or both, alternate at frequencies up to about 10 Mc/s.

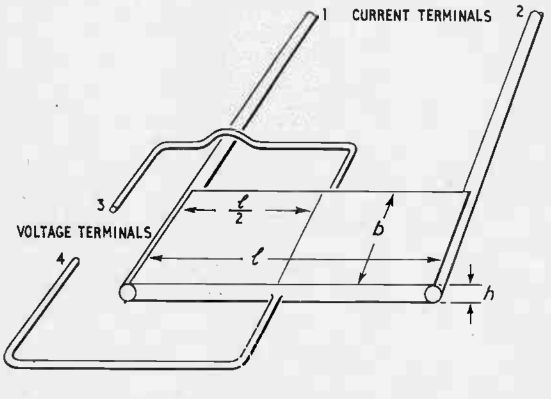

Fig. 1. *Shape of Hall probe. Contacts* 1 *and* 2 *carry the current I through the probe. The Hall voltage V_h is produced across the Hall contacts* 3 *and* 4. *The magnetic field, of flux density B, is perpendicular to the upper surface of the probe*

With indium arsenide, the Hall-voltage V_{ho}, given by Equ. (1), is dependent on temperature as indicated in Fig. 2[3]. The resistance of the Hall probe, as shown in Fig. 1, between the contacts 1 and 2 is indicated by r_i; this resistance is dependent on temperature, as shown in Fig. 3[3]. Similarly, the resistance r_h of the Hall probe of Fig. 1, between the contacts 3 and 4, is also dependent on the temperature, the curve being similar to the one shown in Fig. 3. The resistances r_i and r_h of the Hall probe are also dependent on the magnetic flux density B for constant values of current I and temperature. This is called the 'C.F. Gauss Effect' and is shown in Fig. 4 for Hall probes[1] of different shapes.

Measurements have shown that the Hall contacts, 3 and 4 of Fig. 1, may be displaced to the immediate

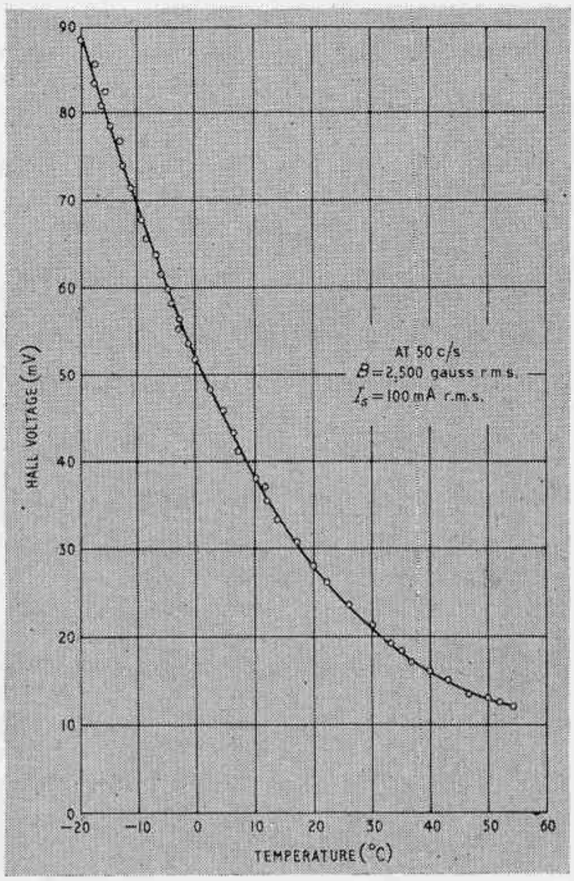

Fig. 2. *Graph showing the variation of the open-circuit Hall voltage V_{ho} with temperature*

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* *Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology.*

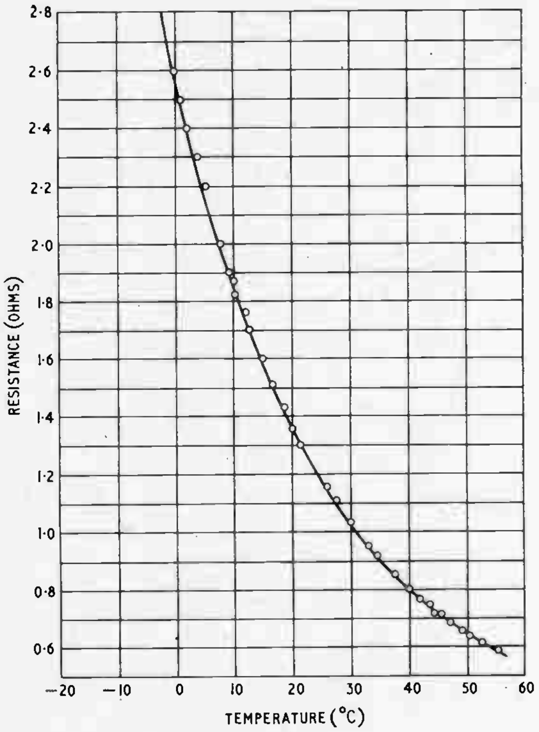


Fig. 3. *Graph showing how the resistance r_i, between the Hall probe contacts* 1 *and* 2 *of Fig.* 1, *is dependent on temperature*

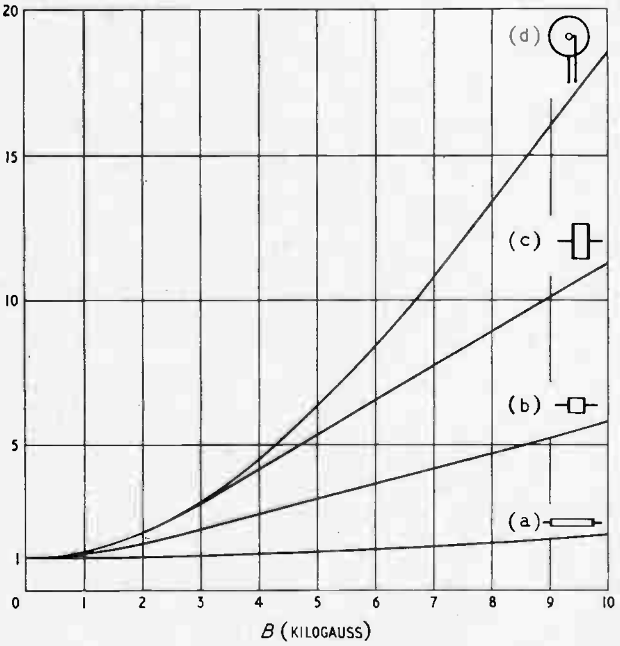


Fig. 4. *Illustrating the 'C. F. Gauss effect' for different Hall probe shapes* (*a*), (*b*), (*c*) *and* (*d*). *Ratio of resistance r_i (of each probe) to its value at zero flux density is plotted against flux density, expressed in kilogauss*

vicinity of the current contacts (see Fig. 5) without affecting the Hall voltage (see Fig. 6)[3].

Wattmeters

It has been shown that the Hall probes of Figs. 1 and 5 may be applied to the design of wattmeters for the measurement of power[4]. Three types of wattmeter will be discussed, viz., 'imaginary' power, real power, and apparent power types.

'Imaginary'-Power Type Wattmeter

A simple arrangement, suitable for measuring the 'imaginary' power, usually referred to as 'reactive volt-amperes', of a single-phase a.c. load circuit Z, is shown in Fig. 7. According to this arrangement, we have

$$\left.\begin{aligned} I &= k_I\, I_{max} \sin(\omega t + \phi) \quad .. \quad .. \quad .. \\ B &= k_V\, V_{max} \sin(\omega t + \pi/2) \quad .. \quad .. \end{aligned}\right\} \quad (2)$$

The coefficient k_I is determined by the shunt R_S and the resistance r, while k_V is determined by the coil and corresponding core of Fig. 7. From Equs. (1) and (2) we obtain

$$V_{ho} = \tfrac{1}{2} \cdot \frac{k_h}{h} F k_I k_V I_{max} V_{max} \{\cos(\phi - \pi/2) - \cos(2\omega t + \phi + \pi/2)\} \quad .. \quad (3)$$

The d.c. part of V_{ho} is proportional to the 'imaginary' power consumed in the impedance Z of Fig. 7.

Real-Power Type Wattmeter

If the current I through the Hall element and the magnetic flux density B are reversed, we obtain a Hall

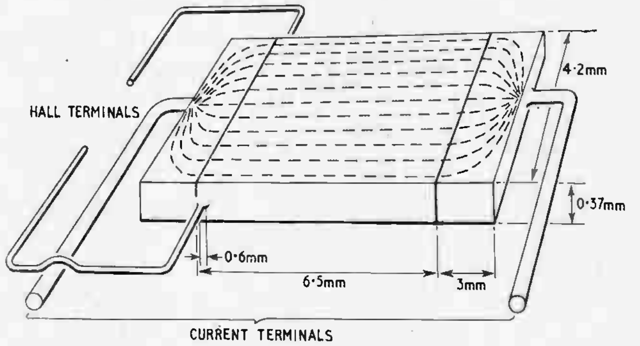
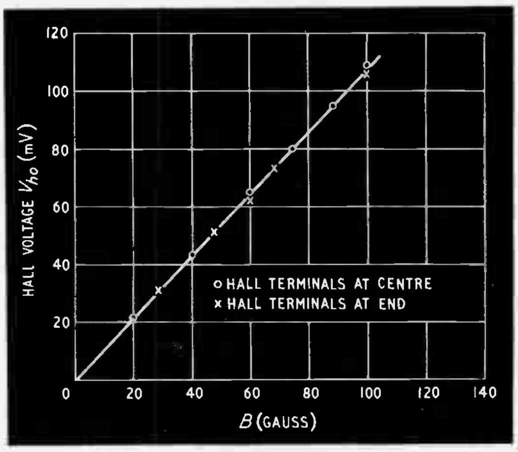


Fig. 5. *Showing Hall probe with its Hall contacts adjacent to one of the current contacts. This arrangement is used in polyphase Hall wattmeters*

Fig. 6. *Linear variation of the open-circuit Hall voltage V_{ho} with magnetic flux density B. The results were obtained with probes shown in Figs.* 1 *and* 5



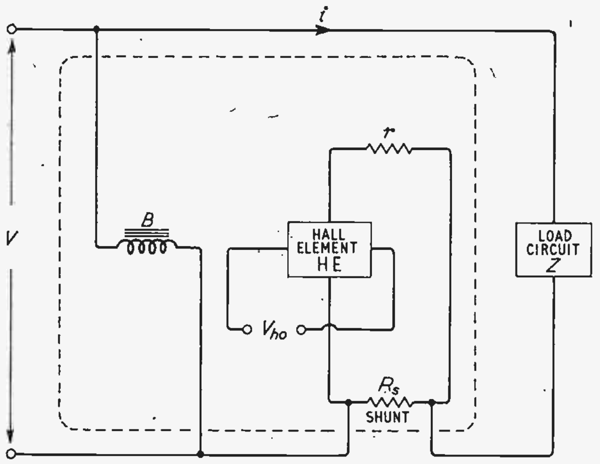


Fig. 7. *Wattmeter arrangement using a Hall element H.E. The current I is supplied via R_s and r, and is proportional to the single-phase current i. The magnetic flux density B is produced by the voltage V between the single phase conductors*

wattmeter suitable for measuring the real power consumed in the impedance of Fig. 7. In this case, the terminals of the shunt R_S must be connected to the coil, causing the flux density B to be produced. The two terminals on the power line must be connected to the current contacts of the Hall probe (see Fig. 1) via the resistance r of suitable value. We then have

$$\left.\begin{aligned} I &= k'_V\, V_{max} \sin \omega t \quad .. \quad .. \quad .. \\ B &= k'_I\, I_{max} \sin (\omega t + \phi) \quad .. \quad .. \quad .. \end{aligned}\right\} \quad (4)$$

Therefore from Equs. (1) and (4) we obtain

$$V_{ho} = \tfrac{1}{2} \frac{k_h}{h} F\, k'_I\, k'_V\, I_{max}\, V_{max}$$

$$\{\cos \phi - \cos (2\omega t + \phi)\} \quad .. \quad .. \quad .. \quad (5)$$

The d.c. component of V_{ho} is proportional to the real power consumed by the impedance Z. This method of connection, for measuring the real power, is shown in Fig. 9 pertaining to a three-phase system[3].

Apparent-Power Type Wattmeter

It is also possible to build a wattmeter for measuring the apparent power (i.e., the product of peak current and peak voltage) consumed in the impedance Z of Fig. 7[3]. An appropriate arrangement is shown in Fig. 8, which may be considered as self-explanatory.

Polyphase Wattmeters

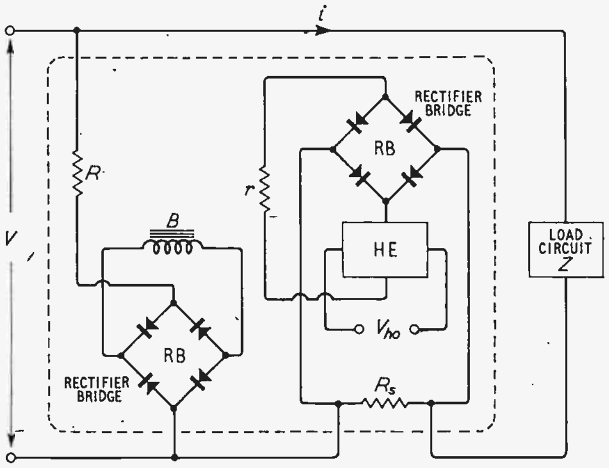
If we try to apply the arrangement of Fig. 7, or its inverse (for measuring the real power) to a polyphase power system, we run into trouble. One difficulty is that considerable a.c. voltages develop across the Hall contacts of the Hall probes. As a result of these voltages considerable currents would be produced, destroying the Hall probes. This is avoided quite simply by the method shown in Fig. 9. Here, three Hall elements (H.E.) are used with their contacts arranged as in Fig. 5. In this way, the a.c. voltages across the Hall contacts are reduced almost to zero. The d.c. components of the Hall voltages of the three elements are in series, and their sum is proportional to the real three-phase power consumed in the three load impedances Z_a, Z_b and Z_c. Another method, which can be used to overcome the difficulty is to use suitable transformers. These transformers, which must be devoid of spurious phase angles, supply the current I and the flux density B of the Hall elements. A Hall wattmeter of this kind is shown in Fig. 9 and, on test, its performance was found to be satisfactory[3].

Temperature Compensation

The temperature dependence of the open-circuit Hall voltage (see Fig. 2) is detrimental to the performance of Hall wattmeters unless it is compensated. The open-circuit Hall voltage may be made nearly independent of temperature at a constant flux density, by using the temperature dependence of r_i (see Fig. 3) as a means of compensation. The current I through the Hall element must increase with rising temperature in such a way that V_{ho} is independent of temperature. An appropriate circuit is shown in Fig. 10. The current i_R is divided into a current i_s through the Hall element H.E. and a current i_{rt} through the auxiliary resistance r_t. This resistance r_t is, however, slightly dependent on temperature, according to a pre-calculated curve. In a practical case, the resistance r_t was composed of manganin wire of 0·0875 ohm (see curve 3 of Fig. 11). By using suitable series combinations for the resistance r_t of Fig. 10, it is possible to obtain a Hall-voltage curve which is independent of temperature over a considerable temperature range to within 0·5% (see curve 5 of Fig. 11)[3].

Obviously, this temperature compensation at a definite flux density B is upset as soon as the flux density is altered. Such an alteration, according to the Gauss effect of Fig. 4, alters the internal resistance r_i, of Fig. 10, between the current contacts of the Hall element H.E. The compensation depends on the particular value of r_i at room temperature, corresponding to the flux density under consideration. It is possible, however, to avoid upsetting the temperature compensa-

Fig. 8. *Circuit of wattmeter suitable for measuring the apparent power (i.e., product of peak current and peak voltage) dissipated in the load-circuit impedance Z. The rectifier bridges supply the flux density B and the current I*



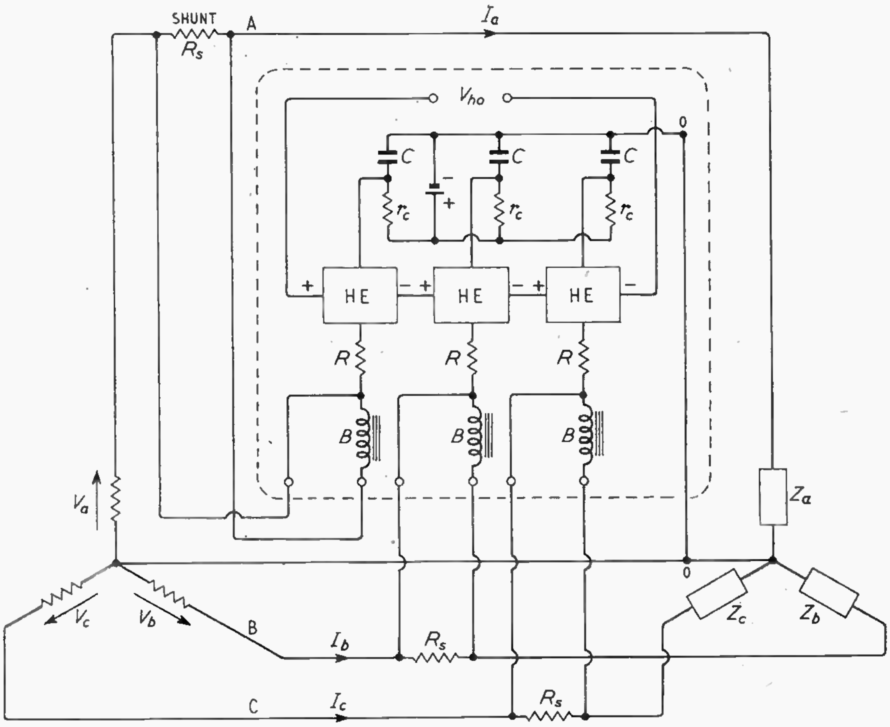


Fig. 9. *Circuit of wattmeter suitable for measuring the real power of a three-phase system consumed in impedances Z_a, Z_b and Z_c. One of the current contacts of each Hall element is connected to earth via a large capacitor C. If electrolytic capacitors are used, the three resistances r_c must be included in order to apply polarizing voltages to their terminals. Resistances R and capacitors C must satisfy the relationship $R \gg 1/\omega C$ or $\omega RC \gg 1$, to ensure that the a.c. through the Hall elements is in phase with the corresponding phase voltages of the three-phase system*

tion if a Hall probe (see Fig. 1) is used whose length l is considerably greater than its width b; i.e., the ratio l/b of the Hall probe is large. Fig. 4 shows that the variation of r_t with flux density is small for a large l/b ratio. We may use a supplementary bias d.c. flux density of fixed value (e.g., 2,500 gauss), on which is superimposed the flux density caused by the current i of Fig. 10. As long as the superimposed alternating flux density is smaller than the fixed d.c. flux density (e.g., as produced by a suitable permanent magnet), the resistance r_t of Fig. 10 will be very nearly constant at constant temperature and current i_s at varying flux densities. The remaining relatively small variations of r_t with flux density will, in this case, have no influence on the d.c. component of the Hall voltage. Hence, in this way, the temperature compensation may be applied satisfactorily.

The internal resistance r_h between the Hall contacts varies with temperature in a way similar to that of the resistance between the current contacts of the Hall element in Fig. 2. This temperature variation of r_h prohibits the full use of the available d.c. power at the Hall contacts. Assuming the external load on these contacts to be r, we obtain a Hall voltage of

$$V_h = V_{ho}\frac{r}{r+r_h} = \frac{V_{ho}}{r_h/r+1} \quad \cdots \quad \cdots \quad \cdots \quad (6)$$

The power P_h consumed in the Hall circuit load resistance r is given by

$$P_h = \frac{P_{ho}}{\frac{1}{2}+\frac{1}{4}\cdot\frac{r_h}{r}+\frac{1}{4}\frac{r}{r_h}} \quad \cdots \quad \cdots \quad \cdots \quad (7)$$

where P_{ho} is the available power at the Hall contacts and equals $V_{ho}^2/4r_h$. Obviously, if $V_h \approx V_{ho}$, then $r \gg r_h$. In this case, the load power P_h is only a very small fraction of the available Hall power P_{ho}. These conditions must be adhered to in order to avoid upsetting the temperature compensation of V_{ho} by the temperature dependence of r_h. The result is that only very small effective load powers P_h, of the order of microwatts, may be obtained. Much larger powers, of the order of milliwatts at least, are needed in order to use the Hall voltage for the regulation of power systems. Powers, in the milliwatt range, may be obtained by using suitable power amplifiers. Employing an appropriate magnetic amplifier, a linear power amplification of about 2,000 times has been obtained[3].

Hall Effect Oscillators

A simplified picture of a Hall effect oscillator is shown in Fig. 12[5]. The open-circuit Hall voltage V_q,

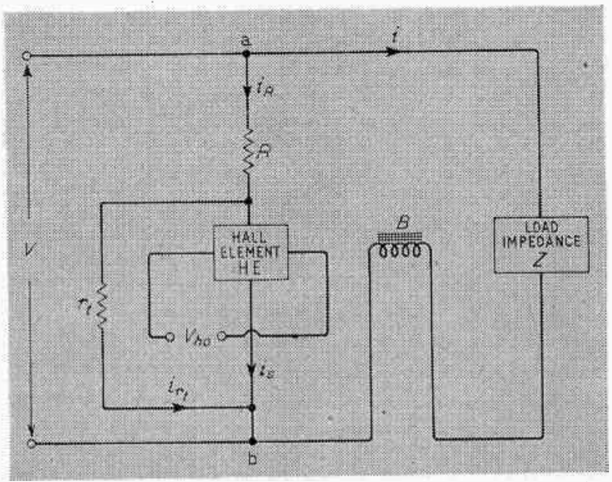


Fig. 10. *Practical circuit for the compensation of the temperature dependence of V_{ho}. A suitable shunt resistance r_t is connected in parallel with the Hall element H.E*

between the Hall contacts 5, 6 and 3, 4 of Fig. 12, is given by Equ. (1). If the remanent flux density of the ferrite core is B_m, then the flux density in the airgap of width g is approximately

$$B = \frac{\mu_0}{g} i_h N + B_m \quad .. \quad .. \quad .. \quad .. \quad (8)$$

where the relative permeability μ_r of the core material is assumed to be large compared with unity, and μ_0 is the permeability of free space. The current i_h is given by

$$i_h = \frac{V_q}{r + r_h} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Writing $V_{ho} = V_q$, and substituting Equs (8) and (9) into Equ. (1), gives

$$V_q = \frac{k_h}{h} F B_m I \, (1 - KI)^{-1} \quad .. \quad .. \quad .. \quad (10)$$

where the feedback factor K is given by

$$K = \frac{\mu_0 \, k_n}{g_h} \cdot F \frac{N}{r + r_h} \quad .. \quad .. \quad .. \quad .. \quad (11)$$

It can be seen that if the product KI is unity, a Hall voltage V_q may be obtained when the remanent flux density B_m is zero. This means that, in this case, oscillations may commence. In a practical case, the value of K was 1·75 (ampere)$^{-1}$.

It may be shown that, if $KI > 1$, oscillations build up, the process continuing until KI reduces to unity. The reduction of K produces several results, such as an increase of r and r_h with rising temperature, additional losses in the core, and mismatch. The oscillations build up in accordance with the following expression:

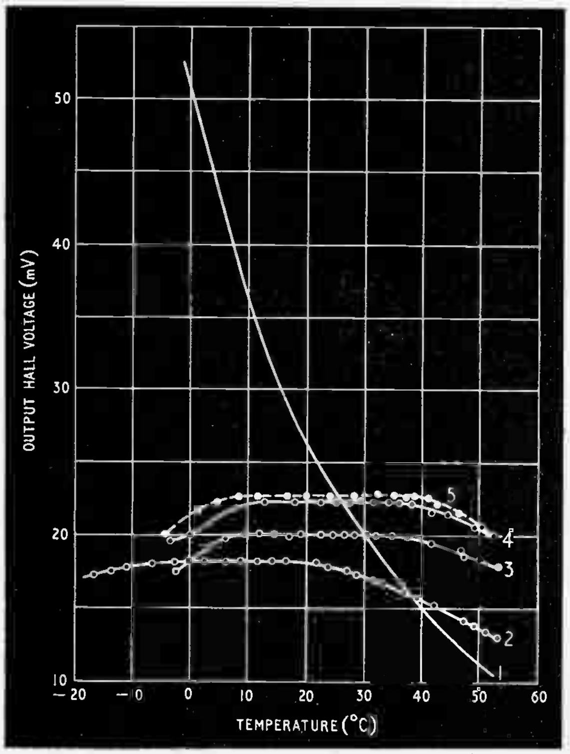
$$V_q = V_m \cos(\omega t + \phi) \exp\left[-\frac{t}{2T}(1 - KI)\right] \quad .. \quad (12)$$

In this equation, $T = L/(r + r_h)$; i.e., the time constant of the circuit. The square of the angular frequency is given by

$$\omega^2 = \frac{1}{LC} - \left(\frac{1 - KI}{2T}\right) \quad .. \quad .. \quad .. \quad .. \quad (13)$$

When $KI = 1$, we obtain $\omega^2 = 1/LC$; this corresponds to the steady state.

Fig. 11. *Temperature-compensated curves of the open-circuit Hall voltage* V_{ho} *for a flux density B of* 2,500 *gauss; Curve* 1, *uncompensated; curve* 2, i_R *(of Fig.* 10) = 0·5 *A*, r_t *(manganin wire)* = 0·157 Ω; *curve* 3, i_R = 0·8 *A*, r_t *(manganin wire)* = 0·0875 Ω; *curve* 4, i_R = 0·7 *A*, r_t *(nickel wire)* = 0·116 Ω at 20 °*C*; *curve* 5, i_R = 0·7 *A*, r_t *consisting of a nickel wire of* 0·103 Ω *at* 20 °*C in series with a manganin wire of* 0·029 Ω



In Fig. 13 some experimental build-up curves, corresponding to Equ. (12), are shown[5]. The calculated values for the exponents are in satisfactory agreement with the experimental values of Fig. 13. The decline in the steady values of oscillatory voltages on reaching their peak value is due to the heating up of the circuit and Hall probe. The values of V_{ho} were obtained at the contacts 3 and 4 of Fig. 12.

The saturation flux densities of the ferrite core under consideration determine the peak levels of the oscillatory voltages shown in Fig. 13. This may be checked from the curves of Fig. 14, which show how saturation occurs at a definite value of supply current I for each curve. Disconnecting the coil N of Fig. 12 from its Hall contacts, the magnetic hysteresis loop of the core was displayed on an oscilloscope screen. One pair of oscilloscope deflector plates was connected to the contacts 3 and 4 of the Hall probe (see Fig. 12), and, an a.c. voltage, proportional to the a.c. through the coil N of Fig. 12, was applied to the other pair of plates. On using one of the frequencies of Fig. 14, the hysteresis loop showed saturation at exactly the V_{ho} value corresponding to Fig. 14.

Experimental investigations were carried out to test the efficiency of the Hall-effect oscillator[5]. Operating the oscillator at 20 °C and 60 c/s, with a supply current I of 1 A, and connecting a 3-ohm resistive load across the contacts 3 and 4 of Fig. 12, the optimum output power at these contacts was obtained. The Hall oscillator was self-oscillating, as indicated in Fig. 14. The power dissipated in the 3-ohm resistive load was 12 mW. The supply power, corresponding to the supply current I, was 810 mW. The Hall probe, between the contacts 1 and 2 of Fig. 12, dissipated a power of 756 mW. Hence, only 54 mW (the difference of the above values) was available for the oscillations. The output power being 12 mW, the efficiency was 12/54 = 0·22. The power dissipated in the Hall probe, in this case 756 mW, may be compared with the cathode-heating power of a valve which is usually not taken into account when considering the efficiency of a valve oscillator.

The fact that the efficiency of oscillation (see Fig. 14) falls off at the higher frequencies is due to the rising losses in the ferrite core (this core being unsuitable for use at high frequencies), and losses incurred in the coil and Hall probe, resulting in a mismatch. It is *not* due to a decline of the Hall coefficient k_h of Equ. (1).

The frequency dependence of k_h was investigated by measuring the Gauss effect (see Fig. 4) at different

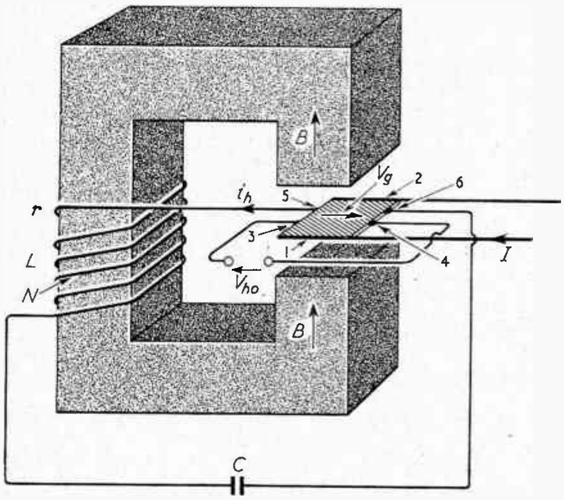


Fig. 12. *Hall probe with current contacts* 1 *and* 2 *through which the d.c. supply current I flows. Hall contacts* 5 *and* 6 *are connected to the feedback coil of N turns, resistance r and inductance L. The Hall current i_h flows through the coil. The open-circuit Hall voltage V_{ho} is produced across contacts* 3 *and* 4. *The capacitor C is connected in series with the feedback coil. The ferrite core causes a magnetic flux B to be produced, which is perpendicular to the current paths of I and i_h*

frequencies[6]. Some relevant values are shown in Figs 15 and 16. The probe of Fig. 15 was of length $l = 2{\cdot}7$ mm, width $b = 6{\cdot}1$ mm and thickness $h = 1{\cdot}6$ mm. Its conductivity (of p-type) was 150 mho/cm. It is seen that a slight decline of the Gauss effect and hence of the Hall coefficient k_h of Equ. (1) is obtained at 10 Mc/s, compared with its value at zero frequency. The probe, used in the measurements of Fig. 16, was of length $l = 3$ mm, of width $b = 5{\cdot}8$ mm and of thickness $h = 0{\cdot}15$ mm. Its conductivity (InSb of n-type) was 90 mhos/cm. As a result of this (see Fig. 16) the Gauss effect is considerably smaller, and hence the Hall coefficient k_h of Equ. (1); this is evident on comparing the values at 300 Mc/s and at 600 Mc/s with those at zero frequency.

Hall Effect Flux Density Meter

Flux density meters, using germanium crystals, have been in use for many years. With the semiconductor compounds, a more sensitive meter may be constructed[7]. An arrangement will be described which has been designed to attain the highest possible sensitivity.

The Hall probe used was of width $b = 2$ mm, of length $l = 4$ mm, and of thickness $h = 0{\cdot}2$ mm. An indium arsenide probe was used in preference to an indium antimonide probe because the Hall coefficient k_h of the former is far less dependent on temperature, at room temperatures. Its Hall coefficient k_h was found to be $0{\cdot}12 \times 10^{-3}$ m³/coulomb. The function F of Equ. (1) being about 0·83, the open-circuit Hall voltage becomes

$$V_{ho} = 0{\cdot}5\,B\,I \quad \ldots\ldots \quad (14)$$

A steady temperature of about 100 °C was produced with a current I of 1·2 A.

Consideration was then given to the use of probes of less thickness h. According to Equ. (1), the open-circuit Hall voltage V_{ho} increases if h is reduced at constant values of B and I. However, if h is reduced, leaving the width b constant, the maximum current I must be reduced to prevent the heat dissipation in the probe from causing an undue rise in temperature. It must be borne in mind that the probe's temperature must remain well below 150 °C in order to avoid disintegration of the material. Considering different possibilities, such as the reduction of I and an increase of b with or without a corresponding increase of l, it was found that only a slight increase of V_{ho}, if any, at a given value of flux density B, can be achieved by reducing the thickness h.

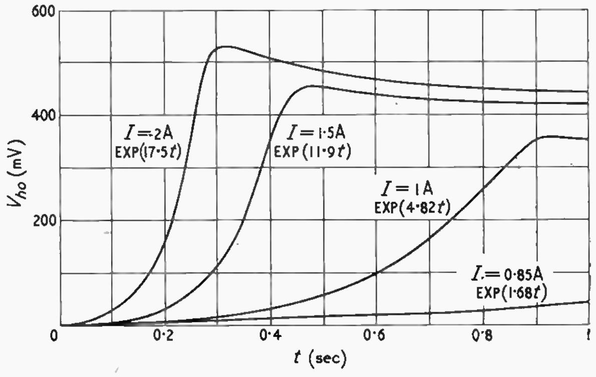
A block diagram of the experimental set-up is shown in Fig. 17. As seen from the figure, a 10-kc/s transistor oscillator[8] is connected across the Hall probe followed by a suitable transistor power output stage. The reasons for using a 10-kc/s a.c. source are as follows. It is easier to build a stable transistor amplifier operating over a range 9 kc/s to 11 kc/s than over the 0 to 1 kc/s range. Amplifiers, either of the valve or transistor type, always show high inherent flicker noise at frequencies f approaching zero. The tapering off of this flicker noise is approximately proportional to $1/f$, and is less pronounced at 10 kc/s. Here, the noise is mainly of the shot and Nyquist type[9,10].

Noise measurements were carried out at the Hall terminals of the Hall probe h.p. of Fig. 17. Some of the results obtained are shown in Fig. 18. Here, the equivalent noise resistance R_{eq} is shown as dependent on the current I (horizontal scale) through the Hall probe. Its dependence on I may be expressed approximately by the equation

$$\frac{R_{eq}\,(I)}{R_{eq}\,(0)} = 1 + 1{\cdot}14\,I^2 \quad \ldots\ldots \quad (15)$$

While this curve of Fig. 18 corresponds to a frequency of 15 kc/s, similar curves taken at lower and higher frequencies showed a tendency for the ratio to drop with rising frequency at a definite value of I. It can be seen from Fig. 18 that the noise at a current of 1·2 A is about two-and-a-half times the Nyquist noise at 15 kc/s. At 10 kc/s, this figure rises to somewhat less than three times the Nyquist noise. The transistor amplifier has a

Fig. 13. *Graphs, drawn from oscilloscope screen pictures for different values of the supply current I, show the variation of the effective value of V_{ho} with time t (see Fig.* 12). *The exponents correspond to Equ.* (12). *The experimental values are in satisfactory agreement with the calculated values derived from Equ.* (12) *for the set-up used*



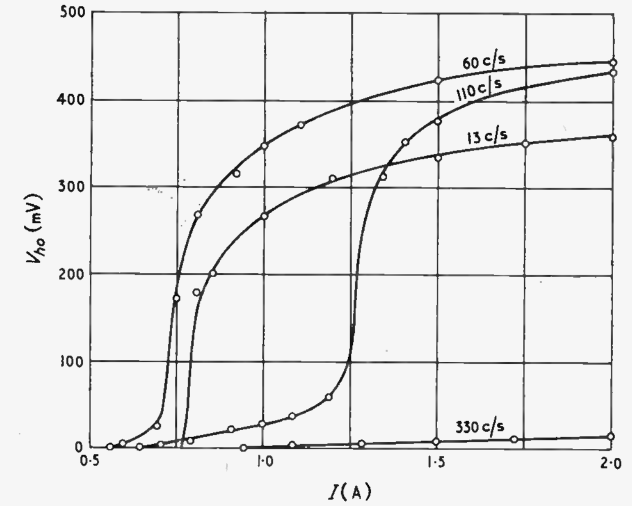


Fig. 14. *The curves, produced at frequencies of* 13, 60, 110 *and* 330 *c/s, show the variation of the effective Hall voltage* V_{ho} *with the d.c. supply current I*

measured[9] noise figure of 2. These figures enable us to calculate the total equivalent effective noise voltage at the Hall terminals of the probe, taking into account the noise of the amplifier and of the probe itself. For the latter, we assume that the equivalent noise resistance is three times the actual resistance between the Hall terminals. The equivalent noise resistance due to the amplifier noise is about twice the actual resistance at the Hall terminals. Hence, the total noise resistance at the Hall terminals of the probe is about five times its effective resistance. The latter value, at a current of 1·2 A, is about 1 ohm. The total mean square effective noise voltage at the Hall terminals is given by the equation

$$V^2_{rms} = 4\,k\,T\,R_{eq\,.\,total}\,\Delta f \qquad \ldots \quad \ldots \quad \ldots (16)$$

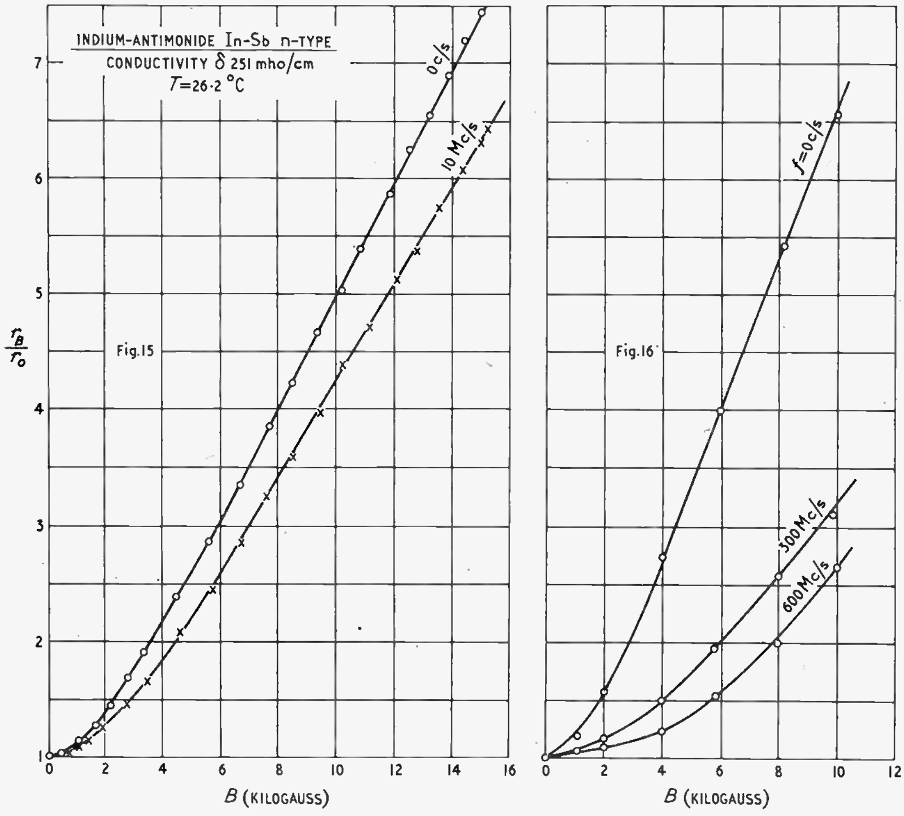
Here, k is Boltzmann's constant ($1{\cdot}38 \times 10^{-23}$ joule/°K), T the temperature in °K, and Δf the effective bandwidth which is 2 kc/s in our case. We obtain, at a temperature T of 393 °K, a square root mean square value of the noise voltage of approximately 0·015 μV.

Assuming that the minimum Hall voltage is equal to the noise voltage, we obtain $V_h = 0{\cdot}015\ \mu$V. If the Hall terminals of the probe are matched to the input of the amplifier, then under optimum load conditions V_h will be half the open-circuit Hall voltage; i.e., 0·03 μV. The corresponding flux density B is obtained from Equ. (14), and is given by

$$B = \frac{2\,V_{ho}}{I} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (17)$$

At a current I of 1·2 A, we obtain an effective value of $B \approx 5 \times 10^{-8}$ volt sec/m² or 5×10^{-4} gauss.

It should be borne in mind that the above figure corresponds to a bandwidth at the detector output (see Fig. 17) of 1 kc/s. Therefore, variations of flux density up to 1 kc/s may be detected. In many cases, only much slower variations of flux density are to be measured. If, for instance, the output of the detector is connected to an indicator, registering up to 0·5 c/s, the corresponding minimum effective value of B would be $\sqrt{2{,}000}$ times smaller, or about 10^{-5} gauss. Furthermore, if suitable pole pieces of mumetal are used, this minimum effective value of B is still about 400 times smaller, or about



Figs. 15 *and* 16. *Illustrating the Gauss effect of indium antimonide Hall probes at different frequencies. The ratio of resistance* r_B *of the Hall probe at a fixed transverse flux density B to the resistance* r_0 *at zero flux density is plotted against the transverse flux density B.* Fig. 15 (*left*) *shows two curves derived at* 0 *c/s and* 10 *Mc/s.* Fig. 16 (*right*) *shows curves at* 0 *c/s,* 300 *Mc/s and* 600 *Mc/s which indicate a significant decline in the Gauss effect*

2×10^{-8} gauss. Comparing this value with the best ones published so far, it is seen to be about 250 times smaller[7].

The use of a transistor oscillator and transistor-detector results in a very compact and portable arrangement, adaptable to many applications.

Hall Effect Mixer Stage

Mixer stages for superheterodyne receivers may be constructed using a Hall probe. The set-up is shown in Fig. 19. The aerial signal causes a current $I_{max} \sin \omega_i t$ to flow through the Hall probe 6. The local oscillator causes a transverse magnetic flux density $B_{max} \sin \omega_{osc} t$ at the Hall probe 6. The resulting Hall voltage, which constitutes the intermediate frequency signal, is given by

$$V_{ho} = \frac{k_h}{h} F \cdot \tfrac{1}{2} I_{max} B_{max} \cos(\omega_{osc} - \omega_i)t \quad .. (18)$$

This is filtered, amplified and detected at stages 11 and 12.

This mixer stage behaves differently from the ones which use mixer valves, transistors or diodes, in that no distortion is involved in the mixing process. The Hall voltage is proportional to the first power of I_{max} and does not contain any higher power components. The Hall voltage does contain a component of angular frequency $(\omega_{osc} + \omega_i)$, but this is filtered out by the intermediate frequency stage 11, which is tuned to $(\omega_{osc} - \omega_i)$. The ferrite core, as a result of its non-linear characteristic, causes some harmonic components of the flux density B to be produced. These, in turn, cause Hall-voltage components of angular frequencies $(n\omega_{osc} \pm \omega_i)$, where $n = 2, 3, 4$, etc. However, these Hall-voltage components are also filtered out by stage 11.

The behaviour of the Hall mixer stage is determined mainly by its gain G and noise figure N. In the case of indium antimonide, the Hall coefficient k_h of Equ. (1) is about $0{\cdot}25 \times 10^{-3}$ m^3/coulomb. In a practical arrangement, h had a value of 0·1 mm, and therefore, if F of Equ. (1) is approximately unity,

$$V_{ho} = 2{\cdot}5\, BI \quad .. \quad .. \quad .. \quad .. (19)$$

The resistance between the Hall contacts, at an effective temperature T_h, is r_h. The available Hall power of angular frequency $(\omega_{osc} - \omega_i)$ at the Hall contacts of the probe is then given by

$$P_h = \frac{\frac{1}{2}\left(\frac{1}{2} \cdot 2{\cdot}5\, B_{max} I_{max}\right)^2}{4 r_h} \quad .. \quad .. \quad .. (20)$$

The available input power of angular frequency ω_i is

$$P_i = \frac{\frac{1}{2} I^2_{max}\, r_0^2}{4 r_0} = \frac{1}{8} \cdot I^2_{max} r_0 \quad .. \quad .. (21)$$

where r_0 is the effective internal resistance of the aerial circuit between the tap 14 on the coil of the resonant aerial circuit and earth (see Fig. 19). The temperature of r_0 is assumed to be the room temperature T_0. The gain of the mixer stage is

$$G = \frac{P_h}{P_i} = \frac{(2{\cdot}5\, B_{max})^2}{4 r_h\, r_0} \quad .. \quad .. \quad .. \quad .. (22)$$

In a practical set-up, the value of B_{max} was 3,000 gauss or 0·3 volt-sec/m^2. In this case, the oscillator connected to the contacts 1 and 2 of Fig. 19 had to supply a power of some 10 mW. Furthermore, r_0 and r_h were both about 1 ohm. Hence, a gain G of about 1/7 or $-$ 8·5 is obtained.

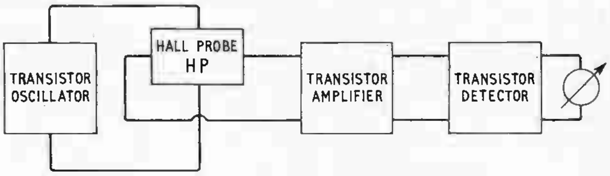


Fig. 17. *The transistor oscillator supplies the Hall probe with a current (of effective value I) at a frequency of* 10 *Mc/s. The amplified output Hall voltage (from the* 9–11-*kc/s transistor amplifier) after detection gives signals of frequencies between* 0 *and* 1,000 *c/s*

The noise figure N may be evaluated as follows. The available input noise power is $kT_0\Delta f$, where k is Boltzmann's constant and Δf is the effective bandwidth under consideration. This gives an equivalent output noise power of $GkT_0\Delta f$. As the Hall resistance r_h causes an available output noise power $kT_h\Delta f$, the total available output noise power becomes $GkT_0\Delta f + kT_h\Delta f$. The noise figure N, obtained by dividing the available output noise power by $GkT_0\Delta f$, is

$$N = 1 + \frac{T_h}{GT_0} \quad .. \quad .. \quad .. \quad .. \quad .. (23)$$

If $T_h = T_0$ and $G = 1/7$, then $N = 8$. If the intermediate frequency stage 11 of Fig. 19 is assumed to have a noise figure N_1, then the overall noise figure will be

$$N_{total} = N + \frac{N_1 - 1}{G} \quad .. \quad .. \quad .. \quad .. (24)$$

Assuming $N_1 = 2$, we obtain $N_{total} \approx 15$. This value compares favourably with the noise figures of known mixer stages in the radio-frequency and high-frequency ranges. Owing to a considerable drop in the Hall voltage above 30 Mc/s, the Hall mixer stage becomes less favourable beyond this frequency.

Automatic gain control is a desirable feature in

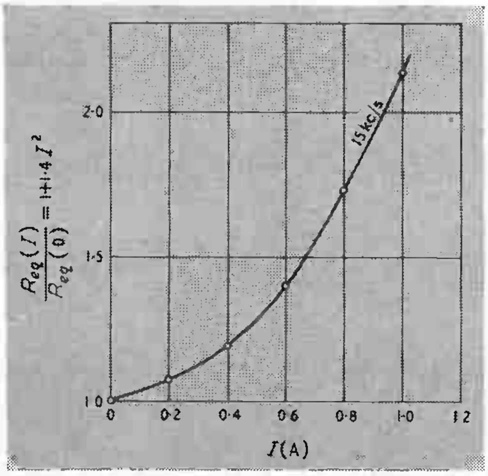


Fig. 18. *Showing the variation of the ratio of the equivalent noise resistance* $R_{eq}(I)$ *to* $R_{eq}(0)$ *with the Hall probe current I, at a frequency of* 15 *kc/s. The value of* $R_{eq}(0)$ *was shown to correspond exactly to the effective resistance between the Hall contacts of the probe*

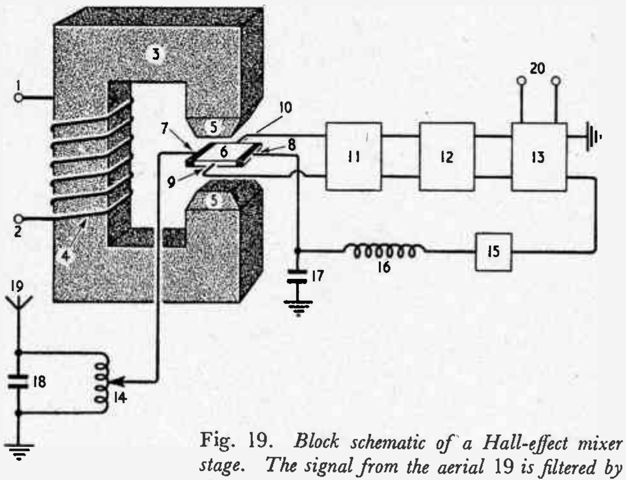


Fig. 19. *Block schematic of a Hall-effect mixer stage. The signal from the aerial* 19 *is filtered by the tank circuit* 18, 14, *and then fed to the Hall probe via the transformer* (*tap* 14) *and contacts* 7 *and* 8. *The capacitor* 17 *closes the h.f. circuit. The Hall probe* 6 *is placed between the pole pieces* 5 *of the ferrite core* 3. *The local oscillator signal is fed to the terminals* 1 *and* 2 *of the coil* 4. *The i.f. output is taken from the Hall contacts* 9 *and* 10, *filtered, amplified and detected at* 11 and 12. *The power rectifier stage* 13, *connected to a power supply* 20, *supplies d.c. output to the Hall probe via a filter* 15, *a choke* 16 *and contacts* 7 *and* 8

mixer stages. This may be obtained with the Hall mixer stages by making use of the temperature dependence of the Hall voltage as in Fig. 2. In this way, the Hall coefficient k_h of Equ. (7) is affected. As this coefficient is independent of B and I, *no distortion* is generated by varying the temperature. Thus, we have a *unique* control, which is entirely free of distortion. This feature compares favourably with the gain controls of all other known mixer stages. The temperature of the Hall probe may be varied by supplying a suitable auxiliary heating current. This may be obtained from a stage 13 (see Fig. 19) which is controlled automatically by the intermediate-frequency output of stage 12. The heating current is supplied to the Hall probe through a suitable filter 15 and a choke 16. The heating time of the Hall probe may be used to obtain a suitable time constant for the gain control.

In a practical arrangement, a small Hall probe was heated up by a heating current of about 1 A and a temperature difference of approximately 60 °C was obtained. According to Fig. 2, a reduction of the Hall voltage by a factor of 5 may be expected at this temperature difference. The time constant was some tenths of a second, depending on the surface cooling of the Hall probe. As the Hall resistance r_h (Fig. 3) is also reduced with rising temperature, a further gain control is obtained by mismatch of the mixer stage to the subsequent intermediate-frequency stage.

The heating current causes a rise in the effective noise temperature T_h of the Hall probe with a resulting increase of the noise figure. This rise in the noise figure does not constitute an impairment of the performance of the mixer stage because the gain control is applied only at high input signal levels.

Acknowledgements

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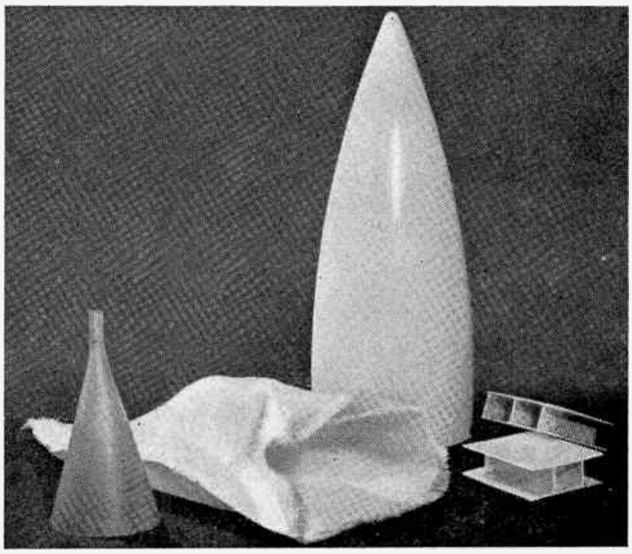
FABRICATION OF RADOMES

A new technique for weaving three-dimensional shapes has provided the manufacturers of radomes with a simple, cheap, and time-saving method for producing these components in reinforced plastics.

A typical reinforced plastic missile nose cone is shown in the illustration, together with one of the contour woven socks from which the cone is manufactured. The illustration also shows a funnel-shaped radar component and a sample of box-form panelling.

The fabrics are being produced by Fothergill & Harvey Ltd., under licence from the Raymond De-Icer & Engineering Company of California.



Triple V.H.F. Reflectometer

MEASUREMENT OF POWER AND REFLECTION COEFFICIENT

By G. H. Millard, B.Sc., A.Inst.P.*

SUMMARY. *A description is given of a multiple reflectometer designed for use in Band I* (41–68 *Mc/s) and Band II* (88–95 *Mc/s) with powers of* 1 *kW or less. Three reflectometers were required for each transmission line and these are combined in one compact unit.*

The reflectometer[1], which samples waves travelling in one particular direction along a transmission line, is a form of asymmetrical directional coupler in which one of the auxiliary outputs has an internal resistive termination. A symbolic representation is shown in Fig. 1; this is an adaptation of a symbol proposed for a symmetrical directional coupler[2]. Most of the power injected at A_1 passes along the straight line A_1A_2 but a proportion follows the curved line A_1B_1 to appear at the auxiliary output B_1. Similarly, a proportion of the power flowing in the direction A_2A_1 is dissipated in the internal resistive termination and, if the reflectometer is perfect, none appears at B_1. Imperfection in this respect is described by the directivity; this may be defined as the ratio of the output at B_1, resulting from power flowing along A_1A_2, to that obtained from the same power flow in the opposite direction, and is usually expressed in decibels. It should be noted that imperfection of the internal termination may reduce the directivity. The output ratio is defined as the square root of the ratio of the power appearing at B_1 to that appearing at A_2, when A_1 represents the input terminals.

It is normal practice in the B.B.C. to insert reflectometers in the aerial feeders at v.h.f. transmitting stations. There are normally two such reflectometers in a unit, one sampling the forward-travelling wave in the feeder to give an indication of the transmitter output power, and the other sampling the reflected wave to give an indication, on a ratiometer, of the reflection coefficient of the aerial and feeder. In addition, a trip mechanism is usually provided in order to protect the transmitters in the event of the aerial or feeder developing a high reflection coefficient. The type of reflectometer used for television and v.h.f. transmissions of 5 kW or greater consists of a small loop inserted in air-spaced coaxial feeder having a 3¼ in. diameter outer conductor. This type of reflectometer would be unduly cumbersome and elaborate for low-power transmissions. For example, the outputs of 0·5 kW, Band I transmitters are carried by 0·5 in. diameter coaxial cable. Connection to the 3¼ in. diameter copper feeder would have to be made with tapering sections, resulting in a heavy unit about 4 ft. in length.

The multiple reflectometer to be described was designed to be cheaper and more convenient while still meeting the essential requirements. It was decided to provide two outputs proportional to the forward-travelling power instead of one, in order to save the cost of an attenuator that would otherwise be required for certain routine tests.

Requirements

The reflectometers for use in Band I were required to pass a combined transmission consisting of 125 W from the sound transmitter and a peak power of 500 W from the vision transmitter. The Band II reflectometers were to carry a single transmission of 1 kW. In each case, two outputs proportional to the forward-travelling power were required corresponding with output ratios of 0·00316 (−50 dB) and 0·00178 (−55 dB). The directivities of these were not required to be high, 20 dB being adequate. In addition, it was required to have an output proportional to power reflected from the aerial, with a directivity greater than 30 dB. The output ratios were to be 0·0100 (−40 dB) for Band I and 0·0178 (−35 dB) for Band II.

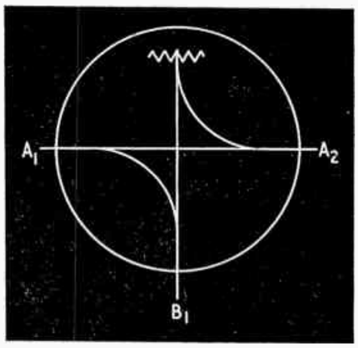


Fig. 1. *Symbol representing a reflectometer*

Each subsidiary line was to have a nominal source impedance of 75 ohms. The main transmission lines

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* *British Broadcasting Corporation*

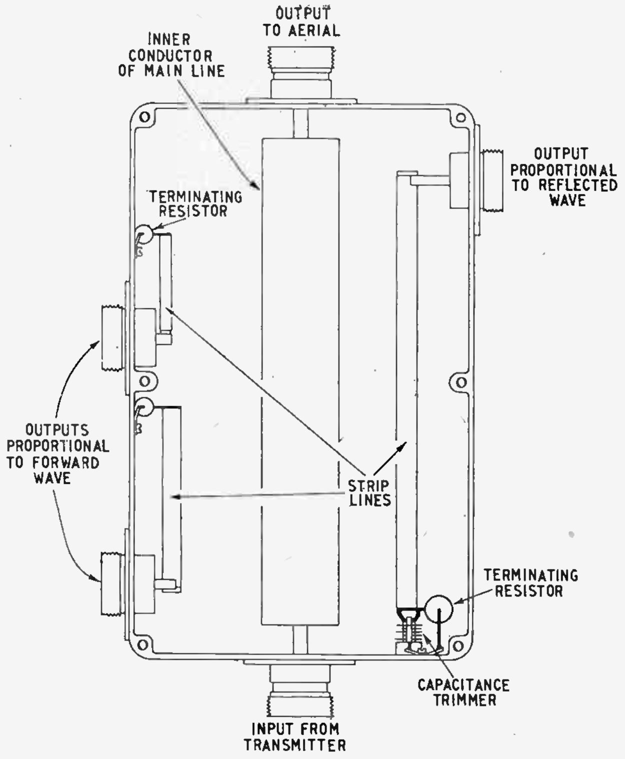

Fig. 2. *Physical arrangement of low-power Band I reflectometer*

were to have a reflection coefficient of less than 0·05 to coaxial cable of 52 ohms characteristic impedance.

Physical Arrangement

A standard die-cast box, 7 in. × 4½ in. × 2 in., which can be bought commercially, is used for the outer conductor. The inner conductor of the main transmission line is supported by 50-ohm coaxial sockets at each end of the box as shown in Fig. 2. The three subsidiary lines have coaxial outputs at the sides, the other ends of these lines being terminated internally.

The inner conductor of the main line is a brass cylinder 1·02 in. in diameter extending for almost the length of the box. Its dimensions were chosen for a characteristic impedance of 52 ohms by means of inductance measurements at 1 Mc/s[2]. Admittance measurements over the frequency range 40–100 Mc/s showed that the reflection coefficient of this arrangement to the coaxial cable UR67 (characteristic impedance 52 ohms) was less than 0·03.

The inner conductors of the subsidiary lines are strips of brass parallel to the inner conductor of the main line. Each strip is connected at one end to the inner conductor of a coaxial socket and at the other to a 75-ohm terminating resistor placed vertically along the side of the box. The subsidiary line which samples the reflected wave in the main line runs the full length of the box, since this line is required to have the highest output ratio. The two lines which sample the forward-travelling wave in the main line are mounted on the opposite side of the box.

Method of Adjustment

With the lay-out adopted, vertical movements of the subsidiary lines (i.e., in a direction normal to the plane of the paper in Fig. 2) varied their characteristic impedance but had little effect upon the output ratio, giving a useful adjustment for the directivity. On the other hand, movements towards the inner conductor of the main line gave a range of adjustment to the output ratio. The technique was, therefore, to adjust firstly for the desired output ratio and then to adjust for directivity.

In order to facilitate the achievement of a high directivity for the reflected-wave line, a small capacitance trimmer was connected across the terminating resistor of this line. Ideally, the entire directivity adjustment could have been carried out by means of reactive and resistive controls on the internal termination, but a pure variable resistance is difficult to achieve at very high frequencies. The alternative used, a variable reactance across a fixed resistor and a line of adjustable impedance, worked well in practice.

Measurement Techniques

Measurement of Output Ratio

A signal generator having a piston attenuator was connected to the main transmission line through a 10-dB 50-ohm resistive attenuator to give the power flow in the direction appropriate to the subsidiary line being measured. A sensitive receiver was connected alternately to the main line and to the subsidiary line through a 20-dB 71-ohm resistive attenuator, the piston attenuator on the signal generator being adjusted to give equal outputs from the receiver. Since there was a mismatch when the receiver was connected to the main line, a correction to the measurements was required. Referring to Fig. 1, when power is applied to A_1, the output ratio may be written

$$\sqrt{\frac{\text{power absorbed by } 75\,\Omega \text{ at } B_1}{\text{power absorbed by } 52\,\Omega \text{ at } A_2}}$$

The ratio actually measured was

$$\sqrt{\frac{\text{power absorbed by } 71\,\Omega \text{ at } B_1}{\text{power absorbed by } 71\,\Omega \text{ at } A_2}}$$

The factor by which the above expressions differ is

$$\sqrt{\frac{\text{power absorbed by } 71\,\Omega \text{ at } A_2}{\text{power absorbed by } 52\,\Omega \text{ at } A_2}} \times \sqrt{\frac{\text{power absorbed by } 75\,\Omega \text{ at } B_1}{\text{power absorbed by } 71\,\Omega \text{ at } B_1}}$$

The second factor may be put equal to unity with negligible error, since the nominal output impedance of B_1 is 75 ohms.

Now, the reflection coefficient, ρ, of a pure resistance of 71 ohms connected to a transmission line of characteristic impedance 52 ohms is 0·156. Accordingly, the correction factor is

$$\sqrt{(1-\rho^2)} = 0{\cdot}987 \text{ or } -0{\cdot}11 \text{ dB}.$$

Measurement of Directivity

A 52-ohm resistive termination was available which had a reflection coefficient to coaxial cable UR67 of about 1% at frequencies below 100 Mc/s. When this termination was used to load the main line, the directivities of the subsidiary lines could be adjusted by setting up for minimum output with the power flow in

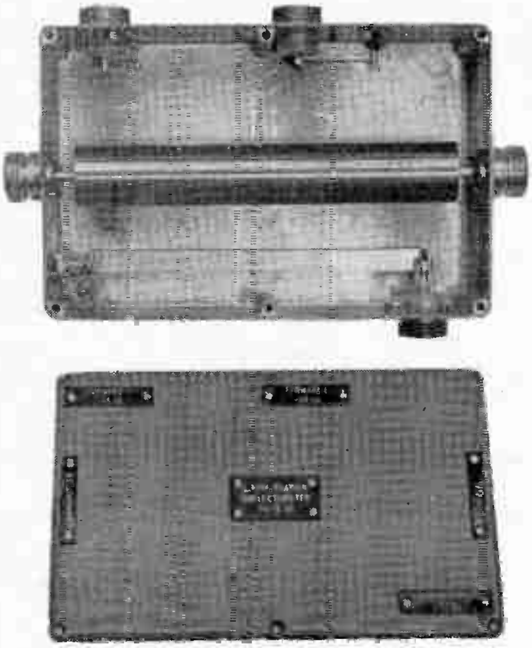

Fig. 3. *Prototype of low-power Band II reflectometer*

the unwanted direction. This procedure was adequate for the two forward-wave lines, but a more refined technique was necessary for the final adjustment of the reflected-wave line. In this case the termination was connected to the main line through about 30 feet of UR67 cable. The output of the reflected-wave line was then the vector sum of two components, one arising from a reflection at the imperfect termination of the main line, and the other arising from imperfect directivity. The output of the reflected-wave line, relative to the level in the main line, was measured over a band of frequencies. Variation in the frequency caused the two components to pass in and out of phase, resulting in a sinusoidal variation of the output. This measurement gave the magnitude of each component ; to avoid any doubt in identifying the components, the match of the load on the main line was degraded and the measurement repeated.

Performance of the Prototypes

Two prototypes were produced, one for each frequency band. A photograph of the Band II reflectometer is shown in Fig. 3. The required ratios were achieved within ± 1 dB at the mid-band frequencies. Since the subsidiary lines are short compared with a wavelength, the output ratios rise with frequency at a rate of 6 dB per octave. Accordingly, the frequency variation was about 4 dB over Band I (41–68 Mc/s) and about 1 dB over Band II (88–95 Mc/s).

No attempt was made to give the forward-wave lines a directivity better than about 20 dB. The directivities of the reflected-wave lines, however, were made as high as possible. The values obtained were 39 dB for the Band I prototype, and 37 dB for the Band II prototype.

In order to check that the main line could carry the required power satisfactorily, it was made to carry a 90-Mc/s pulse of 5-kW peak power ; there was no voltage breakdown. The main line was also made to carry a continuous power of 500 watts at a frequency of 67 Mc/s for a period of about half an hour ; there was no discernible heating.

Conclusions

A cheap and compact multiple reflectometer has been designed for use with low-power transmissions in Bands I and II. A number of units of the Band I design have been produced within the B.B.C. and are now in satisfactory service at low-power transmitting stations.

Acknowledgement

The author wishes to thank the Chief Engineer of the British Broadcasting Corporation for permission to publish this paper.

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METAL SEMICONDUCTOR CONTACTS

'Compression bonding' is the name given to a new Bell Laboratories' technique for attaching leads to semiconductors. The bond is formed by employing a combination of heat and pressure, neither of which is sufficient to cause damage to the semiconductor. Bonds formed in this way can be stronger than the lead itself.

In the case of h.f. transistors where very small leads (0·0004 in. diameter) have to be connected to small areas (0·001 in. × 0·006 in.), the bonding process is very satisfactory. It provides a lead structure which can withstand severe conditions of shock and vibration. It is claimed that this process, when applied to diffused-base germanium transistors with alpha cut-off frequencies in the 500–1,000-Mc/s range, results in a very rugged structure which will withstand 100,000 g in a centrifuge.

Basically, the method consists of pressing the wire (annealed gold has been found to be a good lead material) on the desired area (which has been previously treated with an evaporated and alloyed metallic film, 1,500 to 3,000 angstroms thick) with a half-cylinder shaped tool, and with a force sufficient to produce about an 85% deformation of the wire. During the operation, the temperature is held within the annealing range of the lead material (about 300 °C for gold) while the bond is formed in a non-oxidizing atmosphere. Under these conditions, a good quality metal-semiconductor contact can be made in a matter of seconds.

Details of this bonding process were disclosed in a paper presented recently at the annual convention of the I.R.E. Professional Group on Electron Devices by T. B. Light of Bell Telephone Laboratories.

The Fringe of the Field

By Quantum

THE MASS-SPECTROSCOPE

This is the season for ringing out the old as well as ringing in the new. In every progressive and well-documented field of study it is salutary to take stock, when confronted with some new publication, of what it has displaced; the rapid glance at the not-too-distant past is apt to shake one a little. And, without being so pompous about it, it's rather good fun. Consider, for example, that important branch of applied physics which has so far unaccountably escaped notice in these articles, the rude mechanic art of shove in a cold climate as devised by the Rugby Union. The 1958 "Notes on the Laws of the Game", which supersede the tactlessly entitled "Notes for Guidance of Referees", show how radically things have changed since the days when the standard work was "Why the Whistle Went", that alliterative tribute to W. W. Wakefield (or it may even have been to W. W. Ellis). There are still numbers of grand laws to be finely disregarded, you can still take the ball into your hands and run with it, and Law 15 *B* (8) *b* still treats gravity with a levity quite incompatible with even a coarse regard. But there are great advances in technique; the Union is half a league onward nowadays. I was tempted to a seasonable discussion of this momentous topic, quite appropriate to the fringe of the field, which might have broadened your ideas on binding energy, packing fractions, ground states, and so on, and also have taken your mind off the Test matches, which are merely applied journalism. I could even have waxed poetic about blind-side tactics and touch-judging in the North, particularly at Sale. But, seasonable or not, I felt that this would really hardly be shoving my mass. However, the same general principles shall be my guide. The spirit of pettifogging historical stocktaking will not be denied in this turn-of-the-year discussion of my book-of-the-month, and must be given its head.

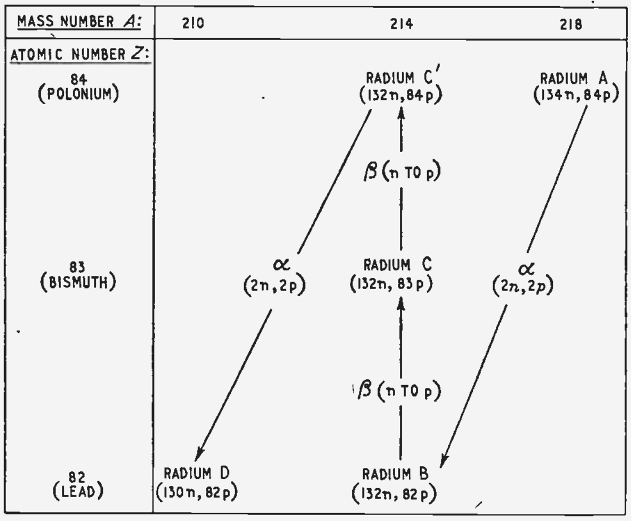


Fig. 1. *Part of Soddy's snakes-and-ladders scheme for the radioactive radium decay series, interpreted in modern terms. Radium A and Radium C′ are isotopes of polonium, Radium C of bismuth, and Radium B and Radium D of lead*

I have just been reading Prof. H. E. Duckworth's "Mass Spectrometry" (Cambridge University Press, 1958), and comparing it with references and memories of about forty years ago. J. A. Crowther, in the 1919 edition of "Ions, Electrons, and Ionizing Radiations", writing of what was then positive-ray analysis, said: "The experimental difficulties are, however, very great, and the method is not likely to come into general use. It has, however, already proved its value in giving evidence of two new gases; one, resembling neon in all its properties but with an atomic weight of 22, the other a gas of atomic or molecular weight 3, which may be either a new element or an allotropic form of hydrogen corresponding to ozone." The last sentence is interesting today! The unidentified stranger cannot have been the radioactive artefact tritium; the abundance of ^{3}He makes this an unlikely candidate, and molecular HD also for the same reason; no doubt it was quadruply ionized ^{12}C. But they were on the look out, even in the early days, for isotopes before they had even accepted the name for them. Chronology is a little haywire here; for Frederick Soddy had, I believe, invented the name in about 1911, and the positive-ray work of J. J. Thomson belongs to about 1912; a third date to fit into the pattern is Moseley's work on atomic numbers in 1913.

Forty years ago, it seemed to the beginner in science that radioactivity was rather more interesting to the chemist than the physicist. J. W. Mellor gave a better account of it than you would find in any physics book, and the whole daring idea of isotopes was explained fully by Soddy in his "Interpretation of Radium", which I was given as a Divinity prize in the belief that it was a commentary on a minor prophet. Neither of these books is with me, nor obtainable anywhere, now. The chemist's atlas was the periodic table, with several gaps now abundantly stopped, and minor snags like iodine and tellurium in the wrong order—for atomic weight determined the sequence—which have long ago been sorted out.

The discovery of several groups of heavy elements, each member of which had the same chemical properties and therefore ought to be in the same place in the table (which is what *isotope* means), but which had different atomic weights and therefore couldn't be, obviously shook the chemists more than it did the physicists. The situation was restored when the atomic number *Z* was made the basis of tabulation, and completely understood from the nuclear side after the discovery of the neutron and from the electron-shell side when the exclusion principle arrived. We now know that the atomic mass number A (which is a whole number, not the actual mass in the $^{16}O = 16$ scale), the number of protons in the nucleus Z, and the number of neutrons N are such that $A = Z + N$. All atoms (or nuclides) with the same Z are isotopes, whatever their A; all nuclides with the same A are isobares, whatever their Z. I have raked all this up because it does seem important to realize that an isotope isn't just something that you separate in a mass-spectrograph or make in a cyclotron or a pile. It originally meant, and still means, one of a group of nuclides which occupy the *same place* in the chemists' periodic table, a place that was really only laid for one. Soddy's contribution, which always seems to have been underrated because he did not follow it up in his later years, was the unravelling of the snakes-and-ladders activities of the α- and β-emitters among the radium and thorium disintegration products. For a β-emission converts a neutron to a proton, and increases Z by one; while an α-emission sheds two neutrons and two protons, putting A down by four and Z by two. Thus, Radium B becomes Radium C^1 after two β-emissions; this ejects an α-particle to become Radium D. Radium B, of atomic weight 214, and Radium D, atomic weight 210, are isotopes—in fact, two of the isotopes of lead (Fig. 1).

Positive Ion Mechanics and Optics

Crowther mentioned the difficulties of positive-ray technique. When the basic equations are written down, with the mere substitution of M for m in the formulae applicable to electrons, one wonders why it is all so much harder than, say, electron optics. The reason is that a homogeneous electron beam is readily obtainable from a suitable cathode; it is not at all easy to get a homogeneous beam of positive ions. The mass spectroscopist is, in fact, like someone who is trying to do an optical interference experiment but cannot start off with a coherent beam of light; he has *two* problems to attend to, the focusing of his apparatus anyhow, and the disciplining of its supply line. But this analogy is not really quite sound, for a spread of velocities among the ions corresponds to optical dispersion; using a non-homogeneous beam is more like having to produce a pure optical spectrum with the help of lenses that can give an unwanted dispersion comparable with that which is actually required of the prism or grating.

(Courtesy, Cambridge University Press)

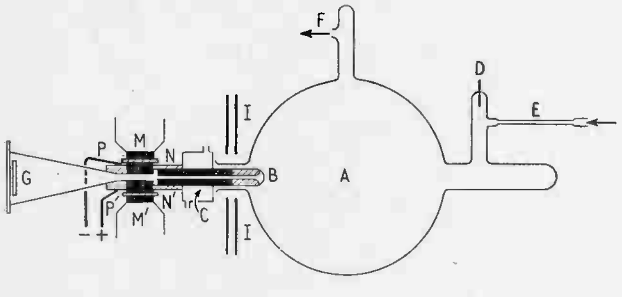


Fig. 2. *Thomson's positive-ray parabola apparatus. The letters indicate the following: A, discharge tube; B, cathode; C, water jacket for cooling cathode; D, anode; E, gas inlet; F, pump lead; G, photographic plate; I, magnetic shield; M, M′, magnetic poles; N, N′, mica for electrical insulation of P, P′, which are pieces of soft iron to serve both as capacitor plates and to define the magnetic field.* (From Aston, 1942)

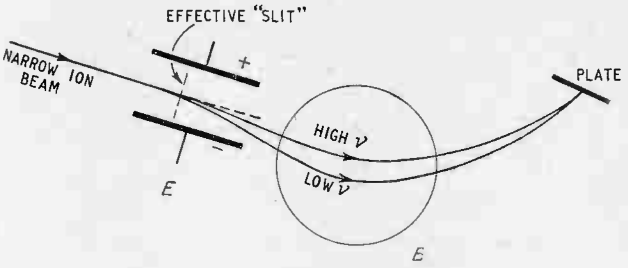


Fig. 3. *Qualitative picture of the focusing and dispersions in Aston's mass-spectrograph. The greater the value of v, the less the downward deflection due to E, and the greater the radius of curvature in B. By suitable arrangement of the dimensions, all ions of the same e/M diverging from an effective slit in E are brought together, whatever their v, to the same place on the plate*

Remembering that we are dealing all the time with positive ions of mass M, each carrying supposedly a single charge e, for acceleration under a p.d. V the velocity v acquired from rest is given by

$$\tfrac{1}{2} M v^2 = e V \quad \ldots \quad (1)$$

The acceleration f parallel to an electric field E is given by

$$M f = E e \quad \ldots \quad (2)$$

from which it follows that an ion moving with velocity v at right angles to E, and traversing a path of length s measured in its original direction, suffers a displacement x parallel to E and towards the positive side of it, such that

$$x = \tfrac{1}{2} E \frac{e}{M} \frac{s^2}{v^2} \quad \ldots \quad (3)$$

In a uniform magnetic field of flux-density B at right-angles to its path, an ion entering the field with velocity v describes a circular arc of radius r, where

$$\frac{M v^2}{r} = B e v \quad \ldots \quad (4)$$

After traversing a distance s measured in the original direction, the displacement at right-angles to the original path is

$$y = s^2 \frac{e}{M} \frac{B}{v} \quad \ldots \quad (5)$$

while from (4) the angular frequency ω for cyclotron resonance in the magnetic field is $\omega = B e/M$, and the cyclotron frequency itself is

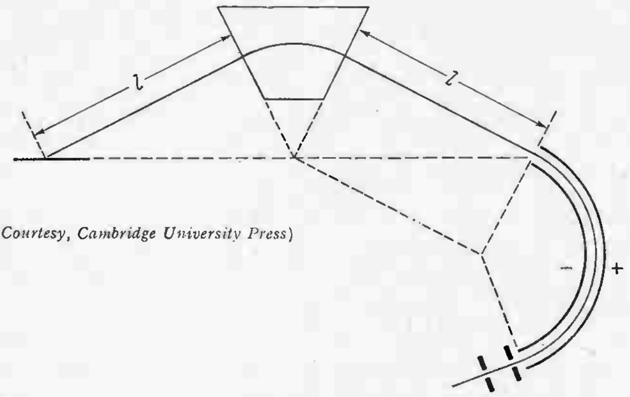
$$f = \frac{B}{2\pi} \frac{e}{M} \quad \ldots \quad (6)$$

Each of these equations contains e/M and, in principle, it ought to be possible to determine e/M for the ion in at least four different and independent ways.

In Thomson's positive-ray apparatus of 1912 (shown in Fig. 2), the electric field E and the magnetic field B

were arranged *parallel* to one another in what we will call the x-direction, and also extended over the same distance s in the path of the ions, which came through the perforated cathode of a discharge-tube containing the gas investigated into the dispersion region, and then impinged on a photographic plate. Eliminating v between (3) and (5), and remembering that s is the same in both equations, $y = 2\, s^2 \frac{B^2}{E} \frac{e}{M} . x$ on emerging from the fields; and as the ions continue in a field-free region the displacements on reaching the plate are proportional to the x and y of this equation. All particles of the same e/M, whatever their v, lie on the same parabolic trace. This calculation, with the appropriate charge instead of e, holds for multiply-charged ions. Continuing the physical-optics analogy, Thomson's method can be likened to a Newton's rings interference experiment, in that coherence conditions look after themselves; thinking of dispersion, it is perhaps really rather more like the crossed-prism experiment of the "Opticks", Book I, prop. ii, theorem 2; there is something about the cunning use of the same s that suggests the Rowland mounting for a concave diffraction grating too. In *principle* it is more elegant than its successors; but it is not capable of high resolution, nor of recording ions present in very small amount because of the wide trace to be covered. That is, it was satisfactory in separating the isotopes of neon and for similar tasks. But to explore the whole spectrum of the periodic table something much more like an optical spectroscope, with a slit and means of producing a sharp ion-image of the slit, is desirable.

In Aston's original mass-spectrograph, slit and plates were at conjugate foci for an ion-optical system (Fig. 3) in which E and B were mutually perpendicular, and so directed that the velocity-dispersion due to E [given in principle by equation (3)], and that of B [similarly represented by equation (5)] were in opposite senses. The magnetic field focuses all ions of the same e/M in the same place; in its dispersion it contrives a double debt to pay, separating ions out with respect to e/M (which is wanted) and also with respect to v (which is not). It is like a prism with a dispersive cylindrical lens attached. The electric field disperses both with respect to e/M and to v, but in this configuration acts like a *diverging* lens. Well, there we are—all the ingredients for an achromatic doublet with respect to v, the electric field playing the part of the diverging flint-glass component and the magnetic field that of the converging crown component. This is the simple slit-lens-prism set-up for getting the first approximation to a pure spectrum. The realization that a suitably shaped radial electric field, between cylindrical, spherical, or even toroidal capacitor plates, could be made to act like a *converging* lens led to the double-focusing arrangement. This is the analogue of the Huygens' eyepiece set-up in optics, which is the only achromatic combination of two *converging* lenses; here distances are critical. Fig. 4 shows the scheme of the Bainbridge-Jordan double-focusing instrument (1936), with the effective object and image distances for both "lenses" indicated. Of course, if you cannot have achromatism the next best thing is to start with a monochromatic beam (in the v sense). An alternative to the electrostatic lens is to have a system of two slits with crossed electric and magnetic fields between, which will transmit ions of only one v (whatever the value of e/M) for a chosen E/B. This is the Wien velocity filter.



(Courtesy, Cambridge University Press)

Fig. 4. *Ion-optical system of the Bainbridge-Jordan double-focusing mass-spectograph. The symbol l denotes an object or image distance. For the upper electrostatic lens-disperser, object and image distances are both zero; for the magnetic lens-disperser, here shown as an equivalent* 60° *prism, each distance is the same, and is three times the distance of the ion beam from the 'vertex' of the prism*

You will find a good account of the ion-optics of the various mass-spectrographs in Duckworth's book. Here I must not let the native hue of resolution (which is what they are in aid of) get sickled o'er with the pale cast of technical detail. Resolving power, $\Delta M/M$ is the important matter. Aston, with his 1925 instrument, got a resolving power of 1/600, and determined M values to one part in 10,000. Since it is impossible for M to be more than about 250, this was a good deal more than would ever be needed to identify isotopes. But by this time the interest had shifted; for *the more accurately the value of M for an ion is known, the more accurately can the binding energy of its nucleus be calculated.* In more recent work, resolving powers of 1/10,000, and M values accurate to 1 part in 100,000 have been obtained. Since the atomic mass unit is equivalent to about 931 MeV, and energy-yielding nuclear changes commonly give something of the order of 10 MeV, this high accuracy in M really only indicates an accuracy of about 10 per cent where it is most needed, in the realm of differences in M. However, work is now right in one fine-structure (or it may even be hyperfine-structure) analogue region.

Accounts of other principles used in mass-spectrometry are given in Chapter 6 of Duckworth's book. There is the time-of-flight system of Smythe and Mattauch, and of Wiley and McLaren, which seems to exploit an arrangement of grids with a klystron-bunching effect. The energy-gain method, which can be likened to the Lamb-Retherford experiment (or perhaps simply to a linear accelerator), uses the absorption of energy from an r.f. field, the resonance frequency of an ion depending on e/M; this does not appear to give resolutions better than about 1 per cent or so, but instruments on this principle have been made small enough to be sent up in rockets. On the other hand, extremely high resolution is available in instruments using the cyclotron-resonance principle. The omegatron of Hipple, Sommer, and

Thomas (which is described more fully, with the theory, in Chapter XXII of Bleaney and Bleaney) really belongs to magnetic resonance spectroscopy, and was used to find e/M for the proton by comparing the cyclotron frequency with the nuclear magnetic resonance frequency. Finally, the Brookhaven mass synchrometer of L. G. Smith, modulating a beam of ions during their circular path in a uniform magnetic field so that they were deflected on to a detector according to their individual cyclotron frequencies, is really an extension of the well-known principle of the cut-off frequency of the magnetron.

Applications of Mass-Spectroscopy

The most fundamental application, which would require a full article to discuss, is the accurate determination of Aston packing-fractions and binding energies. The packing-fraction is $(M - A)/A$, where M is the actual mass of a nucleus in a.m.u., and A the mass-number; this represents the binding energy per nucleon. The curve of packing-fraction against A shows a minimum at about $\mathrm{A} = 60$, from which it was realized in the relatively early days that energy might be released by the fusion of light nuclei, or the fission of heavy ones, a story that is too well known to say more about here. But Aston's work, at moderate resolution, gave a *smooth* curve; with high-resolution instruments it has been shown that there is a fine-structure to it, kinks and changes of gradient that support the view of a shell-structure in the nucleus.

Industrially, the mass-spectroscope is used for hydrocarbon analysis. Under electron bombardment, the vaporized hydrocarbon is dissociated and ionized into a 'cracking pattern' or ions, each substance giving its own relative-abundance spectrum (Fig. 5). By sweeping across the mass spectrum, varying E or H systematically in order to do this, and collecting ions of each M on an electrical detector, the pattern is reproduced on an oscilloscope trace, or by a pen recorder. In fact, adapted in this way, the instrument is as well-known a tool as the infra-red spectrometer.

(Courtesy, Cambridge University Press)

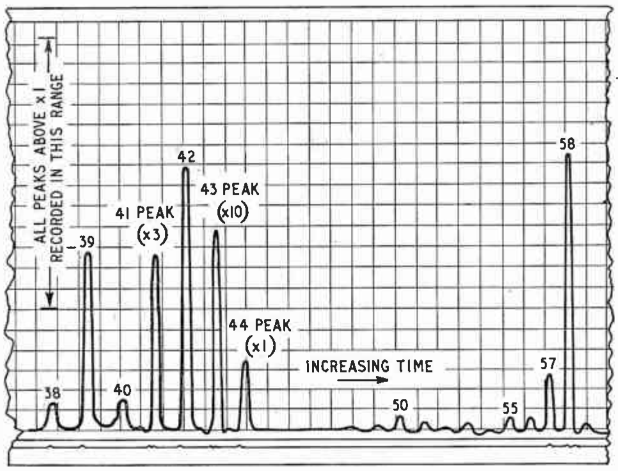


Fig. 5. *Part of the mass-spectrum of n-butane,* $C_4 H_{10}$. $C_3 H_6$, *and* $C_3 H_7$ *fragments, that at* 58 *due to undissociated butane, and the small one at* 59 *the same, but with* ^{12}C *in place of* ^{13}C

A number of other uses are described by Duckworth, but perhaps we might end with another look back into history, at the "Calutron" (named after the California University cyclotron which they pulled to bits in order to make it). This is described in Chapter XII of the famous Smyth Report of 1945. It was a new idea to adapt the mass spectroscope to the collection and production of large quantities of material, and to try to separate U-235 on a usable scale. Before it was tried, it had been believed that on this sort of scale the effects of space-charge would interfere badly. It turned out that they did not, and indeed the separation was successful, and separated uranium isotopes by the milligramme. "If one unit could separate 10 mg a day, 100,000,000 units could separate one ton a day. The questions were of cost and time". This is one of the many solemn thoughts that strike one on re-reading Report; and perhaps it is as well to close on a note of admiration for the scientific imagination that can even visualize mass-spectrometers by the hundred million.

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Reciprocity In Radio-Frequency Measurements

By G. D. Monteath, B.Sc., A.Inst.P., A.M.I.E.E.*

SUMMARY. *It is well known that the reciprocal theorem permits the interchange of source and detector in certain measurements. This article draws attention to advantages to be gained by making an appropriate choice. The interchange of source and detector in standing-wave measurement, a possibility which appears to have been overlooked, is shown to be permissible, and applications are discussed.*

The reciprocal theorem, due to Lord Rayleigh[1], implies that the interchange of a source and a detector will have no effect on the indication given by the detector provided that the impedances of the source and detector are either both zero or both infinite. This well-known result is often used to simplify radio-frequency measurements on linear and passive systems, but some of its applications may be overlooked. For example, it is universally known that the radiation pattern of an aerial is the same in transmission as in reception, and that the source and the detector may be interchanged when measuring impedance with a bridge, but it is less generally appreciated that one alternative may be better than the other. The possibility of using a standing-wave indicator with source and detector interchanged does not seem to have been considered at all.

One situation in which one of two reciprocal alternatives may be preferable occurs when screening is imperfect. If leakage causes unwanted coupling between two components, it is usually better to choose the alternative giving the strongest wanted coupling between these components.

Another occurs when the input to the system under investigation must be restricted to a very low level, so much so that it is difficult to obtain adequate sensitivity. One example is standing-wave measurement at the input to a crystal mixer. The general rule for obtaining the greatest sensitivity for a given input to the unknown is to choose the arrangement in which the receiver is most strongly coupled to the unknown.

Conditions for Interchanging Source and Detector

The fact that the source and detector can be interchanged in null methods of measurement needs no proof but, in certain other types of measurement, the permissibility of interchanging the source and the detector is not always obvious and indeed, in some cases, this interchange may not be permissible. Direct application of the reciprocal theorem permits the interchange of the source and the detector only when both have zero impedance or both have infinite impedance.

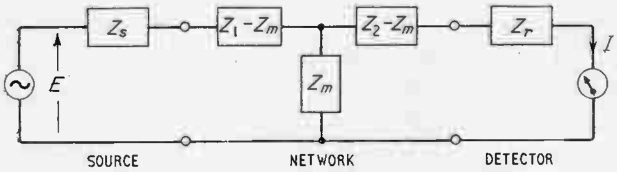

Fig. 1. *A two terminal-pair network connected between a source and a detector*

Fig. 1 shows a source, a detector, and a two terminal-pair network, the latter being characterized by self and mutual impedances Z_1, Z_2 and Z_m. It is easily shown that the ratio of detector current I to source e.m.f. E is

$$I/E = \frac{Z_m}{(Z_1 + Z_s)(Z_2 + Z_r) - Z_m^2} \quad \cdots \quad \cdots \quad (1)$$

If the source and detector are interchanged, Z_1 and Z_2 in this expression are interchanged. I is therefore changed to I' where

$$I'/I = \frac{(Z_1 + Z_s)(Z_2 + Z_r) - Z_m^2}{(Z_2 + Z_s)(Z_1 + Z_r) - Z_m^2} \quad \cdots \quad \cdots \quad (2)$$

This ratio is equal to unity if

(i) $Z_1 = Z_2$,
or (ii) $Z_r = Z_s$

and it is approximately equal to unity if

$$\text{(iii)} \left\{ \begin{array}{l} Z_1 \ll Z_r \gg Z_2 \\ \text{and } Z_1 \ll Z_s \gg Z_2 \end{array} \right.$$

$$\text{or (iv)} \left\{ \begin{array}{l} Z_1 \gg Z_r \ll Z_2 \\ \text{and } Z_1 \gg Z_s \ll Z_2 \end{array} \right.$$

If we are interested only in *changes* in the insertion loss of the network, as in standing-wave or radiation pattern measurement, the interchange of source and detector is also permissible if

(v) changes in Z_1 and Z_2 are negligible, and Z_m is so small that $Z_m{}^2$ is negligible in the denominator of Equ. (1).

I'/I in Equ. (2) is then constant, although it may not be unity.

It is almost self-evident that a source and a detector may be interchanged if their impedances could be interchanged. This statement, illustrated in Fig. 2, is established by supposing a zero-impedance source

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* *British Broadcasting Corporation.*

and detector to be interchanged, leaving the series impedances Z_a and Z_b in position. A theorem due to Wigan[2] corresponds to the limiting case when Z_a and Z_b are respectively zero and infinite.

Radiation Pattern Measurement

The usual procedure is to rotate the aerial under examination and observe changes in the insertion loss between it and a fixed aerial. It is well known that either aerial may be used for transmission, but it is instructive to reconsider this fact in the light of the previous section, where (v) is the relevant condition.

The requirement that Z_m^2 should be negligible implies that signals reflected twice, and therefore traversing the path between the aerials three times, may be neglected. The constancy of Z_1 and Z_2 implies that rotation of the aerial under examination must have no effect on its self-impedance or on that of the second aerial. These conditions would clearly be satisfied in any reasonable measuring arrangement. The use of a non-linear detector does not affect these conclusions, but if its law depends on the impedance to which it is connected, calibration must be carried out under the appropriate conditions.

The choice of the best alternative may be dictated by considerations of screening and interference. For example, if the layout has been determined by other considerations, and if the receiver is poorly screened, the receiver should be connected to the aerial to which it has the greatest stray coupling. On the other hand, one of the two alternative reciprocal arrangements may be less susceptible to incoming interference, or may be less likely to cause interference elsewhere.

Impedance Measurements

Standing-wave measurement in transmission line or waveguide, which is illustrated in Fig. 3, resembles radiation-pattern measurement in that an arrangement giving accurate results must entail weak coupling between the source and detector. The impedances offered to the source and detector must be substantially unaffected by moving the probe along the line, and Z_m will be small, so that condition (v) will be satisfied. If the coupling between the probe and the inner conductor is excessive an error will be caused, whichever of the two reciprocal arrangements is used. The error will be different for the alternative arrangements only in so far as the impedance of the detector differs from that of the source.

There is no reason for the difficulties associated with finite probe coupling to be more severe with the arrangement of Fig. 3 (b) than with that of Fig. 3 (a), but with the arrangement of Fig. 3 (b) the difficulties will appear in a less familiar form, and some aspects may be overlooked. For example, matching appears to suggest itself more strongly in the case of a source than in that of a detector, so that when using the arrangement of Fig. 3 (b) one might be tempted to seek a perfect match between the source and the probe. In so far as this procedure could be successful, it would achieve strong coupling for a small probe penetration; indeed, if the matching could be done with lossless elements, the effect would be the same as if a direct connection were made between the probe and the inner conductor. Considerable error could therefore result.

One difficulty peculiar to the arrangement of Fig. 3 (b) arises if the detector has a non-linear law dependent upon the impedance of the source to which it is connected; e.g., a crystal detector operated at levels above the square-law region. It is customary to calibrate such a detector against the standing-wave pattern measured when the transmission line is short-circuited. This procedure is quite satisfactory with the arrangement of Fig. 3 (a), because the detector will see substantially the same impedance during calibration as during a subsequent impedance measurement, but error could result from the application of this method to the arrangement of Fig. 3 (b). One solution is to connect the detector via an attenuating pad, but some sacrifice of sensitivity would be entailed. Alternatively, care could be taken to ensure that the signal reaching the detector is always sufficiently weak for a square law to apply.

All the difficulties referred to are absent if a signal generator and superheterodyne receiver are used as source and detector, one of these being connected by a flexible cable to the probe. Measurements are made by adjusting a piston attenuator in the signal generator so as always to restore the receiver input to a fixed level[3]. Whichever of the arrangements of Fig. 3 is used, the use of a receiver and signal generator is very convenient, since it avoids detector calibration, enables a wide range of standing-wave ratios to be measured, and gives a high sensitivity, so that probe tuning is often unnecessary. At frequencies up to at least 1,000 Mc/s, the trailing cable causes no appreciable error; indeed, Buchanan[4] has used a cable and probe with a slotted waveguide at 24,000 Mc/s. As an alternative, a klystron oscillator could be mounted directly on the probe carriage.

The less conventional arrangement of Fig. 3 (b) is always to be preferred when it is necessary to restrict the input to the unknown impedance, for example, when the unknown is the input of a crystal mixer, or an aerial which could cause interference by radiation. In the latter case, it also minimizes the effect of any stray coupling between the aerial and the receiver.

The considerations determining which reciprocal

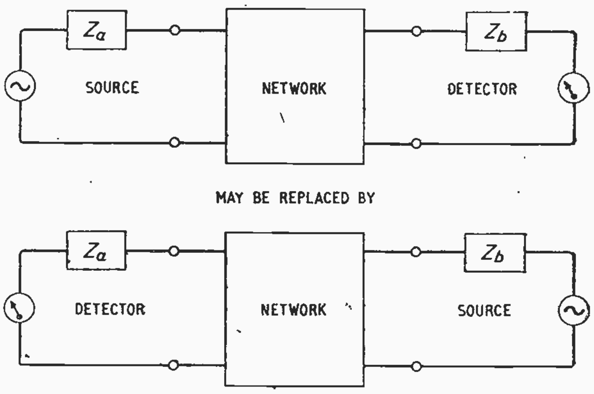

Fig. 2. *Illustrating the reciprocal theorem. Source and detector of zero impedance interchanged, leaving impedances Z_a and Z_b in position*

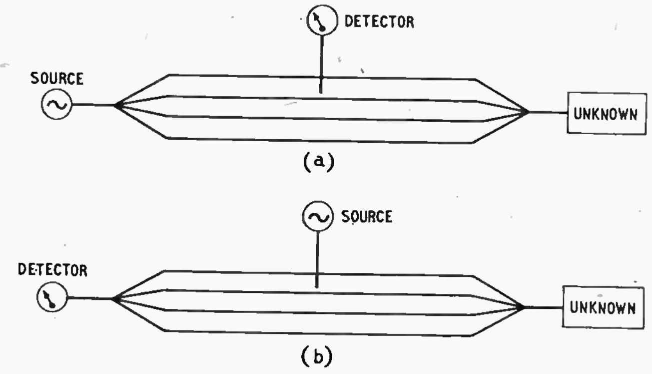

Fig. 3. *Standing-wave measurement. The conventional arrangement is shown at (a) with the reciprocal at (b)*

alternative is to be used for standing-wave measurements also apply to the use of any impedance measuring device in which either the source or the detector is strongly coupled to the unknown while the other is coupled weakly. One example is a commercial admittance meter described by Thurston[5]. Basically, this consists of three transmission lines connected in parallel to the source. These lines are terminated respectively by a standard conductance, a standard susceptance and the unknown admittance. The current entering each branch is sampled by a loop, whose coupling can be varied, and the three loops are connected in parallel to a detector so that a null is obtained when the sum of the voltages induced in them is zero. At the lowest frequency, at which the admittance meter is used (about 60 Mc/s), the coupling between the loops and the associated transmission lines is weak, so that the sensitivity is low. The source is, however, connected directly to the unknown. The low sensitivity and strong coupling between the source and unknown has been found to cause an error when measuring the admittance of an aerial, using a poorly screened receiver. In every case, interchange of the source and the detector effected a cure.

Conclusion

A theoretical investigation can often be simplified by supposing a source and a detector to be interchanged, and it is always advantageous to keep both reciprocal possibilities constantly in mind, so that the most convenient can be used. The need to preserve the same flexibility in devising methods of measurement is less commonly appreciated; indeed, one of two reciprocal arrangements is often chosen simply because it is the one most amenable to theoretical treatment.

In commercial instruments, such as impedance bridges, it is the universal practice to label two pairs of terminals 'Source' and 'Detector' respectively, and the user tends to follow the maker's instructions without question. It would be better if the terminals were labelled 'Source or Detector' and 'Detector or Source'. A user so misguided as to attempt a measurement with two sources or two detectors need hardly be considered.

Acknowledgement

The author is indebted to the Chief Engineer of the British Broadcasting Corporation for permission to publish this article.

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CLIP-ON D.C. MILLIAMMETER

A clip-on probe (illustrated) forms part of the Hewlett-Packard d.c. milliammeter type 428A which is designed to facilitate the measurement of direct current.

The probe is sensitive to the magnetic field produced by the current being measured. As it extracts no energy from the field, the milliammeter has the unusual characteristic that it introduces no resistance (and less than 0·5 μH) into the circuit being measured. In transistor work, where low-impedance circuits are common, this property is of great value.

Illustrating the construction of the probe in which the flanges on the body open a set of split jaws for clipping on the conductor to be measured. A spring return closes the jaws when the flanges are released. The measurement sensitivity can be increased by looping the conductor through the probe several times so as to increase the effective magnetizing force of the measured current. The probe accepts single conductors of up to $\frac{3}{16}$ in. (overall diameter)

A specially developed magnetic amplifier is contained within the probe and provides an a.c. output (at 40 kc/s) which is proportional to the magnetizing force produced by the direct current being measured. The level of a.c. in the probe is small enough to have negligible effect on most circuits with which the equipment is used.

The construction of the probe is such that no effects from the earth's field are noticed (except on the most sensitive range of the milliammeter) and the magnetic sensing elements in the probe are sufficiently free from remanence or residual induction effects for residual readings to be negligible.

COMPUTER TRAINING COURSES

Training courses for the users of Emidec computers have been arranged by E.M.I. Electronics Ltd. These courses, which are held at Hayes, Middlesex, last for three weeks; the next course is scheduled to commence on 12th January 1959.

Demodulation and Detection

By D. A. Bell. Ph.D., M.I.E.E.*

A sporadic debate on etymology over the past quarter of a century has tended to divert attention from the fact that there is a very real difference between 'demodulator' and one specific use of the word 'detector', the specific use being one that was significant in the early days of wireless telegraphy and is again significant now in the light of Shannon's theory of communication. First to dispose of the etymological argument, that the word demodulator should mean 'destroyer of modulation' rather than 'extractor of modulation', let us say that the meaning of words is determined by their usage ; and by usage 'demodulator' now means a device for extracting from a modulated signal the original modulating waveform. The demodulator must then be designed to match the modulator (Ville[1] has used the term 'conjugate' to describe the relation between modulator and demodulator) and will take various forms according to the type of modulation. We shall therefore call those devices which employ a local carrier and those devices which give an output proportional to the envelope of the modulated signal coherent and incoherent demodulators respectively, reserving the term 'detector' for a different *function*, rather than a different *mode of operation*.

In communication theory, both the modulator and the demodulator may be regarded as special cases of re-coding devices. By 're-coding' we mean, in general, the changing of the physical representation of a message, and within this definition both transducers and modulators would have to be included as re-coding devices. It is perhaps more usual to reserve the term 'coding' for a device which conserves the information but either changes the redundancy in the signal or makes an exchange of signal-to-noise ratio against bandwidth; but a frequency-modulator increases the redundancy of the signal, and therefore is certainly a coding device, and some varieties of envelope demodulator change the signal-to-noise ratio and so come near to re-coding. (The process of *coding* is best represented by the geometric analogy of *mapping* ; which in the mathematical sense means that there is a one-to-one correspondence between each of a set of points in one space and the corresponding members of a set in another space, though the latter space does not necessarily have the same number of dimensions as the first space.) A demodulator which employs a local oscillator (the homodyne or its practical version, the synchrodyne) may be thought of as the exact inverse or conjugate of an amplitude modulator ; the modulator changes the signal frequency from a band $0 - n$ c/s to $f_c \pm n$ c/s, and the demodulator frequency-changes back from carrier-frequency f_c to carrier-frequency zero. Thus, a superheterodyne receiver equipped with a homodyne demodulator might fairly be said to have first and second frequency-changers, whereas at one time it would have been incorrectly described as having first and second detectors. The combination of frequency-discriminator and rectifiers constitutes a demodulator of frequency-modulated signals, as also does a cycle-counting circuit.

Detection

The word *detection* is reserved here for the device which, in the simplest case, decides whether or not a signal is present or, in general, decides which one of a pre-arranged ensemble of possible signals is present. The coherer was clearly a true detector but not a demodulator since it could not respond even to the on/off modulation of telegraphy. (After detecting a signal it required re-setting by mechanical vibration, so that it could not of itself show the cessation of a signal.) The crystal rectifier which followed the coherer naturally inherited the name of detector but, since it gave an output which bore a functional relationship to the input it was, in fact, a demodulator as well. At this stage the term rectifier also appears. A rectifier will always function as a detector, and in combination with a suitable smoothing circuit it will also act as a demodulator of a.m. signals, though it is in practice difficult to make it a distortionless detector of 100% a.m.

In this article it is *function* which is being classified into demodulation, rectification and detection, and the necessity for classifying *function* rather than *physical device* can be seen by considering the behaviour of a diode rectifier in relation to amplitude-modulated signals. For a small input the diode characteristic may be taken as approximately square-law. It therefore behaves as a square-law rectifier and as an incoherent demodulator of the envelope type with the signal-to-noise ratio the same in the output as in the input. If the input level is raised, the diode becomes a linear rectifier; but if signal and noise are of comparable intensity it remains an incoherent demodulator, though with a complicated relationship between output and input signal-to-noise ratios. But if the signal-to-noise ratio is very high, the linear rectifier used as demodulator behaves like a coherent demodulator. Thus, the one physical device may behave as a square-law or linear rectifier and (with its associated smoothing circuit) as incoherent or coherent demodulator, according to the input-signal conditions.

The distinction between detection and demodulation

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can be illustrated from R. A. Smith's study[2] of three types of circuit; namely, the square-law rectifier, linear rectifier, and coherent demodulator. He used the symbol S for ratio of signal or message[†] amplitude to r.m.s. noise, but we shall here use P for signal-to-noise power ratio, with the same suffixes as Smith used for two formulations of the ratio. Then P_0 is defined as the ratio of 'power in the d.c. component of output due to signal' to 'noise power in the absence of signal'. Although the postulated separation of the d.c. component from the rest of the output is debatable, P_0 is clearly a *detection* criterion since it compares one of the effects of a signal with the background which is present in the absence of a signal. His other criterion P_s is the ratio of d.c. signal power to noise power in the presence of a signal; i.e., it is the signal-to-noise ratio in the *demodulated signal*, and is not concerned with discrimination between signal and no-signal conditions. Both P_0 and P_s can be computed for various types of circuit device and levels of signal-to-noise at the input; results for square-law, linear and coherent detectors, with small and large input signal-to-noise ratios and with various subsequent bandwidth limitations, have been tabulated by R. A. Smith.[‡]

It appears that a coherent device is the best demodulator, since P_s is twice the input signal-to-noise ratio when the latter is large; but a square-law device is a better detector of the presence of a strong signal since P_0 is then equal to the square of the input signal-to-noise ratio. (The coherent device is naturally the better detector if the signal is initially below the noise.)

The classic applications and studies of detection have been related to simple on/off signals, whether for telegraphy or radar. But another important application of detection arises from Shannon's theory of the possibility of communicating with negligible risk of error, provided that the messages to be communicated are selected from a pre-arranged group of possible messages which is finite in number. In the geometrical analogy of the error-free communication process, the final stage (see, for example, reference 3) involves the identification of the received signal with that one of the possible messages which is most likely to have originated the signal in question. This positive identification, the replacement of a set of probabilities relating the received signal to various possible messages by a statement that with probability approaching unity one specific message was transmitted and all others may be ignored, can properly be called the *detection* of the message. The manner in which one makes a firm choice of a particular one out of a set of possibilities is governed by a *decision function*; a great deal of mathematical work has been done on decision functions, some of which are intended to be related to human behaviour in the strategies of games, business enterprise and war. But the decision-making function of a detector in a communication system has to be mechanized and, in practice, it is worthwhile considering the very simplest forms of decision-function, such as the use of a threshold or critical value of some parameter.

Cross-Correlation and Detection

The first clear example of this approach was the 'pre-detection filter'[4] proposed for radar, the word detection, in that case, being used precisely in the sense defined in this article. The decision function was simply a comparison of instantaneous powers (squared amplitudes), for the filter was designed to maximize the signal-to-noise ratio at a particular instant; i.e., to convert the incoming message into a high-amplitude impulse which would stand out above the noise. The decision-making could then be instrumented in the form of a slicer, though in the radar case a human operator might be employed as the amplitude-discriminator. It is worth noting that the final act of 'detection' or 'decision making' is always a strongly non-linear process, so that if the electrical chain consists only of linear elements, ending perhaps in a visual display, the human observer provides the non-linear element. An example of a signal-detection system in which the decision is instrumented is the 'auto-alarm' used to keep watch for distress signals at sea. Here a specific signal (twelve long dashes) is detected, and not merely the presence of some signal, and action results from reception of the specific signal only.

When testing for specific signals an obvious technique in principle is to compare the received signal, $f_1(t)$, with a local replica $f_2(t)$ of the message which is believed to be probably responsible for the signal, and this comparison can be carried out by evaluating the cross-correlation function

$$\psi_{12} = \int f_1(t) \, . \, f_2(t) \, dt \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

This function maximizes when $f_1(t) \equiv f_2(t)$ and so provides a means of discrimination: if f_1 is compared with $f_2, f_3 \ldots$, the function f_n which gives the greatest value of ψ_{1n} is the most similar to f_1. Such detection might well be called cross-correlation detection, though with some risk of confusion with schemes of detection which are based on various types of *auto*-correlation process and which are often referred to as 'correlation detectors'. For the frequency analysis of signals, auto-correlation and Fourier analysis are closely equivalent as has been pointed out by Golay[5] and as is indeed obvious from the directness of the mathematical transformation between power spectrum and auto-correlation function. A true Fourier analysis is theoretically superior since it retains phase information which is suppressed in the power spectrum, but in practice it is very difficult to perform a true Fourier analysis including values of relative initial phases of the several components. The auto-correlation technique is valuable for signals of very low frequency, since noise can then be minimized by extending the integration period of the auto-correlation over a long time, whereas it would not be practicable to construct a frequency filter having a bandwidth of a fraction of a cycle per second. However, we are not here concerned with auto-correlation processes.

In order to instrument cross-correlation detection using Equ. (1), let us translate from the time plane to the frequency plane using the Fourier transforms

$$\left.\begin{aligned} F(j\omega) &= \int_0^\infty f(t) \, e^{j\omega t} \, dt \\ f(t) &= \frac{1}{2\pi} \int_0^\infty F(j\omega) \, e^{-j\omega t} \, d\omega \end{aligned}\right\} \qquad \ldots \quad \ldots \quad (2)$$

[†] The terminology is that *message* corresponds to that which was originally transmitted, whereas *signal* is that which is received; i.e., message plus noise.

[‡] In Reference[2], but note that his S_0 and S_s refer to *amplitude* ratios.

The integral of the product of Fourier transforms enters into Parseval's theorem, and the following treatment of Equ. (1) parallels the derivation of Parseval's theorem for Fourier integrals which has been given by Jeffreys and Jeffreys[6]. First replace $f_1(t)$ by its Fourier transform.

$$\psi_{12} = \int f_2(t) . \left[\frac{1}{2\pi} \int_0^\infty F_1(j\omega) \, e^{-j\omega t} \, d\omega \right] dt \qquad .. \qquad (3)$$

Then change the order of integration and write Equ. (3) as

$$\psi_{12} = \frac{1}{2\pi} \int_0^\infty d\omega \int_0^\infty F_1(j\omega) \, f_2(t) \, e^{-j\omega t} \, dt \qquad .. \qquad .. \qquad (4)$$

But $\int f_2(t) \; e^{-j\omega t} \; dt$ differs from $F_2(j\omega)$ only in having $-j\omega t$ instead of $+j\omega t$ in the exponent, so that on changing the variable from $+\omega$ to $-\omega$ it may be written $F_2\,(-j\omega)$, and Equ. (4) becomes

$$\psi_{12} = \frac{1}{2\pi} \int_0^\infty F_2(-j\omega) . \; F_1(j\omega) \; d\omega \qquad .. \qquad .. \qquad (5)$$

Now ψ_{12} maximizes when $f_1 = f_2$ or in the frequency plane $F_2 = F_1{}^*$ (complex conjugates).

The first limitation of this treatment is that it ignores the effect of noise. This is not too serious because random noise is, by definition, uncorrelated with any specific signal, and the effect of adding random noise $f_\nu(t)$ to f_1 is, therefore, to make an addition to ψ_{12} which will, on the average, be the same for all f_2 and the ratio of $\int f_\nu(t)\, f_2(t)\, dt$ to $\int f_1(t)\, f_2(t)\, dt$ will tend to zero for t large compared with $1/\omega_\nu$. The second limitation is that since it has been derived via Fourier transforms the integrals must all extend to an upper limit of infinity, whereas in practice Equ. (1) will be evaluated over a finite range of time.

Now the pre-detection filter work already cited-arrives at a figure of merit

$$M = \frac{\left| \int_{-\infty}^{\infty} F_2(\omega) . \; F_1(\omega) \; e^{j\omega t_0} \; d\omega \right|^2}{\int_{-\infty}^{\infty} | F_1(\omega) \; |^2 \; d\omega \int_{-\infty}^{\infty} | F_2(\omega) \; |^2 \; d\omega} \qquad .. \qquad .. \qquad (6)$$

which is shown to maximize when

$$F_2(\omega) = [F_1(\omega) \; e^{j\omega t_0}]^* \qquad .. \qquad .. \qquad .. \qquad (7)$$

This is the same as the condition found for maximizing Equ. (5) except that the rotation of phases of the constituents of $F_1(\omega)$ by $e^{j\omega t_0}$ represents a shift in the origin of time from zero to t_0.

The purpose of carrying out this comparison is to show that the 'pre-detection filter' is, in fact, the same as a device which can carry out detection by performing the cross-correlation of the received signal with the supposed message. Thus, the construction of a cross-correlation detector is converted to the problem of designing a filter of suitable characteristics. The physical interpretation of Equ. (7) is well known: $F_2(\omega)$ is the spectral response of a filter such that its impulse response is the mirror image with respect to t_0 of $f_1(t)$. This makes it legitimate to regard the filter as holding a replica of the message, as is required for a cross-correlation process. The filter is still, strictly speaking, a pre-detection filter, and not in itself a detector. It remains to perform the act of detection by means of some threshold type of device, which records 'message identified' if the value of M in Equ. (6) exceeds some specified value. The choice of this value by the user then determines the probabilities of the two kinds of error, identifying a message falsely or failing to identify a message when it does match the filter. The magnitudes of these probabilities depend also on the signal-to-noise ratio and on the distribution of the amplitudes of the noise output from the pre-detection filter, which need not always be a Gaussian distribution.

Equ. (6) assumes white noise but, if the spectral distribution of the noise power is not uniform, the design problem can often be solved by postulating the inclusion of a filter which will convert the given noise to white noise[7]. Equ. (7) does not guarantee the physical realizability of the filter and this problem, especially acute when the 'memory' of the filter is to be limited to a prescribed time, is also discussed by Zadeh and Ragazzini.[7]

Band-Limiting Filters

It is also important to distinguish these pre-detection filters, one of which must be provided specifically for each of the messages to be identified in the received signals, from filters, such as Wiener's optimum filter, which are designed to improve *average* signal-to-noise ratio. The latter are designed to be statistically optimum for a whole class of messages, a class described in terms of the mean power spectrum resulting from the transmission in random order of all messages in the class.

An intermediate class is that designed to maximise the signal/noise energy ratio for *a specific signal*, and an example of this type for a rectangular-pulse signal has been described by Heaps and Isaacs[8]: assuming the use of a Butterworth type of filter, they computed the values of design parameters that would maximise the ratio of signal energy to noise energy. This does not solve the problem of detection, and one would presumably follow this filter with a power responsive device (e.g., square-law rectifier) and an integrator with memory time of the order of the pulse duration. Since this complete system includes a non-linear device before the integrator, it is not possible to combine the latter with the filter. The impulse response of an ideal low-pass filter is a sin x/x function, so that a low-pass filter does not fulfil the criterion of formula (7), that the pulse response of the filter should be the image of the signal wave-form.

If the pulse is symmetrical its frequency function will contain only even powers of $j\omega$ and $F \equiv F^*$. Then if one ignored the factor $e^{j\omega t_0}$ it would appear that the filter was merely the one which had a spectral shape similar to that of the signal. Even this would not mean a clear transmission of all components of the signal, but the reduction of transmission of frequencies at which the signal components are weaker could be explained as a means of discrimination against signals of other forms. But in general, with an asymmetric signal, the spectrum of the conjugate filter is not the same as the spectrum of the signal. The most important feature is that the pre-detection filter is designed to maximize the *instantaneous* ratio of squared amplitudes of signal and noise at time t_0, not the mean-square ratio; and for this purpose the filter must operate on the relative phases of the components, and not merely on amplitudes, and the phase-shift $e^{j\omega t_0}$ cannot be ignored.

Conclusions

It is shown that a distinction can be drawn between *demodulation,* which is merely a slight re-coding of the signal, and *detection* which has a special significance in relation to Shannon's theory of communication. Detection is usually performed by picking that one of a group of alternatives which has the largest value of some parameter. It is then necessary to construct a set of devices of which each will maximize the parameter in question when the in-coming signal is identical with that for which the individual device is designed. It appears moreover that the pre-detection filter of Van Vleck and Middleton can be regarded as carrying out a cross-correlation of the received signal with the message against which it is to be tested.

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MATHEMATICAL TOOLS

By Computer

Solving Linear Simultaneous Equations

It is quite often necessary to solve numerical simultaneous equations where the number of equations and unknowns is considerable, say 4 or more; an obvious case of this is in multiple-correlation analysis associated with statistical investigations. If the number of equations and unknowns is very large, say 10 or more, the best course is probably to obtain a solution by means of an electronic computer, which is well adapted for this purpose. Here we shall discuss the matter in terms of solving 4 equations with 4 unknowns, but in a perfectly general way.

Mathematically, the solution of n linear equations for n unknowns can be written down immediately by means of 'determinants', and the solution is unique provided that a certain 'determinant' associated with the coefficients of the equations is different from zero. If this 'determinant' is small, the equations are said to be 'ill-conditioned' (that is, one of the equations is nearly a linear combination of the others) and therefore small errors in the data associated with the equations may produce far greater errors in the results. But the numerical evaluation of these 'determinants' is a somewhat tedious process, and is likely to involve differences between nearly equal quantities.

Various well-known methods are available for the numerical solution of equations; we have chosen the 'Escalator' method described below, due to Morris[1], because it can easily be reduced to a standard computational routine involving only elementary arithmetical processes, and because there are many points in the process where checks have to be satisfied so that errors can be discovered before too much time is wasted. If the obvious method of 'pivotal condensation' is used, so that one variable is successively eliminated from all the equations in turn, a numerical error may remain undetected until the whole process of solution is complete, and every number may have to be examined before the incorrect one is discovered. In the 'Escalator' process to be described, the search for the error is more restricted, as will be explained.

We shall only discuss the case where the equations are symmetrical, that is to say, the coefficient of the rth unknown in the sth equation is the same as the coefficient of the sth unknown in the rth equation. Equations not having this property can easily be replaced by an equivalent symmetrical set as explained in the Appendix.

We shall now solve the symmetrical equations:

$$x+0\cdot 08711\,y+0\cdot 42521\,z+0\cdot 37688\,t+0\cdot 28009=0 \tag{1}$$

$$0\cdot 08711\,x+y+0\cdot 16249\,z+0\cdot 14094\,t+0\cdot 23215=0 \tag{2}$$

$$0\cdot 42521\,x+0\cdot 16249\,y+z+0\cdot 91117\,t-0\cdot 40936=0 \tag{3}$$

$$0\cdot 37688\,x+0\cdot 14094\,y+0\cdot 91117\,z+t-0\cdot 26253=0 \tag{4}$$

If the normal 'pivotal condensation' procedure is used, Equ. (1) is multiplied first by 0·08711 and subtracted from Equ. (2), then by 0·42521 and subtracted from Equ. (3) and, lastly, by 0·37688 and subtracted from Equ. (4). We thus obtain a set of three new linear simultaneous equations in y, z and t from which x is absent. Then y is eliminated from these equations, and we derive two simultaneous equations in z and t for which the solution is straightforward. But the values of z and t thus derived, and the values of x and y which can be deduced from these, must be checked in all the original equations, and if there is an error, we have 14 coefficients to check when trying to trace the error [$\frac{1}{6}n(n-1)(n+4)-2$ such coefficients if originally we had n equations for n unknowns].

The 'Escalator' process to be described works in the opposite direction. The first step is to complete Table 1 in the manner we now explain. In Equ. (1), suppose that the number term 0·28009 is omitted, that z and t are both zero and that y is unity. Then the value of x would be quite definite, and equal to —0·08711. We shall call X_2 the value of x derived in this somewhat peculiar way. Row 1 of Table 1 is simply the coefficients of the various terms of Equ. (1) multiplied by X_2; row 2 of Table 1 is the coefficients of Equ. (2) in order as they stand. Row 3 is obtained by adding rows 1 and 2. The zero in the first column of row 3 is not an accident; it is our guarantee that we have obtained X_2 correctly. The quantities in the other columns of row 3 are given names q_2, s_{23}, s_{24} and s_{25} because they will be required at a later stage. There is also a check on the correctness of q_2 (row 3, column 2) which should be given by

$$q_2 = a_{22} - (a_{12}^2/a_{11}) \quad \ldots \quad (5)$$

where a_{11} denotes the coefficient of the first unknown x in Equ. (1), here 1, a_{22} the coefficient of the second unknown y in Equ. (2), here also 1, and a_{12} the coefficient of x in Equ. (2) [or of y in Equ. (1) since we have assumed the given equations to be symmetrical]; this last coefficient is here 0·08711; in the notation of Equ. (5),

$$X_2 = -a_{12}/a_{11} \quad \ldots \quad (6)$$

TABLE 1

Initial Stage of 'Escalator' Tabulations

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4	Column No. 5	Remarks
1	—0·08711	—0·007588	—0·037040	—0·032830	—0·024399	Coefficients of Equ. (1) × X_2
2	0·08711	1	0·16249	0·14094	0·23215	Coefficients of Equ. 2
3	0	0·992412 $=q_2$	0·125450 $=s_{23}$	0·108110 $=s_{24}$	0·207751 $=s_{25}$	Row 1 + Row 2

For the next stage we require to find the values X_3 of x and Y_3 of y which would satisfy Equs. (1) and (2) if z were replaced by 1, t by zero and the number terms were omitted. The trick for doing this is to calculate

$$X_3 = -(a_{13}/a_{11}) - s_{23}(X_2/q_2) \quad \ldots \quad (7)$$

$$Y_3 = -s_{23}/q_2 \quad \ldots \quad (8)$$

where a_{13} is the coefficient of the third unknown z in Equ. (1) or of the first unknown x in Equ. (3), here 0·42521. X_3 and Y_3 are in fact (see Reference 1) the values of x and y required; their numerical values for Equs. (1)–(4), obtained from Equs. (7) and (8), are —0·414199 and —0·126409 respectively.

We are now in a position to complete the four rows of Table 2, which are numbered 4 to 7 because they are really a continuation of Table 1 above.

Row 4 consists of the coefficients of Equ. (1) multiplied in turn by X_3 (derived from Equ. (7) and having here the value —0·414199) while row 5 consists of the coefficients of Equ. (2) multiplied in turn by Y_3 (derived from Equ. (8) and having here the value —0·126409). Row 6 is merely the coefficients of Equ. (3) copied in order, and row 7 is the sum of rows 4–6. The zeros in the first two columns of row 7 guarantee that X_3 and Y_3 are correct. If these zeros do not appear, the most likely trouble is an incorrect value of s_{23}; an incorrect value of X_2 or q_2 would have been detected earlier as already indicated. The value of q_3 can be checked from the alternative formula

$$q_3 = a_{33} - \frac{a_{13}^2}{a_{11}} - \frac{s_{23}^2}{q_2} \quad \ldots \quad (9)$$

where a_{33} is the coefficient of the third unknown z in Equ. (3), here 1, and the quantities named q_3, s_{34} and s_{35} will all be required later.

We are now ready to proceed to the next stage, for which we require the values X_4 of x, Y_4 of y and Z_4 of z

TABLE 2

Second Stage of 'Escalator' Tabulations

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4	Column No. 5	Remarks
4	—0·414199	—0·036081	—0·176122	—0·156103	—0·116013	Cfts. Equ. (1) × X_3
5	—0·011011	—0·126409	—0·020540	—0·077816	—0·029346	Cfts. Equ. (2) × Y_3
6	0·42521	0·16249	1	0·91117	—0·40936	Cfts. Equ. (3)
7	0	0	0·803338 $=q_3$	0·737251 $=s_{34}$	—0·554719 $=s_{35}$	Sum

TABLE 3

Orderly Arrangement of the Quantities X_2, Y_3, etc.

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4	Column No. 5
1	1	−0·08711	−0·414199		
2	0	1	−0·126409		
3	0	0	1		
4	0	0	0	1	

satisfying Equs. (1), (2), and (3) if t is taken as unity and the number term is omitted. The formulae for X_4, Y_4 and Z_4 are:

$$\left.\begin{aligned} -X_4 &= (a_{14}/a_{11}) + (s_{24}X_2/q_2) + (s_{34}X_3/q_3) \\ -Y_4 &= \qquad\qquad\quad (s_{24}/q_2) + (s_{34}Y_3/q_3) \\ -Z_4 &= \qquad\qquad\qquad\qquad\qquad\quad (s_{34}/q_3) \end{aligned}\right\} \quad .. \quad (10)$$

where a_{14} is the coefficient of the fourth unknown t in Equ. (1) or of x in Equ. (4), here 0·37688. All the other quantities in Equ. (10) are known, and therefore the determination of X_4, Y_4 and Z_4, like that of X_3 and Y_3 from Equs. (7) and (8), is perfectly straightforward. Nevertheless, it is extremely easy to make mistakes, and the arrangement of Table 3, which includes various quantities we have so far obtained, is helpful.

In Table 3 the number in row 1 and column 1 is always unity; the Table has one more column than the number of unknowns and the same number of rows as the number of unknowns. All rows of column 1 below row 1 contains zeros. The number X_2 is placed in row 1, column 2, and the number 1 in row 2, column 2 and in all other places where the row-number and column-number are equal; zeros occupy any position where the row-number is greater than the column-number. In column 3, the successive entries are X_3, Y_3 1 and 0, and X_4, Y_4, Z_4 will be entered in column 4, rows 1, 2 and 3 when they have been calculated from Equ. (13). The quantities to be entered in column 5 will be the respective values of x, y, z and t satisfying Equ. (1), and they will be derived by means of formulae similar to Equ. (13) to be obtained shortly.

It is convenient to tabulate separately (in Table 4) the quantities obtained by dividing column 1 of Table 3 by a_{11}, column 2 by q_2, column 3 by q_3, and so on; at this stage we only know the first three columns of Table 4 which gives the results of this division.

In Table 4 the number of rows and columns is the same as the number of unknowns; as column 5 of Table 3 is the solution, we do not need to proceed further and divide by a quantity to be called q_5, though this would be done if there were more equations and unknowns. Finally, we construct Table 5, which also has the same number of rows and columns as there are unknowns; at this stage we can complete all of it except the number s_{45} in the bottom right-hand corner. Column 1 is the coefficients of Equ. (1) in order, including the number-term but omitting a_{11}, the coefficient of x. Column 2 has a zero in the first row (and zeros likewise occupy all places in which the column-number exceeds the row-number) but the remaining entries are the quantities we have already determined in Table 1 and labelled s_{23}, s_{24} and s_{25}; the non-zero elements of column 3 of Table 5 are already determined in Table 2 and labelled s_{34}, s_{35}. Table 5 at this stage is thus

TABLE 4

Elements of Table 3 with first Column divided by a_{11}, second by q_2, third by q_3, and so on

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4
1	1	−0·087776	−0·515597	
2	0	1·007646	−0·157355	
3	0	0	1·244806	
4	0	0	0	

TABLE 5

Orderly Arrangement of Coefficients of Equ. (1), and the Quantities Named s_{28}, s_{34}, etc.

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4
1	0·08711	0	0	0
2	0·42521	0·125450	0	0
3	0·37688	0·108110	0·737521	0
4	0·28009	0·207751	−0·554719	

and the point of this tabulation is that Equs. (7), (8) and (10) can be handled most easily by aligning a row of Table 5 with a row of Table 4, multiplying numbers in the same column of the two chosen rows, adding the results of these multiplications and changing the sign of the sum. This operation can be performed on a calculating machine without recording anything except the numbers in Tables 4 and 5 and the final answer, and when a number of divisions by the same divisor are required, as in the columns of Table 4, time can be saved by multiplying by the reciprocal of that divisor which is itself an entry in Table 4. Thus, to obtain X_3 [Equ. (7)] take row 2 of Table 5 with row 1 of Table 4

$$-X_3 = 0{\cdot}42521 \times 1 + 0{\cdot}125450 \times (-0{\cdot}087776) \quad .. \quad (11)$$

and to obtain Y_3, take row 2 of Table 5 with row 2 of Table 4, so that

$$-Y_3 = 0{\cdot}42521 \times 0 + 0{\cdot}125450 \times (1{\cdot}007646) \quad (12)$$

Similarly, to obtain X_4 [Equ. (10)] take row 3 of Table 5 with row 1 of Table 4, so that

$$\begin{aligned} -X_4 = 0{\cdot}37688 \times 1 &+ 0{\cdot}108110 \times (-0{\cdot}087776) \\ &+ 0{\cdot}737521 \times (-0{\cdot}515597) \quad .. \quad (13) \end{aligned}$$

X_4 is thus 0·012734.

To obtain Y_4, take row 3 of Table 5 with row 2 of Table 4, which gives

$$-Y_4 = 0{\cdot}108110 \times 1{\cdot}007646 + 0{\cdot}737521 \times (-0{\cdot}157355) \quad .. \quad .. \quad (14)$$

so that Y_4 is 0·007074. The row in Table 5 has the number one less than the suffix of the quantity sought, thus it is the third row when we seek X_4, Y_4 or Z_4 and the second when we seek X_3 or Y_3. The row in Table 4 is 1 if we seek an X, 2 if we seek a Y, 3 if we seek a Z, and so on.

Having obtained X_4, Y_4 and Z_4, we can complete Table 6, which is a continuation of Tables 1 and 2, so that the rows are numbered from 8 upwards.

As at the corresponding point of earlier stages, we now have powerful checks on the calculation. If the first three columns of row 12 of Table 6 are not zero, there is an error in X_4, Y_4 or Z_4; the most likely cause of this

TABLE
Third Stage of 'Escalator' Tabulations

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4	Column No. 5	Remarks
8	0·012734	0·001109	0·005415	0·004799	0·003567	Cfts Equ (1) × X_4
9	0·000616	0·007074	0·001149	0·000997	0·001642	„ „ (2) × Y_4
10	−0·390230	−0·149123	−0·917734	−0·836212	0·375684	„ „ (3) × Z_4
11	0·37688	0·14094	0·91117	1	−0·26523	Cfts Equ. (4)
12	0	0	0	0·169584 $=q_4$	0·118363 $=s_{45}$	Sum

is that s_{24} or s_{34} is wrong, since an error in X_3, Y_3, q_3 or s_{23} would have been detected at an earlier stage. There is also a check on the value of q_4 analogous to Equ. (9), namely

$$q_4 = a_{44} - \frac{a_{14}^2}{a_{11}} - \frac{s_{24}^2}{q_2} - \frac{s_{34}^2}{q_3} \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

a_{44} being the coefficient of the fourth unknown t in the fourth equation. [Those familiar with determinants may like to know that the determinant of the coefficients of the unknowns of Equs. (1)–(4) is $a_{11}q_2q_3q_4$.] The numerators of the terms in Equ. (15) are in row 3 of Table 5 just as those of the terms in Equ. (9) are in row 2 of Table 5; the relevant row number in Table 5 is one less than the suffix of the q concerned. The quantity s_{45} is next written in the blank space (row 4, column 4) of Table 5. We are now ready to calculate the solution, which is

$$x = X_5, y = Y_5, z = Z_5, t = T_5 \quad \ldots \quad \ldots \quad (16)$$

where X_5, Y_5, Z_5 and T_5 are given by equations analogous to Equ. (10), namely

$$\left.\begin{aligned} -X_5 &= (a_{15}/a_{11}) + (s_{25}X_2/q_2) + (s_{35}X_3/q_3) + (s_{45}X_4/q_4) \\ -Y_5 &= (s_{25}/q_2) + (s_{35}Y_3/q_3) + (s_{45}Y_4/q_4) \\ -Z_5 &= (s_{35}/q_3) + (s_{45}Z_4/q_4) \\ -T_5 &= (s_{45}/q_4) \end{aligned}\right\} \quad (17)$$

a_{15} being the number-term in Equ. (1), here 0·28009. X_5, Y_5, Z_5 and T_5 can be obtained direct from Equ. (17), but it simplifies the calculation and avoids errors to complete Table 3 by writing X_4, Y_4 and Z_4 in column 4, and to complete Table 4 by dividing these quantities by q_4 and putting the results $X_4/q_4 = 0{\cdot}075090$, $Y_4/q_4 = 0{\cdot}041714$, $Z_4/q_4 = -5{\cdot}411677$ in the first three rows of column 4, and $1/q_4 = 5{\cdot}896782$ in the fourth row. X_5, Y_5, Z_5 and T_5 are then obtained by multiplying elements of the fourth row of Table 5 (now completed with the value 0·118363 of s_{45} in the fourth column) in turn by corresponding elements of the first, second, third and fourth rows of Table 4 (in which there is now a fourth column as just explained) summing the products and reversing the sign. To check the final solution, it is easiest to complete the first four columns only of Table 7, which is a continuation of Tables 1, 2 and 6 so that the rows are numbered 13–18. The zeros in row 18 are the final check; if they do not appear, the error is most likely to be in s_{25}, s_{35} or s_{45}, since errors in other quantities would have been detected earlier. Row 17 has the number-term of Equ. (1) in column 1, that of Equ. (2) in column 2, and so on.

APPENDIX
'Normalizing' Unsymmetrical Equations to make them Symmetrical

Suppose that the original unsymmetrical equations are

$$\left.\begin{aligned} a_1x + b_1y + c_1z &= d_1 \\ a_2x + b_2y + c_2z &= d_2 \\ a_3x + b_3y + c_3z &= d_3 \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (A1)$$

Multiply each of the equations (A1) by the coefficient of x in that equation and add; the resulting equation is

$$(a_1^2+a_2^2+a_3^2)x + (a_1b_1+a_2b_2+a_3b_3)y + (a_1c_1+a_2c_2+a_3c_3)z = a_1d_1+a_2d_2+a_3d_3 \quad \ldots \quad \ldots \quad (A2)$$

Now multiply each of the equations (A1) by the coefficient of y in that equation and add, to obtain

$$(a_1b_1+a_2b_2+a_3b_3)x + (b_1^2+b_2^2+b_3^2)y + (b_1c_1+b_2c_2+b_3c_3)z = b_1d_1+b_2d_2+b_3d_3 \quad \ldots \quad \ldots \quad (A3)$$

and, finally, multiply each of the equations (A1) by the coefficient of z in that equation and add, to obtain

$$(a_1c_1+a_2c_2+a_3c_3)x + (b_1c_1+b_2c_2+b_3c_3)y + (c_1^2+c_2^2+c_3^2)z = c_1d_1+c_2d_2+c_3d_3 \quad \ldots \quad \ldots \quad (A4)$$

The equations (A2), (A3) and (A4) are a symmetrical set equivalent to the original equations (A1).

The 'Escalator' process described in the main text can be adapted (see Reference 1) to unsymmetrical equations, but is simpler with symmetrical equations. It should, however, be noted that the checks on the arithmetic described in the main text are only effective on the assumption that Equs. (A2)–(A4) are correctly derived from Equ. (A1). Although the process of derivation is straightforward, mistakes are easily made, and therefore Equs. (A2)–(A4) must be very carefully checked in a numerical case before the 'Escalator' process is begun.

REFERENCE

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TABLE 7
Final Check of the 'Escalator' Solution

Row No.	Column No. 1	Column No. 2	Column No. 3	Column No. 4	Column No. 5	Remarks
13	−0·556754	−0·048499	−0·236737	−0·209829	Not Needed	Cfts. Equ. (1) × X_5
14	−0·026269	−0·301564	−0·049001	−0·042502		„ „ (2) × Y_5
15	0·565980	0·216284	1·331061	1·212823		„ „ (3) × Z_5
16	−0·263048	−0·098371	−0·635961	−0·697961		„ „ (4) × T_5
17	0·28009	0·23215	−0·40936	−0·26253		Number terms
18	0	0	0	0		Sum

Modified Two-Hole Directional Coupler

By W. Geoffrey Voss, B.Sc.*

SUMMARY. *This article describes a modification to the two-hole, interference type, co-directional coupler in a rectangular waveguide system which enables it to be tuned to give perfect directivity at any wavelength within a wide frequency range.*

Two distinct types of directional coupler are possible in a rectangular waveguide system and many articles have been devoted to their description. The first is the contra-directional type, which uses a single coupling element in the broad side of the guide and is directive due to the intrinsic directive property of the hole (Bethe-hole coupler[1]). The second is the co-directional, interference type, directive because of the phase opposition of waves coupled through two holes in the side wall of the waveguide. Both suffer from the same limitation that they are perfectly directive at only one wavelength, the design wavelength.

This article is concerned only with the latter type and it will be shown that in this case the limitation can be overcome by the introduction of two phase shifters, placed between the two coupling elements, one in each guide.

The Modified Coupler

Fig. 1 shows two waveguides joined by two coupling elements of equal size in their narrow side and placed a distance S apart. A and B are two phase shifters.

With A and B set to give zero phase shift, the system becomes theoretically perfectly directive when

$$S = \lambda_g/4$$

where λ_g is the guide wavelength; i.e., when the phase difference of the two waves, due to the hole spacing, is

$$(2n - 1) . \pi/2, \text{ where } n = 1, 2, \ldots . \text{ etc.}$$

If the phase shifters are set to arbitrary values θ_1 and θ_2 in the primary and secondary guide respectively, the condition for the backward wave to cancel is

$$2\psi + \theta_1 + \theta_2 = (2n - 1)\,\pi \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

where $n = 1, 2, \ldots\ldots$ etc.

The phase relation of the waves coupled in the forward sense by the elements, considered at some plane beyond the elements, is

(a) $\psi + \theta_2$ for the wave coupled through (X)
(b) $\psi + \theta_1$ for the wave coupled through (Y)

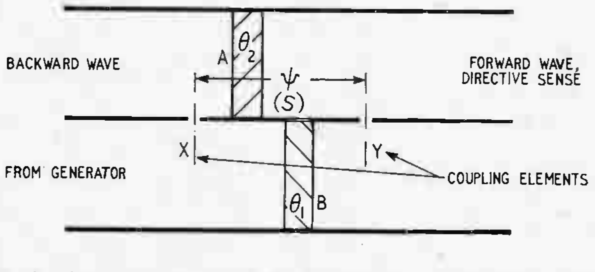

Fig. 1. *Schematic diagram of the modified two-hole directional coupler. A and B are phase shifters and the coupling elements are placed a distance S apart. (The system is assumed to be matched at all four terminations)*

Thus the forward waves add in phase if

$$\theta_1 = \theta_2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

The theoretical condition for perfect directivity then becomes

$$\psi + \theta = (2n - 1)\,\frac{\pi}{2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

Thus, if the distance S is chosen to give a phase shift ψ_0, and at this spacing the coupler is perfectly directive at the shortest wavelength required, θ can be increased from zero to θ_{max}, to give the maximum wavelength at which perfect directivity is required.

For most applications it will be seen that both phase shifters need to be calibrated. However, with a perfect travelling wave in the primary guide, only one phase shifter has to be calibrated; the other can be simply tuned to give maximum output in the directive sense of the secondary guide.

Rosen and Bangert[2] have shown that a simple coupler of the interference type is directive at a wavelength slightly longer than the theoretical value, due to the effect of interaction between the coupling holes. In the simple coupler, other practical difficulties arise from mechanical errors in the hole diameters and their spacing. These errors cease to be of importance with the modified coupler, as any error can then be compensated for by the phase shifters.

It has also been shown[2] that as the elements are spaced further apart ($n = 2$, 3, etc.) the bandwidth of the resulting directivity curve decreases considerably. As the introduction of the phase shifters between the elements necessitates their greater spacing, the system will be directive over a narrow bandwidth. This effect is of advantage as it makes the tuning system more sensitive.

It will be seen that this modification of the interference coupler makes its operation at any wavelength within the frequency band of the particular system possible, when used for the measuring devices described by Mumford[3]. With both the phase shifters calibrated, the system can be used as a wavemeter. In all cases, the accuracy and sensitivity of the coupler will depend on the accuracy of the calibration of the phase shifters and the hole spacing used.

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High-Q Echo Boxes

DESIGN OF TUNABLE CYLINDRICAL TYPES

By A. Cunliffe, B.Sc., Ph.D., A.R.C.S., and R. N. Gould, M.Sc., Ph.D.*

SUMMARY. *The performance of tunable H_0 cylindrical echo boxes may deteriorate sharply when the length of the cavity is approximately that for which, theoretically, another mode has the same resonant frequency as the operating mode. The cause of this effect is investigated, and the modes for which the effect is likely to be greatest are named. Cavities should be designed so that the frequency-length curves for these particular modes do not cross the frequency-length curve for the operating mode.*

Cavities used for supplying artificial echoes to radar sets are known as echo boxes. The transmitter pulse introduces energy into the cavity and the resulting oscillations in the cavity persist after the transmitter pulse, decaying exponentially. Energy can therefore be fed from the cavity to the receiver after the transmitter pulse has ended, giving effectively an artificial echo which can be used for general testing and tuning of the radar installation. The exponential decay of oscillations within the cavity is more gradual the higher the Q value of the cavity, and to obtain a usable artificial echo, the Q value must be so high that it is necessary to use higher-order modes of oscillation rather than the mode with the lowest frequency of oscillation. Cylindrical cavities operating in the H_{01n} modes are normally used, where n is 1 or 2 for wavelengths in the ten-centimetre region, or 4 to 8 for wavelengths in the three-centimetre region.

These H_{01n} cavities are normally tuned by having one of the plane ends adjustable, moving like a piston in a cylinder. The piston does not need to make actual contact with the cylinder wall since there is no flow of current between a plane end and the curved cylinder wall in an H_{01n} resonator. It is difficult to ensure that the piston and the other plane end of the cavity remain accurately parallel as the cavity is tuned, and this lack of parallelism can give rise to a decrease in Q value of the H_{01n} resonator. This decrease in Q is associated with the coincidence in the natural frequencies of the H_{01n} and E_{11n} modes of oscillation of the perfectly cylindrical cavity[1,2]. The reduction in Q value associated with this coincidence in frequencies can be effectively eliminated by making either the piston or the other end of the cavity slightly spherical, rather than perfectly plane. Alternatively, a circular groove, with the centre of the circle on the axis of the cavity can be cut either in the piston or the other plane end. These deformations from the perfectly cylindrical shape cause the frequencies of the H_{01n} and E_{11n} modes to separate, and the frequencies remain separated when the cavity length is altered by adjustment of the piston position.

With the frequencies of the H_{01n} and E_{11n} modes so separated, there may be certain tuning positions near which the Q value is very much reduced. Such reductions in Q value are associated with the coincidence in frequencies of the H_{01n} and other modes of oscillation of the undistorted cavity. With the high-order modes needed for echo boxes, and particularly for echo boxes operating in the three-centimetre region of wavelength, if the echo box is to be tunable over a reasonable frequency band, the frequency-length curve for the operating mode must theoretically cross the frequency-length curves of other modes. It is found experimentally that the amount of reduction of the Q value, and the tuning range over which the reduction is effective, depend on the mode which 'crosses' the operating mode, on how much the cavity departs in shape from the perfect cylinder, and on whether the coupling system couples to the crossing mode as well as to the operating mode. In general, the reduction in Q value and the tuning range over which the reduction is important, are both least when the cavity does not depart appreciably from the perfect cylindrical shape and when the coupling system does not couple to the crossing mode.

Cause of the Decrease in Q Value

Formulae for the frequencies and for the electric and magnetic-field configurations of the modes of natural oscillations of perfectly cylindrical cavities are given in the standard text books[3]. If a cavity is distorted from the perfectly cylindrical form, either by deforming the walls or by coupling another circuit to the cavity (e.g., a waveguide circuit may be coupled to the cavity via an iris in the cavity wall), the modified cavity will have a different set of frequencies and field configurations for its modes of natural oscillation. If the distortion is only slight (i.e., if the walls almost form a cylinder, or the coupling to the external circuit is loose), the modified cavity will have modes with frequencies which do not

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differ very much from the frequencies of the perfectly cylindrical or undistorted cavity. It might also be expected that the field configurations of the modified cavity will be very much like the field configurations of the modes of the perfect cylinder, important differences appearing only in parts of the cavity close to where the distortions are introduced (near where the wall is deformed from the perfectly cylindrical shape or near the iris). Corresponding to a mode of the undistorted cavity which has a frequency well separated from the frequencies of other modes, there is indeed a mode in the modified cavity with a slightly changed frequency and a field configuration very similar to that of the mode in the perfectly cylindrical cavity. Consider, however, two modes of the undistorted cavity which have frequencies close together. The modified cavity, if the distortion is only slight, will have two modes of natural oscillation with frequencies close to the frequencies of the two modes of the undistorted cavity. It will be seen later that with some types of distortion the field configurations of these two modes of the modified cavity will not be similar respectively to the field configurations of the two separate modes of the undistorted cavity. Each of the two modes of the modified cavity may have field configurations which are similar to what would be obtained by mixing the field configurations of the separate modes of the perfectly cylindrical cavity, in different proportions for the two modes.

A cavity which is not provided with a slightly spherical end or a circular groove to separate the frequencies of the H_{01n} and E_{11n} modes, and which has non-parallel ends, for example, will not have a mode of natural oscillation with a field configuration similar to that of the H_{01n} mode alone. If the distortion is small, there will be modes of oscillation with frequencies close to the frequency of the H_{01n} or E_{11n} modes of a perfectly cylindrical cavity of the same length and radius, but these modes of the modified cavity will have field configurations which are effectively mixtures of the field configurations of the H_{01n} and E_{11n} modes of the undistorted cavity. If the modified cavity has a piston at one end, which does not make electrical contact with the curved cylinder wall, the modes of the modified cavity will have reduced Q values; for the part of the field configurations of the modes of the modified cavity, similar to the field configurations of the E_{11n} modes of the perfectly cylindrical cavity, will require current to flow between the plane ends and the curved cylinder wall, and lack of contact here will depress the Q values. The undistorted cavity has three modes with identical frequencies, the H_{01n} mode and two E_{11n} modes. Two E_{11n} modes with field configurations rotated with respect to each other through an angle of $\pi/2$ about the axis of the cylinder can be considered to be distinct modes. The modified cavity with non-parallel ends will have three modes with frequencies close to the common frequency of the three modes of the undistorted cavity, and the Q values of all three of these modes of the modified cavity will be reduced if there is a gap between the piston and the cylinder wall.

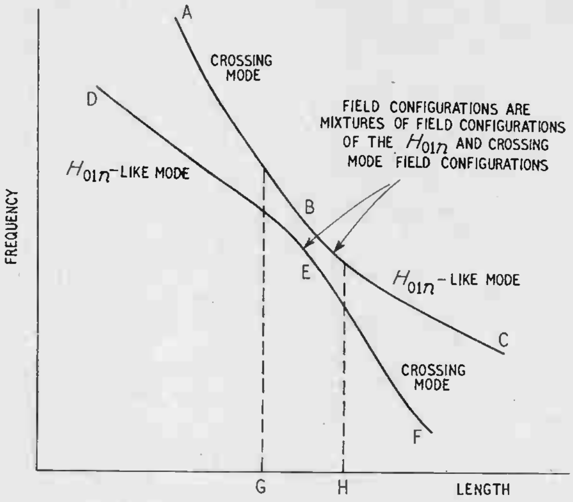

Fig. 1. *Frequency-length curves of the H_{01n}-like mode and the crossing mode*

A slightly spherical end, or a circular groove cut into an end with the centre of the circle on the axis of the cavity are distortions which have a different effect from the distortion where the ends of the cylinder are not parallel. With a slightly spherical end or circular groove distortion there will be three modes in the modified cavity with frequencies fairly close to the common frequency of the H_{01n} and E_{11n} modes of the perfectly cylindrical cavity. One of these modes will have a field configuration very much like the field configuration of the H_{01n} mode of the perfectly cylindrical cavity, the field configuration only being appreciably different near the slightly spherical end or groove. The other two modes will have a common frequency which is different from the frequency of the first mode, and these two modes will have field configurations very much like the field configurations of the E_{11n} modes of the perfectly cylindrical cavity. If there is a gap between piston and cylinder wall, the Q value of the H_{01n}-like mode will not be affected to any extent, but the Q value of the E_{11n}-like modes will be reduced. Now regard the cavity with a slightly spherical end or circular groove as the undistorted cavity and this cavity with an additional distortion where the ends become slightly non-parallel as the modified cavity. The undistorted cavity has an H_{01n}-like mode separated in frequency from two E_{11n}-like modes. If the additional distortion, where the end plates become slightly non-parallel, is not large compared with the slightly spherical end or circular groove distortion, the modified cavity will still have an H_{01n}-like mode, and the Q value of this mode will not be affected to any extent by a gap between piston and cylinder. This is because the H_{01n}-like mode of the undistorted cavity is separated in frequency from the E_{11n}-like modes. This frequency separation is not large, however, and if the modified cavity has ends which are appreciably non-parallel, there will not be a mode which can be regarded as being purely of the H_{01n}-like form.

The frequencies of the H_{01n}-like and E_{11n}-like modes of a cavity, which has only the slightly spherical end or circular groove distortions, remain separated when the cavity length is altered. But the frequency-length curves for other modes can intersect the frequency-length curve for the H_{01n}-like mode and at the lengths where the intersections occur, some other mode will have the same frequency as the H_{01n}-like mode. With certain additional types of distortion, for lengths near to the value for which in the absence of the additional distortion the H_{01n}-like mode and the other mode have the same frequency, it can happen that the modified cavity will have no pure H_{01n}-like mode. For such lengths the cavity may have modes with field configurations which are effectively mixtures between the field configurations of the H_{01n}-like mode and the other mode. The Q values of these modes will be small if there is no electrical contact between piston and cylinder and the field configuration of the mode which 'crosses' the H_{01n}-like mode requires current to flow across this gap. The modified cavity will have an H_{01n}-like mode for lengths which are not close to the length for which, without the additional distortion, the frequencies of the H_{01n}-like mode and the crossing mode coincide. With the additional distortion, the frequency-length curves of the H_{01n}-like mode and the other mode will not intersect. Instead, the effect illustrated in Fig. 1 will take place, where there are transitions near the points B and E between an H_{01n}-like field configuration and a field configuration characteristic of the crossing mode. It can be shown theoretically that the distances BE and GH are greater the greater the additional distortion.

Now consider a cavity which is distorted by having a waveguide circuit coupled to it. It will be seen in the next section that if, in theory, there is coupling between the waveguide circuit and both the H_{01n} and an E_{11n} mode, this type of distortion can act like the non-parallel end distortion. Similarly if, instead of there being coupling between the waveguide and the E_{11n} mode, there is coupling between the waveguide and both the H_{01n} and a crossing mode, the modified cavity may have no pure H_{01n}-like mode for lengths close to the length for which the frequencies of the H_{01n} and the crossing mode would coincide if there were no distortion. The modified cavity will, in fact, have modes with field configurations which are effectively mixtures of the field configurations of the H_{01n} and the crossing mode, for such lengths. The Q-values of these modes will be small if there is a gap between piston and cylinder wall and the field configuration of the crossing mode requires current to flow across this gap.

If the crossing mode is of the H_{0pq} type, it might be expected that a gap between piston and cylinder wall would not reduce the Q values appreciably, for a field configuration which is effectively a mixture of the H_{01n} and H_{0pq} modes would not require current to flow between piston and cylinder. As well as the H_{01n}-H_{0pq} crossing however, there will be H_{01n}-E_{1pq}, H_{0pq}-E_{11n} and E_{11n}-E_{1pq} crossings. The slightly spherical end or circular groove distortion will change the position slightly of, for example, the H_{01n}-E_{1pq} crossing, but cannot eliminate it. It is clear therefore that an additional distortion (e.g., non-parallel ends or coupling to a waveguide circuit by both the H_{01n} and an E_{1pq} mode), will lead to there being no pure H_{01n}-like mode near the crossing region, with consequent depression of Q values.

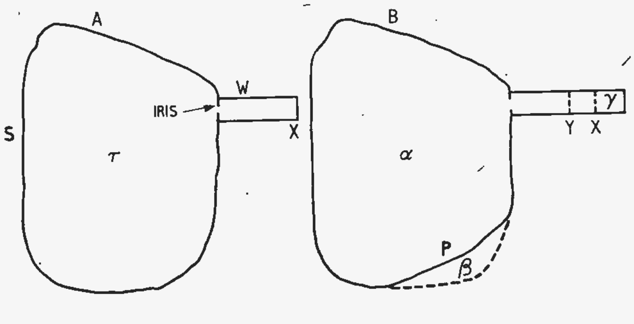

Fig. 2. *Illustrating the undistorted cavity A, and the modified cavity B*

The Field Configurations in the Modified Cavity

Expressions for the electric and magnetic field configurations of the modes of natural oscillation of a modified cavity can be obtained in terms of the modes of oscillation of the undistorted cavity by the use of perturbation theory. Several methods of applying perturbation theory to cavities have been described[1,2,4,5,6,7,8] and the results of a comprehensive account by Slater[9] will be used here.

Let A of Fig. 2 be the undistorted or unperturbed cavity. It will be supposed that the cavity consists of a homogeneous medium of permittivity ϵ and permeability μ throughout a region τ which is completely enclosed by perfectly conducting walls **S**. The waveguide W with the short circuit at X is considered to be part of the cavity, while the medium within the waveguide is the same as elsewhere in the cavity.

The electric and magnetic fields, within the cavity when it is oscillating only in a single mode, can be written respectively as

$$\mathbf{E} = A\mathbf{e}_n \sin(\omega_n t + B),\ \mathbf{H} = A\mathbf{h}_n \cos(\omega_n t + B) \quad (1)$$

where A and B are constants determining the amplitude and phase of the oscillations, the eigen-functions $\mathbf{e}_n$ and $\mathbf{h}_n$ are vector functions of position but not of time, and the eigen-value ω_n is a constant. The eigen-functions determine the electric and magnetic field patterns and the eigen-value determines the frequency of the particular mode of oscillation. The equations

$$\nabla \times \mathbf{e}_n = \mu\omega_n\mathbf{h}_n,\ \nabla \times \mathbf{h}_n = \epsilon\omega_n\mathbf{e}_n \quad \ldots \quad (2)$$

hold throughout the volume τ of the cavity, and at the boundary **S**,

$$\mathbf{e}_n \times d\mathbf{S} = 0 = \mathbf{h}_n \cdot d\mathbf{S} \quad \ldots \quad (3)$$

Let $\mathbf{e}_m$, $\mathbf{h}_m$ and ω_m be the eigen-functions and eigen-values of another mode. Suppose that ω_n and ω_m are equal, or nearly equal, and that these two eigen-values are well separated from all other eigen-values of the cavity. It will be supposed that the eigen-functions are normalized so that

$$\epsilon \int \mathbf{e}_n \cdot \mathbf{e}_m d\tau = \delta_{nm} = \mu \int \mathbf{h}_n \cdot \mathbf{h}_m d\tau \quad \ldots \quad (4)$$

Let B of Fig. 2 be the modified or perturbed cavity. B is formed from A by pushing in the wall of A through a small volume β leaving the volume α unchanged, and by adding an extra small volume γ to the waveguide circuit. Then this modified cavity will have two modes with magnetic field patterns which, except close to the distortions, are given approximately by the equations

$$\mathbf{h} = \text{constant}\left[\mathbf{h}_n + \left(\frac{\omega_m I_{mn}}{A-B}\right)\mathbf{h}_m\right],$$

$$\mathbf{h} = \text{constant}\left[-\left(\frac{\omega_n I_{nm}}{A-B}\right)\mathbf{h}_n + \mathbf{h}_m\right] \quad .. \quad (5)$$

The angular frequencies of these two modes are given approximately by the equations

$$\omega^2 = \omega_n{}^2 + A - B - \omega_n I_{nn},$$

$$\omega^2 = \omega_m{}^2 - A + B - \omega_m I_{mm} \quad .. \quad .. \quad (6)$$

respectively. In these equations

$$I_{ij} = \int_P \mathbf{e}_i \times \mathbf{h}_j \,.\, d\mathbf{S} - \int_Y \mathbf{e}_i \times \mathbf{h}_j \,.\, d\mathbf{S} \quad .. \quad .. \quad (7)$$

the first integral being taken over the surface separating the regions α and β and the second integral over a plane at Y such that the volume between Y and X is equal to γ. In both integrals $d\mathbf{S}$ is positive if it has the direction of the outward normal for the volume, $\alpha - \gamma$.

$$A = (\omega_n I_{nn} - \omega_m I_{mm} + \omega_m{}^2 - \omega_n{}^2)/2 \quad .. \quad (8)$$

$$B = \sqrt{A^2 + \omega_n \omega_m I_{nm} I_{mn}} \quad .. \quad .. \quad .. \quad (9)$$

It can be seen from Equ. (5) that the magnetic field patterns of the modes in the modified cavity will be mixtures of the magnetic field patterns of the modes in the undistorted cavity unless $I_{nm} = 0 = I_{mn}$. These integrals are each composed of two parts, one part arising from the distortion where the cavity wall is pushed in, and the other from the modification to the waveguide circuit. Both types of modification can cause the modes of the modified cavity to have magnetic field patterns which are effectively mixtures of the magnetic field patterns of the modes of the undistorted cavity. It can be seen that if either $\mathbf{e}_n = 0 = \mathbf{h}_n$ over Y or $\mathbf{e}_m = 0 = \mathbf{h}_m$ over Y, there will be no contribution to I_{nm} or I_{mn} due to the modification to the waveguide circuit; i.e., unless both modes couple to the waveguide circuit, a further addition to the waveguide circuit will not be the cause of the magnetic field patterns of the modes of the modified cavity being effectively mixtures of the magnetic field patterns of the modes of the undistorted cavity.

To avoid the reductions in Q value of tunable H_{01n} cylindrical cavities it is, therefore, desirable to design the coupling system so that there is no coupling to the modes which 'cross' the operating mode. Distortions from the perfectly cylindrical shape, where a portion of the cavity wall is pushed in slightly, will not be important if

$$\int_P \mathbf{e}_i \times \mathbf{h}_j \cdot d\mathbf{S} = 0 = \int_P \mathbf{e}_j \times \mathbf{h}_i \,.\, d\mathbf{S}$$

where i denotes the H_{01n} mode and j denotes the crossing mode. The most troublesome distortion in practice is that in which the ends of the cylinder are not parallel. In the next section, the values of the above integrals will be considered for this distortion for the different crossing modes, with a view to finding which crossings are most likely to lead to a lowering of the Q value.

Non-Parallel Ends and the Different Crossing Modes

Let $P_{ij} = \int_P \mathbf{e}_i \times h_j \,.\, d\mathbf{S}$. This integral can be re-written as $\int_\beta (\epsilon\omega_j \mathbf{e}_i \,.\, \mathbf{e}_j - \mu\omega_i \mathbf{h}_i \,.\, \mathbf{h}_j)\, d\tau$, the volume integral being taken throughout the volume β through which the wall is pushed in. If $\omega_i = \omega_j$, then $P_{ij} = P_{ji}$.

Consider a cylindrical cavity which is provided with some mechanism, such as a slightly spherical end to separate the frequencies of the H_{01n} and E_{11n} modes, and which is further distorted by tilting one of the plane ends, as in Fig. 3. If the distortion is small, the axial co-ordinate for the eigen-functions may be taken to be that of the end of the cavity for all points within the volume β. There are four types of crossing which involve the H_{01n} mode; H_{01n}-H_{0pq} crossings; H_{01n}-E_{0pq} crossings, H_{01n}-H_{rpq} crossings and H_{01n}-E_{rpq} crossings. Let i denote the H_{01n} mode and j the crossing mode, and suppose that the cavity length is such that $\omega_i = \omega_j$.

It is found[10] that P_{ij} does not vanish for H_{01n}-H_{0pq} crossings. By using Equs. (6) it can be shown that the effect illustrated in Fig. 1 occurs, the frequency-length curves for the two modes of the modified cavity not intersecting. This is an undesirable effect when the cavity is to be used as an echo box, but the Q values will remain high. There will be H_{01n}-E_{1pq} and H_{0pq}-E_{11n} crossings nearby, where, as will be seen below, the Q values will drop, and experimentally these might be confused with the H_{01n}-H_{0pq} crossing.

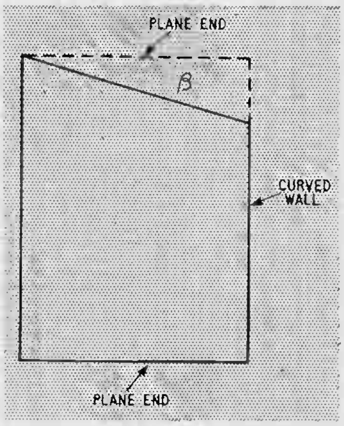

Fig. 3. *A distorted cylindrical cavity with non-parallel plane ends*

P_{ij} vanishes for the H_{01n}-E_{0pq} crossings, because the H_{01n} mode has no electric field and its magnetic field is at right angles to that of an E_{0pq} mode at the end of the cavity. The Q value of an H_{01n} cavity will, therefore, not be lowered at an H_{01n}-E_{0pq} crossing by this distortion.

There are two H_{rpq} modes, where $r \neq 0$, with field configurations which are rotated with respect to each other through a right angle about the axis of the cylinder. It is convenient to consider the field patterns of these two H_{rpq} modes lined up in directions which are parallel and perpendicular to the axis of tilt of the end plate. P_{ij} for one of the modes can then be shown to be of the form constant $\int_{\phi=0}^{\pi} \sin r\,\phi \sin \phi \, d\phi$, and for the other mode P_{ij} is of the form constant $\int_{\phi=0}^{\pi} \cos r\,\phi \sin \phi \, d\phi$. In the preceding section, formulae were written down for the field configurations of modes of the modified cavity when there were only two modes with frequencies close together in the undistorted cavity. The extension to this case, where there are three modes with frequencies close together, is straightforward. The performance of the echo box will be impaired at the crossing unless both values of P_{ij} vanish.

$\int_{\phi=0}^{\pi} \sin r\phi \sin\phi \, d\phi$ vanishes when $r > 1$ but not when $r = 1$.

$\int_{\phi=0}^{\pi} \cos r\phi \sin\phi \, d\phi$ vanishes when r is an odd integer, but not when r is even.

There are similarly two E_{rpq} modes when $r \neq 0$, and with the field patterns of these modes lined up as above the P_{ij} integrals for the E_{rpq} modes take the same form as for the H_{rpq} modes.

It follows that this type of distortion will lead to a deterioration in performance of the echo box at H_{01n}-H_{rpq} and H_{01n}-E_{rpq} crossings when r is unity and when r is even. The distortion will not lead to a deterioration in performance when r is an odd integer greater than unity.

Conclusions

H_{01n} tunable echo boxes should be provided with some mechanism (e.g., a slightly spherical end or a circular groove cut in one end) to separate the frequencies of the H_{01n} and E_{11n} modes in such a manner that the modes remain of separate H_{01n}-like and E_{11n}-like form. The coupling system should couple only to the H_{01n} mode and not to any crossing mode. The cavity should be designed so that there are no H_{01n}-H_{0pq} crossings and no H_{01n}-H_{rpq} or H_{01n}-E_{rpq} crossings where r is unity or even, in the operating frequency band.

Acknowledgements

The authors wish to thank Mr. C. M. Burrell, Mr. J. Fitzsimmons and Dr. L. Bailey, for initiating the work and for very valuable discussions during the progress of the work.

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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.

A Matter of Nomenclature

SIR,—In the Editorial of your October issue you refer to the difficulty of finding a suitable term to describe reproduced sound that is not stereophonic. "Monaural" is perhaps permissible on the ground that sound heard over a single channel sounds as if it were heard by one ear, although two ears are normally used; on the other hand, this is a hybrid word, the correct form of which would be "monotic". Alternatives that have been suggested are "monodic" (one channel), "monophonic" (one voice) and "platyphonic" (flat sound, as opposed to solid sound).

The only one of these that seems at all likely to catch on is monophonic. Its secondary meaning of one sound is inappropriate, but its primary sense of one voice seems to be just what we need—if we forget about tweeters. Monophonic also fits in quite well with the current, but rather ugly, abbreviation "mono".

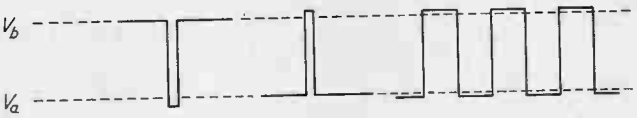
Ossulton Way,
London, N.2.
12th November 1958.

E. L. E. PAWLEY.

Oscilloscope Calibration

SIR,—With reference to the excellent editorial review of "Modern Oscilloscope Practice" in your Journal (June 1958, pp. 212–225) we would like to add a postscript to the paragraph on voltage measurements, to mention yet another source of error which is not commonly realized but which can be a serious nuisance. We refer to the fact that d.c.-coupled amplifiers have what may be argued as a somewhat lower gain for d.c. than a.c. input. The effect of this when applied to the Y axis of a cathode-ray tube is to cause, under certain conditions, errors in voltage measurement which may be as large as 10% of the voltage corresponding to maximum undistorted deflection.

One particularly noticeable effect is illustrated in the figure. Pulses of the same amplitude but differing in mark–space ratio appear at different levels on the screen of the cathode-ray tube. Taking this to its limit, a d.c. calibrating signal (infinite mark–space ratio) does not appear at the same level as a single short pulse of the same amplitude. The duration for which the d.c. calibrating signal is applied is, however, of importance. If, with the aid of a long-persistance tube and a very slow time-base, the deflection is observed from the time of its application, there will be an overshoot of approximately 10% decaying to the steady deflection with a time constant of the order of 1 second. The deflection immediately after the application of the calibrating signal appears to be the same as the deflection caused by the short pulse but the deflection a few seconds later, which is the level normally taken for calibration purposes, will be somewhat less. This response to a step input is characteristic of a network with a higher gain to a.c. signals than to d.c. signals.



100-c/s signals of equal amplitude but with large, small and unity mark–space ratios

The cause of this behaviour is not yet clear to us. Our first thoughts, unsupported by any experimental evidence, are that it may be due to a fall in emission for one of the many reasons discussed in literature [see, for example, A. S. Eisenstein, "Advances in Physics", Vol. 1, p. 1, (1948)], a change in working point due to thermal effect or even a progressive change in charge distribution inside the amplier valves when d.c. signal is applied.

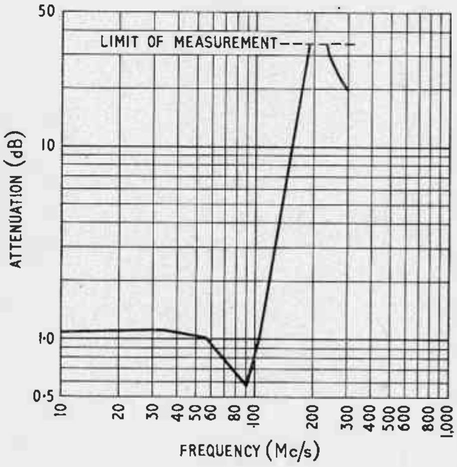
The error can be reduced—as has been done by a number of manufacturers—by the application of positive d.c.-coupled feedback with a large capacitor shunting it, so that the feedback is effective at d.c. falling off to a negligible level at very low frequencies. The primary purpose of this letter is to draw attention to the fact that a number of d.c. oscilloscopes have no correcting network and, with at least one notable exception, those that have incorporated it make no mention of this error, the purpose of even the existence of the correcting network and the possible need to adjust it.

The above described shift must not be confused with that which may be caused by a long pulse charging up a coupling capacitor to the trigger or sync circuit and discharging during the 'space' time through the output impedance of the source.

Cavendish Laboratory, H. V. Beck
University of Cambridge. C. W. McCutchen
26th November 1958. M. H. N. Potok

Miniature Delay Lines

Sir,—In an article which appeared in the October 1958 issue of your journal, Mr. Reinhold Gerharz has described some techniques for the manufacture of very-high-frequency delay lines. I should like to point out that quite satisfactory results at high frequencies may be achieved through the use of conventional techniques provided the characteristic impedance of the lines be made low enough. Under such conditions the distributed capacitance of the inductive elements is quite small in comparison with the total capacitance. Furthermore, since the required inductance is small, the line losses, which are generally attributable to undesirable characteristics of the inductors, are minimized.



The figure illustrates the frequency response characteristics of a typical line. The total delay is 100 millimicroseconds, and the impedance is 20 ohms. A hybrid type of construction is employed where the inductor is distributed and the capacitances are lumped. The line comprises 100 sections; the total length is 24 inches, the width and thickness, ¾ inch. The pulse response is excellent, with little stretching.

Epsco Inc., R. M. Zilberstein.
Boston,
Massachusetts, U.S.A.
14th November 1958.

New Books

Control System Components

By J. E. Gibson, Ph.D. and F. B. Tuteur Ph.D. Pp. 493 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 93s.

This book is derived mainly from a post-graduate course on control-system components given at Yale University and from special courses given to mechanical and electrical engineers. It is more of a text-book than a handbook and each chapter ends with a set of problems for the interested reader to solve.

The thirteen chapters may be grouped into four sections, 1–7, 8–9, 10–11, 12–13 concerned with electrical, mechanical, hydraulic and pneumatic components respectively.

Chapter 1, after a short description of types of resistors, capacitors and inductors, deals with network theory. Approximate and exact methods for constructing a network to have a given transfer function (i.e., ratio of output to input voltage) are given and a numerical example is solved. The chapter concludes with a discussion on the reliability of electrical components.

Chapter 2, on d.c. amplification, contains sections on familiar valve circuitry and also analyses the transistor d.c. amplifier. The 'chopper' type of d.c. amplifier, in which modulation to some form of a.c. is used before amplification, is described rather briefly. The final sections revert to the design of four types of valve d.c. amplifier.

Hard-valve and transistor power amplifiers are noted in Chapter 3 which is devoted mainly to thyratron and magnetic power amplifiers. The sections concerned with the latter give a useful account of their action and limitations.

Chapter 4 provides a detailed account of d.c. motors and generators and their control gear as used in control systems.

Chapter 5 deals with synchros, defined as devices to convert angular positions of a rotating shaft into corresponding electrical signals.

Modulators and demodulators are the topics discussed in Chapter 6. These are necessary where transmission of information is by a.c., as, for example, in the 'chopper' type of d.c. amplifier. Demodulation is not necessary if use is made of the a.c. motors described in Chapter 7.

Although there are many applications in which electronic devices are and will remain supreme, mechanical devices are sometimes simpler and more satisfactory. Chapter 8, after a short introduction covering the analogy between electrical and mechanical systems, is concerned with the use and design of gear trains. Chapter 9 is, I feel, rather muddled and, after more about gears and a mention of strain gauges, is devoted in the main to the gyroscope.

Chapters 10 and 11 are both on hydraulic systems, the former with pump-controlled and the latter with valve-controlled types. The theory of hydraulic transmission lines shows a considerable similarity to the better-known electrical lines while there is a corresponding connection between hydraulic and electronic valves.

The final chapters deal with pneumatic systems. The analogy between pneumatic and electrical components is emphasized and examples are given of several pneumatic devices.

The book is well produced and is complete with author and subject indexes. Having read this book, I wish it had been in existence sixteen years ago when, freshly down from college, I was required to help develop servo-mechanisms for aircraft instruments.

G.A.M.

Introduction to Nonlinear Analysis

By W. J. Cunningham. Pp. 349 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 74s.

This book is one of the McGraw-Hill Electrical and Electronic Engineering Series of some fifty volumes. The author is Professor of Electrical Engineering at Yale University, and he has used the subject matter for several years for a course to graduate students.

It is not an easy subject, but we are forced sometimes to face the fact that a system of linear equations, or other equations for which an exact analytical solution is known, is inadequate as a description of a practical physical system. One of the main difficulties is that the principle of superposition does not apply to nonlinear equations.

Only techniques requiring pencil, paper, slide rule and perhaps a desk calculating machine are considered. Electronic computers can advantageously be used for solving nonlinear problems, but discussion of them is outside the scope of this book.

Chapter 2 summarizes numerical methods which are very generally applicable. If an analytical solution is available, even approximately, this is superior to a numerical solution, which requires separate determination for each particular relevant set of values of any parameters involved. Nevertheless, if the independent variable is given a sufficiently small increment, h, a solution for a particular numerical case can be found step by step by using Taylor's series, or modifications thereof, to turn the original differential equation into a difference equation; the value of the dependent variable y is evaluated successively for values x_0, $x_0 + h$, $x_0 + 2h$, . . . $x_0 + nh$ of the independent variable. If h is small, errors can usually be kept small, but n, the number of steps required in the calculation, is correspondingly increased. If however h is large, the amount of calculation is reduced at the expense of accuracy.

In Chapter 3, graphical methods, notably the method of isoclines [lines of equal slope], are considered. It is often extremely useful to get an approximate idea of the nature of the solution by such methods even if subsequently more accurate determination is necessary by numerical or analytical methods. Starting from any point, the system of isoclines determines approximately the slope of the solution-curve there and, hence, a neighbouring point on the solution-curve-isoclines cannot intersect except at singular points, which are discussed in Chapter 5; they are points of stable or unstable equilibrium of the system. Chapter 4 deals briefly with equations having known exact solutions; the general technique is where possible to reduce the given nonlinear system to a linear system or other system having a known exact solution and approximately applicable over a certain range. Chapter 6 deals with methods such as variation of parameters and perturbation methods and the principle of harmonic balance; either the nonlinear terms are here regarded as small, or the discrepancies from an equation having a known analytical solution are small.

Systems driven by a simple-harmonic forcing function are considered in Chapter 7. In Chapter 8 differential-difference equations are introduced; these apply to systems in which fixed time-delays exist. Chapter 9 is concerned with linear differential equations having variable coefficients, notably the Mathieu equation, and in Chapter 10 stability is discussed. A bibliography having 81 references is included; their relevance to the various chapters is briefly indicated. This bibliography is not intended to be exhaustive, but is rather "a collection of material which has been found useful in preparing the manuscript".

As already indicated, the subject is not easy, and should not in the reviewer's opinion be attempted by an engineer who distrusts and dislikes mathematics, because it is inevitably mathematical. But for the minority who are prepared to face the subject, Professor Cunningham's book is helpful, and could guide a graduate engineer with a sufficient aptitude for and enjoyment of mathematics to a proper appreciation of how nonlinear problems can be tackled.

J.W.H.

Space-Charge Waves and Slow Electromagnetic Waves

By A. H. W. Beck, B.Sc.(Eng.), A.M.I.E.E. Pp. 396 + xi. Pergamon Press Ltd., 4 & 5 Fitzroy Square, London, W.1. Price 90s.

As a result of extensive research in the field of microwave engineering, if only in radar and guided weapons, it is not surprising that many new books dealing with the theory and practical applications of microwave electronics are being published. This book, which is the eighth volume in Pergamon's International Series of Monographs on Electronics and Instrumentation, provides the reader with an analytical treatment of the operation of microwave valves, the underlying theory being based on the interaction of two wave systems. This theoretical approach is derived partly from a knowledge of Maxwellian electrodynamics from which the idea of space-charge waves being used to interact with slow circuit waves is developed.

A general introductory chapter describing qualitatively the types of velocity-modulated valves now in existence is followed by a chapter on Maxwellian electrodynamics.

Slow-wave structures and space-charge theory are dealt with in the next two chapters followed by a chapter on the matching of input conditions with space-charge waves.

The reader, satisfied that he is fully conversant with the space-charge wave theory, can proceed to the next four chapters which apply this theory to various types of microwave valves, including klystrons, travelling-wave amplifiers, backward-wave oscillators and carcinotrons.

The latter part of the book is devoted to a chapter on the noise phenomena in space-charge wave devices, and the book ends with twelve appendixes dealing with measurements on circuits, the theory of the maintenance of electron beams, Llewellyn's electronic equations and several mathematical analyses.

The treatment is of a highly mathematical character and is likely to appeal only to those specialist engineers who are both deeply interested in the subject and are at least half mathematicians. The author admits in his preface that he has previously been accused of using too advanced mathematics. He says that this is incorrect. Perhaps it is. If one has vector analysis, the wave equations, Bessel functions and determinants so much at one's finger-tips as to be second nature one can read the book. None of these things is in itself very difficult and the charge should perhaps be that the author uses too much mathematics rather than too advanced. After all, a book in which the story was told mainly in school algebra could be understood by any engineer, but it would still be a difficult one.

B.J.M.

Vacuum-Tube and Semiconductor Electronics

By Jacob Millman. Pp. 664 + ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 77s. 6d.

The main object of this book is to provide the physics and electrical engineering undergraduate with a fully-integrated treatment of the theory and practical applications of valves and semiconductors.

Chapters 1 and 2 deal with the basic concepts and applications of the motion of charged particles in electric and magnetic fields.

The following chapter considers the electronic theory of metals and semiconductors and, in the section dealing with the energy distribution of metals, the author makes a useful analogy between the energy distribution function and the distribution function in age of the people in the U.S.A. to ensure that the reader has grasped the full significance of the energies possessed by electrons in a metal.

Valve-diode and semiconductor-diode characteristics are treated in Chapters 4 and 5, followed by a chapter on the photoelectric theory which shows how photo devices can be used in circuits.

Valve-triode characteristics, triode linear equivalent circuits, transistor characteristics, multi-electrode valves and h.f. transistors are dealt with in Chapters 7 to 11.

The remaining eight chapters are devoted to electrical discharges in gases, commercial gas tubes, rectifiers, untuned voltage amplifiers, audio power amplifiers, feedback amplifiers, sinusoidal oscillators and power supplies.

Throughout the text much attention has been devoted to the inclusion of worked examples and a bibliography of references is provided at the end of each chapter.

The latter part of the book includes appendixes on constants, conversion factors, etc., and the book ends with nearly 600 numerical and theoretical problems (unfortunately without answers) for the reader to solve.

Regardless of the fact that a great deal of the subject matter has been dealt with in an earlier book "Electronics" by the author and Dr. S. Seely, this book can be recommended as a suitable textbook for the electronics student, or even as a work of reference for the professional engineer.

B.J.M.

The Story of the Ionosphere

By J. A. Harrison, M.A., M.Ed., Ph.D. Pp. 103. Hulton Educational Publications Ltd., 161-166 Fleet Street, London, E.C.4. Price 10s. 6d.

The alternative title "or Exploring with Wireless Waves" is better indicative of the contents of this book than the main one. Two chapters only of the eight deal with the ionosphere. In the main, the book is a popular account of the history of wireless, starting with the nature of waves, going on to an account of Hertz's work and Marconi's transatlantic communication, and ending the first part with the Fleming diode and a brief mention of the triode. Then comes the ionosphere part, followed by chapters on radar and radio astronomy. It is difficult for one who knows a good deal about the subject to judge how such a book will appeal to one who knows nothing about it. However, it is felt that the book may well interest non-technical people and give them some idea of how wireless works.

W.T.C.

Notes on Applied Science No. 19

Pp. 16. Published for D.S.I.R. by Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 1s. 8d., post paid.

Deals with the calibration of attenuators, signal generators, voltmeters and ammeters at radio frequencies.

ABSTRACTS AND REFERENCES INDEX

The Index to the Abstracts and References published throughout 1958 is in course of preparation and will be available with the March 1959 Issue. As usual, a selected list of the journals scanned for abstracting, with publishers' addresses will be included.

NEW BRIT. I.R.E. PRESIDENT

Professor E. E. Zepler, Ph.D., has been elected President of the British Institution of Radio Engineers. Professor Zepler came to this country from Germany in 1935 and joined Marconi's Wireless Telegraph Company at Chelmsford where he remained until 1940.

After lecturing for two years at University College, Southampton, and a further three years at Cavendish Laboratory, Cambridge, he returned to Southampton as head of the electronics department. Since 1949 he has occupied the Chair of Electronics at Southampton University.

Several of Professor Zepler's articles have been published in *Wireless Engineer*, the first on "Oscillation Hysteresis in Grid Leak Detectors" appearing in the August 1946 issue.

SYMPOSIUM ON MILLIMETRE WAVES

The Microwave Research Institute of the Polytechnic Institute of Brooklyn will hold its ninth international symposium on "Millimetre Waves" in the Auditorium of the Engineering Societies Building, 33 West 39th Street, New York City, U.S.A., from 31st March to 2nd April 1959 inclusive.

NEW DATES FOR TRANSISTOR CONVENTION

The International Convention on Transistors and associated semiconductor devices, organized by the I.E.E., will be held at Earls Court, London, from 21st to 27th May 1959, and not on the dates previously announced.

An exhibition, covering the manufacture, application, research, development and other aspects of transistors and associated semiconductor devices, will also be held at Earls Court, and will remain open during the period of the Convention.

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

CONVENTION ON THERMONUCLEAR PROCESSES

The Institution of Electrical Engineers is organizing a convention on thermonuclear processes to be held in London on the 29th and 30th April 1959. Further details will be announced.

NEW JOURNAL ON DATA PROCESSING

The new quarterly journal *Data Processing*, published by Iliffe & Sons Ltd., answers the demand from management and technical executives for a comprehensive survey, with accurately-assessed data, on all modern methods of business and industrial automatic control.

The cost of a 12-months' subscription is £4, and only £8 for 3 years. In the U.S.A. and Canada the rates are $12 and $24.

MEETINGS

I.E.E.

5th January. "The Application of Transistors to Line Communication Equipment", by H. T. Prior, B.Sc., D. J. R. Chapman, B.Sc. and A. A. M. Whitehead, B.Sc.

6th January. "Silicone Electrical Insulation", by J. H. Davis.

19th January. "High-Current-Density Thermionic Emitters", by A. H. W. Beck, B.Sc. (Eng.).

20th January. "D.C. Amplifiers", discussion to be opened by K. Kandiah, M.A.

21st January. "Dielectric Materials—Trends and Prospects", by C. G. Garton.

22nd January. "Subscriber Trunk Dialling", by D. A. Barron, M.Sc.

26th January. "Automation", Faraday lecture by H. A. Thomas, D.Sc., at 6 p.m. at Royal Festival Hall, by ticket only.

29th–30th January. "Long-Distance Transmission by Waveguide". Forms for registering for this Convention obtainable from Secretary.

These meetings will be held at the Institution of Electrical Engineers, Victoria Embankment, London, W.C.2, and will commence at 5.30, except where otherwise stated.

Brit. I.R.E.

23rd January. Inaugural meeting of the Institution's Medical Electronics Group. An Address will be given by Professor A. V. Hill, C.H., M.A., Sc.D., F.R.S.

28th January. "Theoretical Aspects of Mechanical Speech Recognition", by Professor D. B. Fry, Ph.D.; "The Design and Operation of the Mechanical Speech Recognizer at University College, London", by P. Denes.

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

Television Society

9th January. "A Television Link with America?" discussion meeting to be held at Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2., at 7 p.m.

22nd January. "Modern Optics in Relation to Television", Fleming Memorial Lecture by G. H. Cook, F.B.K.S., to be held at 7 o'clock at the Royal Institution, Albemarle Street, London, W.1.

Society of Instrument Technology

14th January. "Data Storage Media", Symposium.

27th January. "Flow Measurement", Symposium.

Both these meetings will be held at Manson House, Portland Place, London, W.1, and will commence at 6 o'clock.

Institute of Navigation

16th January. "Blind Landing Problems", by W. J. Charnley, at 5.15 at The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.

Women's Engineering Society

30th January. "Some Uses of Electron Beams in Metallurgy", by Mrs. I. H. Hardwich, M.A., A.M.I.E.E., A.Inst.P., at 7 o'clock at Hope House, 45 Great Peter Street, Westminster, London, S.W.1.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for November* 1958

Date 1958 November	MSF 60 kc/s 2030 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	− 1	N.M.
2	− 1	N.M.
3	− 1	− 2
4	− 1	− 2
5	− 1	− 1
6	− 1	N.M.
7	0	− 1
8	0	N.M.
9	0	N.M.
10	0	0
11	0	0
12	0	+ 1
13	0	+ 1
14	0	+ 1
15†	0	N.M.
16	0	N.M.
17	0	+ 2
18	0	+ 2
19	0	+ 2
20	0	+ 2
21	0	+ 3
22	0	N.M.
23	0	N.M.
24	0	+ 3
25	0	+ 4
26	0	+ 4
27	0	+ 4
28	0	+ 4
29	0	N.M.
30	0	N.M.

**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.*

†*From the* 15*th November MSF* 60 *kc/s resumed transmission at* 14·29–15·30 *G.M.T.*

New Products

Crystal Clock

A portable transistorized crystal clock type TSA33 has been produced by Venner Electronics Ltd. The instrument incorporates a Venner 50-c/s synchronous clock with seconds hand. If desired, the clock can be omitted and replaced by a 12-V 50-c/s synchronous motor.

Two models are available which operate at ambient temperatures ranging from 15 °C to 25 °C and from 34 °C to 50 °C respectively. Each model provides standard output frequencies of 10 kc/s, 1 kc/s, 100 c/s and 50 c/s, and a frequency accuracy within 1 sec per week is claimed.

Provision is made for mains, accumulator, or dry-battery operation.

Venner Electronics Ltd.,
Kingston By-Pass, New Malden, Surrey.

Tape Cartridge

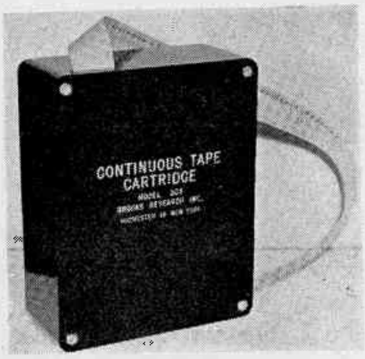
A continuous tape cartridge, type 203, designed to accept up to 100 feet of paper or plastic-punched tape for computer, automatic-test equipment, and similar applications, has been produced by Brooks Research Inc., U.S.A. The makers state that the cartridge obviates the need for rewinding tape at the end of a programme, as the tape can be spliced at the ends to form a continuous loop.

No mechanical connection to the tape driving mechanism need be made, power for the tape motion being derived from the tape reader itself.

Mounted in a phenolic container, the unit may be attached to the tape-reader panel by two simple slide fasteners.

The cartridge weighs 22 oz, measures 6⅞ in. × 5⅜ in. × 2⅜ in., and is said to be capable of storing up to 12,000 codes.

Brooks Research Inc.,
Post Office Box 67, *Rochester* 10, *N.Y., U.S.A.*



Stabilized Power Supply

A new power supply, type 41, designed primarily for incorporating in scientific equipment and instruments, as well as being a general purpose laboratory unit, is available from Automa Engineering Group Ltd.

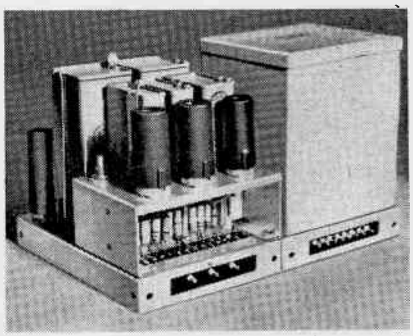
The circuit employs a grid-controlled series system with a neon reference tube, fluctuations being fed back through a high-gain d.c. amplifier to the control valve.

Details from the makers' specification include:—

Input 200–250 V in 10-V steps, 40–60 c/s, with permissible variation of ± 7%.
Output 250 or 350 V, 0–100 mA, ± 1%.
Output impedance d.c. of 0·1 Ω.
Noise and ripple level less than 250 μV.

The equipment comprises two units which may be assembled in two ways, giving alternative sizes 5 in. × 14 in. × 6½ in. or 10 in. × 7 in. × 6½ in.

Automa Engineering Group Ltd.,
Cherry Tree Rise, Buckhurst Hill, Essex.

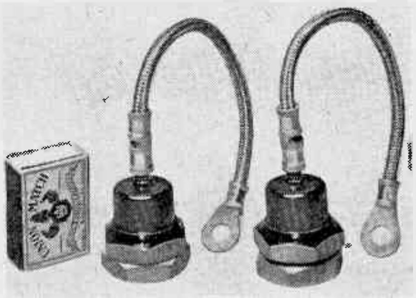


Silicon Rectifiers

A number of silicon rectifier devices have been announced by Standard Telephones & Cables. Brief descriptions of the components and some details from the makers' advance specifications are given below.

High-Power Silicon Rectifiers (Illustrated)

This new series, Type RS8, of five high-power silicon rectifiers incorporates an alloyed silicon junction in a stud-ended hermetically-sealed case. Intended for use with a heat sink, the rectifiers are claimed to have a maximum d.c. output current rating (half-wave resistive load) of 100 A at a maximum rated stud temperature of 100 °C. The five rectifiers are rated in 50-V steps from a maximum p.i.v. of 50 V to 300 V inclusive (up to 100 °C) and have a forward voltage drop, at 100 A and 100 °C of only 1.2 V. At the rated p.i.v., the maximum reverse current is stated to be 50 mA at the maximum junction temperature and 25 mA at 25 °C. The maximum mean dissipation (with a suitable heat sink) is of the order of 150 W.



Silicon Junction Power Rectifiers

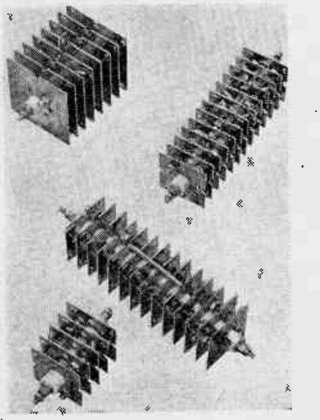
This series, designated RS5, comprises six silicon power rectifiers which incorporate a diffused silicon junction in a stud-ended hermetically-sealed case. They are suitable for either single or polyphase connection in equipments where a cooling surface is available.

The six rectifiers have p.i.v. ratings ranging from 50 V to 400 V and, depending upon the heat sink and cooling method employed, are capable of providing a maximum d.c. output current of 5 A at ambient temperatures up to 100 °C.

Max. forward voltage drop at 5 A d.c. (25 °C)	.. ..	1.3 V
Max. reverse current at p.i.v.	25 °C	100 μA
	100 °C	750 μA
Max. surge current (peak)	..	100 A
Max. mean dissipation at 100 °C		7.5 W
Max. storage temperature	..	150 °C
Min. storage temperature	..	−40 °C

Silicon Rectifier Stacks

These spindle-mounted rectifier stacks (as shown in the photograph) incorporate the RS3 and RS5 series of silicon junction power



rectifier and are suitable for natural convection or forced air cooling.

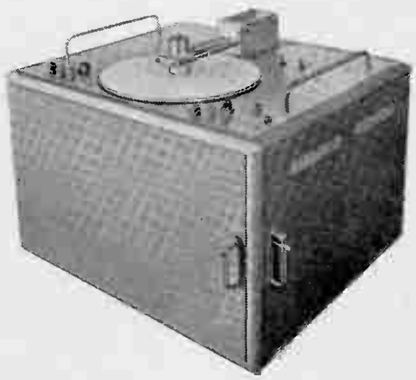
At present, the RS3 series is available with p.i.v. ratings ranging from 50 V to 800 V, while the RS5 series (described above) is available with p.i.v. ratings up to 400 V.

The makers claim that the design of these rectifier stacks offers many advantages such as low weight, small size, higher ambient operating temperatures (up to 100 °C), and storage temperatures ranging from −40 °C to +150 °C.

Standard Telephones & Cables Ltd.,
Rectifier Division, Edinburgh Way, Harlow, Essex.

Automatic Nyquist Diagram Plotter

An instrument designed for the accurate testing of d.c., a.c., hydraulic and pneumatic systems has been produced by Servo Consultants Ltd. The makers' claim that a Nyquist frequency plot of any servomechanism or network can be obtained within 5 minutes, this time being reduced to 90 seconds, if the response below 1 c/s is not required.



Details from the manufacturers' specification include:

Frequency range: 0·25 – 100 c/s.
Frequency markers: 0·25, 0·5, 1, 2, 4, 8, 16, 32, 64 c/s, with 3 smaller markers within each octave.
Graph paper size: 30 cm diameter.
Amplitude accuracy: 1%.
Phase-angle measurement accuracy: 1°.
Internal carrier frequency: 2 kc/s.
Carrier frequency when operated from external source: 50 – 3,000 c/s.
Output carrier modulation: 100%.

Servo Consultants Ltd.,
17 Woodfield Road, London, W.9.

Eddy-Current Non-Destructive Tester

This instrument, called the Elcotector, has been designed to carry out non-destructive measurements on all types of component coatings. It will also measure the thickness of most metallic and non-metallic foils, both in the static and moving state.

An eddy-current inspection device, it employs a special oscillator cast in a synthetic resin block which renders it impervious to changes of humidity. Readings are indicated on a built-in valve voltmeter.

Two types of probe are available, a pencil type for general-purpose testing and an extended type for use in narrow confined spaces.

East Lancashire Chemical Co. Ltd.,
Fairfield, Manchester.

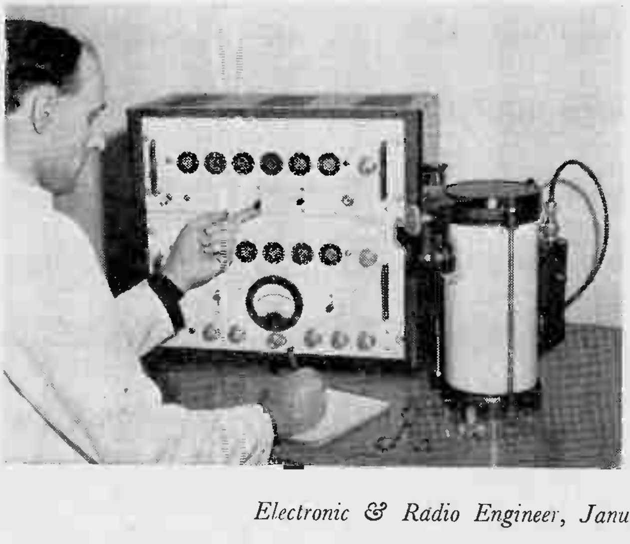
Scintillation Counter

The SC/LP scintillation counter has been developed for the counting of millimicrocurie quantities of very low-energy beta-radiation emitters, such as tritium and carbon-14. It is intended primarily for use with liquid phosphors but can be converted to accommodate both plastic phosphors and sodium iodide crystals. The equipment comprises a $1\frac{1}{2}$ in. thick (minimum) lead castle fitted with a light-tight door. This door is interlocked with an e.h.t. switch and a shutter assembly to protect the photomultiplier and to prevent external light increasing the background count.

A standard 30 c.c. glass jar is normally used as the container for the liquid phosphor into which is placed the radioactive source to be counted. On closing the castle door, the shutter opens automatically and operates a microswitch. This allows samples to be changed in normal room lighting while maintaining a low level of background counts. When the glass container is positioned in the castle, its bottom surface is adjacent to the sensitive surface of the photomultiplier tube and a good optical contact is maintained by means of a liquid light guide. The shutter assembly can be replaced by a crystal mount which will accept sodium iodide crystals of $1\frac{1}{2}$ in. diameter × 1 in., or plastic phosphors of similar dimensions. Inverting the unit enables liquid and solid samples to be investigated with a minimum air gap between the sample and detector.

For counting purposes, the instrument is used in conjunction with a linear amplifier and suitable scaling apparatus incorporating discriminator and e.h.t. supplies. Efficiencies of 10% for tritium and 50% for carbon-14 have been obtained at room temperature, but higher counting efficiencies can be achieved if the castle is cooled in a standard refrigerator.

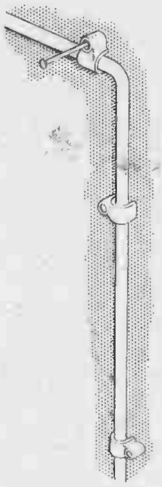
Panax Equipment Ltd.,
Holmthorpe Industrial Estate, Redhill, Surrey.



Plastic Cable Clips

A new type of wire and cable clip, made of unbreakable polystyrene with a pre-fixed hardened steel nail, has been introduced by Perihel Ltd.

With this new wiring accessory, the installation of coaxial, telephone and other round cables can be effected quickly at a relatively low cost, while providing a neater and cleaner finish.



The clips are available in seven sizes for cable diameters of 3 mm up to 9 mm, and in three colours of ivory, brown and grey.

Perihel Ltd.,
146 *New Cavendish Street, London, W.*1.

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

534.2-8-14 1

Measurement of Ultrasonic Velocity in Fluids under Pressure by the Pulse Method.—A. V. J. Martin. (*J. Rech. Centre nat. Rech. sci.*, Dec. 1957, No. 41, pp. 251–272.)

534.23+621.396.677.3 2

The Signal/Noise Performance of Electroacoustic Strip Arrays.—D. G. Tucker. (*Acustica*, 1958, Vol. 8, No. 1, pp. 53–62.) The directivity curves and signal/noise performance of 14 different arrangements of a strip array are discussed and tabulated. See also 994 of 1958.

534.641 : 621.395.92 3

An Artificial Ear for Insert Earphones.—J. Y. Morton. (*Acustica*, 1958, Vol. 8, No. 1, pp. 33–36.) "The artificial ear is designed to have the mean acoustical impedance of 19 normal ears. It includes an ear-mould simulator, physical representation of the ear-canal, and acoustical elements to represent the ear-drum impedance. One of these elements is a series acoustical resistance." See also 1001 of 1958 (Morton & Jones).

534.78 4

The 'Characterization' of Speech and Diction.—A. Moles. (*Ann. Télécommun.*, Jan. 1957, Vol. 12, No. 1, pp. 21–32.) Problems of voice identification and typification are examined theoretically and discussed on a statistical basis with particular reference to the use of visual representation of speech.

534.78 : 534.844 5

The Subjective Masking of Short-Time-Delayed Echoes by their Primary Sounds and their Contribution to the Intelligibility of Speech.—J. P. A. Lochner & J. F. Burger. (*Acustica*, 1958, Vol. 8, No. 1, pp. 1–10.) Experiments are described on the masking of single echoes under non-reverberant conditions using speech and pulsed tones as signals. Results of articulation tests are given which were carried out to determine the integration characteristics of the hearing mechanism for speech.

534.839 6

Techniques for Measuring and Evaluating Noise.—J. J. Hamrick. (*J. audio Engng Soc.*, Jan. 1958, Vol. 6, No. 1, pp. 19–25.) Measuring techniques and equipment for listening tests are described.

621.395.612.45.1 : 621.395.625 7

The Application of Velocity Microphones to Stereophonic Recording.—E. R. Madsen. (*J. audio Engng Soc.*, April 1957, Vol. 5, No. 2, pp. 79–85.) Single and two-channel reproduction systems are discussed. Suitable ribbon microphones are described.

621.395.623.743 8

The Electrostatic Loudspeaker.—D. T. N. Williamson. (*J. Instn elect. Engrs*, Aug. 1957, Vol. 3, No. 32, pp. 460–463.) A survey is made of progress in the field, concluding with an appraisal of the technique of constant-charge operation [see e.g. 2825 of 1955 (Walker)].

AERIALS AND TRANSMISSION LINES

621.315.212.029.62 9

Mutually Coupled *CR*-Type Directional Coupler.—S. Kurokawa, T. Takahashi & M. Arai. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 127–133.) Principles and construction are given of a directional coupler having a nearly constant coupling coefficient over the range 30–100 Mc/s without frequency compensation. Performance details are presented.

621.372.2 10

Investigations with a Model Surface-Wave Transmission Line.—G. Goubau & C. E. Sharp. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 222–227. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 895.)

621.372.2 : 621.318.134 11
: 621.385.029.6

Designing with Ferrite Isolators.—W. A. Hughes. (*Canad. Electronics Engng*, Feb. 1958, Vol. 2, No. 2, pp. 28–31.) An example is given of the method of selection of an isolator for a particular magnetron on the basis of maximum permissible frequency pulling. The magnetron Rieke diagram is used in conjunction with a nomogram for unilateral isolation.

621.372.2 : 621.396.11 12

Return Loss: Part 3.—T. Roddam. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 539–540.) Application of a transmission-line concept to problems of multipath propagation [see e.g. 2257 of 1958 (Bernath & Brand)]. Part 2: 346 of 1958.

621.372.8 13

Guided Wave Propagation in Submillimetric Region.—A. E. Karbowiak. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1706–1711.) An analysis of e.m. wave propagation in waveguides in the range 30–300 kMc/s and above. In the case of metallic waveguides at frequencies f well above cut-off the attenuation of TE_0 modes is always proportional to $f^{-3/2}$, but for TH waves the proportionality changes from $f^{1/2}$ to $f^{-5/2}$ as the frequency increases.

621.372.8 14

Step Discontinuities in Waveguides.—W. E. Williams. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 191–198. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 894.)

621.372.832.8 : 537.311.33 15

The Semiconductor Hall-Effect Circulator.—(*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 118–122.) A magnetic field is applied perpendicular to a circular slice of semiconductor to give circulator action between six equally spaced contacts at the edge. The effect of load changes on forward and reverse loss is discussed and construction methods and practical limitations are described.

621.396.67 16

Medium-Wave Aerials for Simultaneous Transmission of Broadcast Programmes.—L. Leng. (*Brown Boveri Rev.*, May 1958, Vol. 45, No. 5, pp. 218–223.) The simultaneous operation of two transmitters over a single aerial is considered with reference to existing installations.

621.396.674.3 17

The Exact Solution of the Field Intensities from a Linear Radiating Source.—R. N. Ghose. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 237–238.)

621.396.677.3 + 534.23 18

The Signal/Noise Performance of Electroacoustic Strip Arrays.—Tucker. (See 2.)

621.396.677.43 19

Studies on a Rhombic Antenna with Cylindrical Helices as the Arms.—A. K. Sen. (*Indian J. Phys.*, July 1958, Vol. 32, No. 7, pp. 303–316.) The use of helices is intended to reduce the area required for the erection of a rhombic aerial. Theoretical expressions are derived for the input impedance and directivity, and the results are compared with observed data using an aerial operating between 300 and 900 Mc/s.

621.396.677.5 20

Radiation Field of an Elliptic Loop Antenna with a Constant Current.—S. C. Loh & J. Y. Wong. (*Canad. J. Phys.*, June 1958, Vol. 36, No. 6, pp. 672–676.)

621.396.677.71 21

Radiation from a Fine Slot Traversed by a Travelling Wave.—J. Ernest. (*C. R. Acad. Sci., Paris*, 9th April 1958, Vol. 246, No. 14, pp. 2113–2116.) The conditions for radiation from a rectangular slot in a filled and unfilled waveguide are calculated.

621.396.677.71 22

Experimental Investigation of Radiation from a Fine Slot Traversed by a Travelling Wave.—J. Ernest. (*C. R. Acad. Sci., Paris*, 14th April 1958, Vol. 246, No. 15, pp. 2236–2239.) Radiation diagrams for a slot in a filled and unfilled waveguide are given. See 21 above.

621.396.677.833 23

A Simple Solution to the Problem of the Cylindrical Antenna.—J. G. Chaney. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 217–221. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 895.)

621.396.677.85 24

General Solution of the Luneberg Lens Problem.—S. P. Morgan. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1358–1368.) "The general solution is obtained for the index of refraction of a variable-index spherical lens which will form perfect geometrical images of the points of two given concentric spheres on each other."

AUTOMATIC COMPUTERS

681.142 25

Electronic Computers and the Engineer.—(*Electronic Radio Engr*, Nov. 1958, Vol. 35, No. 11, pp. 420–423.) An outline of the form and function of the main units of a digital computer. The basic techniques of programming are described.

681.142 26

The Physical Realization of an Electronic Digital Computer: Input and Output.—A. D. Booth. (*Electronic Engng*, Oct. 1958, Vol. 30, No. 368, pp. 570–574.) Continuation of a series of papers (447 of 1953 and back references). Electromechanical input-output equipment is described.

681.142 27

Universal Tape Amplifiers for Digital Data Systems.—R. F. Shaw. (*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 91–93.) Read- and write-amplifiers covering a wide range of input characteristics can be used with both return-to-zero and non-return-to-zero pulse techniques at rates up to 22 kc/s.

681.142 28

Designing Ultrasonic Delay Lines.—I. C. Miller & C. W. Sharek. (*Electronic Ind.*, July 1958, Vol. 17, No. 7, pp. 72–76 . . 118.) The significant operational characteristics of delay lines and their influence on design are examined.

681.142 29

The Minimality of Rectifier Nets with Multiple Outputs Incompletely Specified.—R. McNaughton & B. Mitchell. (*J. Franklin Inst.*, Dec. 1957, Vol. 264, No. 6, pp. 457–480.)

681.142 30

A Computer Oriented toward Spatial Problems.—S. H. Unger. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1744–1750.) A stored-program computer is described which can handle spatial problems by operating directly on information in planar form without scanning or other methods for transforming the problem into some other domain.

681.142 31

Analogue Computation.—E. L. Thomas. (*Brit. Commun. Electronics*, May 1958, Vol. 5, No. 5, pp. 348–358.) A survey of the principles, practice and current applications of computers, together with a tabulated analysis of British models.

681.142 32

New Job for an Old Method: Capacitor Storage used in Analogue Memory.—W. S. Kozak. (*Canad. Electronics Engng*, Feb. 1958, Vol. 2, No. 2, pp. 38–42.) A delay of 10 s for signals of frequency up to 10 c/s with negligible attenuation and an error of 1 % can be obtained by the unit described.

681.142 : 517.512.2 33

Fourier Analysis by a General Purpose Electronic Analogue Computer.—N. S. Nagaraja. (*J. Instn Telecommun. Engrs, India*, June 1958, Vol. 4, No. 3, pp. 130–136.) A method for use with a high-speed differential analyser is described in which the period of the harmonic function, generated by the computer, is kept constant, and the period of the arbitrary function made variable.

681.142 : 621.3.087.9 34

Electron Gun operates High-Speed Printer.—J. T. McNaney. (*Electronics*, 26th Sept. 1958, Vol. 31, No. 39, pp. 74–77.) A method of converting pulse-code data from a shaped-beam c.r. tube into printed records on untreated paper at a rate of 10^6 characters/min is described.

681.142 : 621.316.82 35

Resistance Potentiometers as Function Generators.—R. W. Williams & H. Marchant. (*Electronic Engng*, Oct. 1958, Vol. 30, No. 368, pp. 579–585.) "The generation of functions of a single variable by means of resistance potentiometers is reviewed." See also 37 of 1958 (Shen).

681.142 : 621.318.57 : 621.318.134 36

End-Fired Memory uses Ferrite Plates.—V. L. Newhouse, N. R. Kornfield & M. M. Kaufman. (*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 100–103.) Magnetic flux is induced around corresponding holes in two adjacent plates and in this way irregularities in the material from hole to hole do not affect the accuracy. Storing and switching systems using plates are described and suitable transistor drive and regeneration circuits are given.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21 37

Impedance—Matched or Optimum?—W. B. Snow. (*J. audio Engng Soc.*, April 1957, Vol. 5, No. 2, pp. 66–70.) Examples are given of situations, where matched impedance is the ideal condition, and of cases in which other impedance relations prove to be better.

621.3.049.75 38

Through-Connections for Printed Wiring.—J. W. Buckelew & E. D. Knob. (*Bell Lab. Rec.*, Oct. 1958, Vol. 36, No. 10, pp. 368–370.) In the process described a metal eyelet is passed through the printed-circuit board, staked and soldered in one operation.

621.316.549 39

A Voltage-Sensitive Switch.—K. O. Otley, R. F. Shoemaker & P. J. Franklin. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1723–1730.) The controlled dielectric breakdown of oxide films has been investigated. A small, shock-resistant device requiring no additional power supply has been developed which consists of an oxide layer between an Al base and a Ag film. Before breakdown the device has a resistance of at least $10^7\,\Omega$; on switching the resistance drops to about $1\,\Omega$. Some circuit applications are described.

621.316.825 40

Solving Thermistor Problems.—R. S. Goodyear. (*Electronic Ind.*, July 1958, Vol. 17, No. 7, pp. 51–55 . . 118.) The methods of designing circuits incorporating thermistors are illustrated by giving three particular cases.

621.316.86 41

Nonlinear Properties of Carbon Resistors.—C. E. Mulders. (*Tijdschr. ned. Radiogenoot.*, Nov. 1957, Vol. 22, No. 6, pp. 337–347. In English.) A method of determining the nonlinear characteristics is proposed which is based on the measurement of harmonics produced when a pure sinusoidal current flows through the resistor.

621.316.89 42

The Versatile Vamistor.—R. C. Langford. (*Canad. Electronics Engng*, Feb. 1958, Vol. 2, No. 2, pp. 19–23.) Construction, specification and performance data of a resistor incorporating the best features of wire-wound and carbon types.

621.318.57 : 621.314.7 43

Designing Transistor Circuits—Switching Statics.—R. B. Hurley (*Electronic Equipm. Engng*, June 1958, Vol. 6, No. 6, pp. 42–46.) The basic switching properties of junction transistors are examined, and examples of single and multiple switches are given.

621.318.57 : 621.396.669 44

An Electronic Transmit-Receive Switch.—A. S. McNicol. (*R.S.G.B. Bull.*, July 1958, Vol. 34, No. 1, pp. 22–23.) Incoming signals are fed via the tank circuit to a r.f. pentode which is cut-off by grid bias when the transmitter is on. See also 1347 of 1958.

621.319.4 : 621.3.082.6 45

A Temperature-Sensitive Ceramic Reactance Element.—C. V. Ganapathy, C. S. Rangan, R. Krishnan & T. V. Ramamurti. (*J. Instn Telecommun. Engrs, India*, June 1958, Vol. 4, No. 3, pp. 168–174.) A sintered mixture of $BaTiO_3$ and CeO gives, within the range −50°C to 100°C, a nearly linear, repeatable capacitance/temperature curve with a 50% change of capacitance. The effects of different percentages of CeO are examined and applications are mentioned.

621.319.43 : 621.314.63 46

A Voltage-Variable Capacitor.—G. F. Straube. (*Electronic Ind.*, May & July 1958, Vol. 17, Nos. 5 & 7, pp. 69–73 & 77–80.) Details of a subminiature Si junction device with a variable capacitance controllable by bias voltage are given. Applications of the 'varicap' to an f.m. transmitter, an amplifier and a variable filter are described. See also 1677 of 1958 (Keizer).

621.372.5 47

Response of a Capacitance-Resistance Divider to the Step Function, Exponential Function and Ramp Function.—S. Turk. (*Electronic Engng*, Oct. 1958, Vol. 30, No. 368, pp. 608–611.)

621.372.5 : 621.314.7 48

A Four-Pole Analysis for Transistors.—B. J. Alcock. (*Electronic Engng*, Oct. 1958, Vol. 30, No. 368, pp. 592–594.) "An algebraically convenient analysis for four-pole networks is presented, based upon determinant techniques; it is applied to transistor circuits and introduces two new comprehensive parameters for the transistor."

621.372.54 49

Additional Tables for Design of Optimum Ladder Networks.—L. Weinberg. (*J. Franklin Inst.*, July & Aug. 1957, Vol. 264, Nos. 1 & 2, pp. 7–23 & 127–138.) Tables for networks with characteristics given by Butterworth, Tchebycheff and Bessel polynomials are presented; they are classified on the basis of the conductance ratio. See also *Proc. nat. Electronics Conf., Chicago*, 1956, Vol. 12, pp. 794–817, and 702 of 1958.

621.372.54 50

Estimation of Dissipative Effects in Tchebycheff Symmetrical Filters.—D. C. Pawsey. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1763–1764.) Comment on 2053 of 1957 (Grossman).

621.372.542.2 51

Low-Pass *RC* Filter with Optimum Response.—D. D. Nye, Jr. (*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 104–105.) Formulae and graphs for one- and two-section filters are given.

621.372.543.2 : 538.652 52

Narrow-Band Magnetostrictive Filters.—A. P. Thiele. (*Electronic Radio Engr*, Nov. 1958, Vol. 35, No. 11, pp. 402–411.) 20 filters, each with a 3-dB bandwidth of 7 c/s, are arranged with inputs in series and outputs isolated from one another, to form a bank with a centre frequency of about 40 kc/s. Design and construction are described and applications of single resonators and arrays are considered. Temperature effects are reduced by using cobalt-substituted ferrites.

621.372.543.2 : 621.372.852.1 53

A Waveguide Filter Theory.—M. van Sliedregt. (*Tijdschr. ned. Radiogenoot.*, Nov. 1957, Vol. 22, No. 6, pp. 375–389. In English.) On the basis of the matrix theory used by Hessel et al. (3374 of 1949) formulae are derived giving insertion loss and envelope delay for multistage filters having as parameters the ratios of the bandwidth of the stages to the bandwidth of the total filter.

621.373 54

Wide-Range *RC* Oscillator.—C. G. Mayo & J. W. Head. (*Electronic Radio Engr*, Nov. 1958, Vol. 35, No. 11, pp. 412–416.) A frequency range exceeding 1 000 : 1 in a single sweep can be obtained with the tone source described. Elements having continuously distributed series resistance and shunt capacitance are used, which can be made from standard high-stability resistors.

621.373.431.1 55

Negatively-Biased Multivibrator.—A. Bar-Lev. (*Electronic Radio Engr*, Nov. 1958, Vol. 35, No. 11, pp. 436–440.) Description of a multivibrator which has its grids returned to a negative bias. It has high frequency stability and is suitable for use as a variable frequency source or a source of special wave-shapes, and can be frequency modulated. The frequency varies linearly with bias voltage.

621.373.431.2 : 621.314.7 56

Transistor Blocking-Oscillator Circuits.—A. S. Daddario. (*Electronic Equipm. Engng*, June 1958, Vol. 6, No. 6, pp. 55–58.) Circuit characteristics are examined and methods for varying the output waveform are discussed.

621.374.3 : 621.318.57 57

Increased Sensitivity of Trigger Circuits.—M. M. Vojinović. (*Bull. Inst. nuclear Sci. 'Boris Kidrich'*, March 1958, Vol. 8, pp. 123–131.) In the cathode-coupled monostable multivibrator described, the use of a crystal diode in the feedback loop allows one valve to amplify the trigger pulse before it is applied to the other valve. The circuit is sensitive to a few millivolts.

621.374.32 58

Latching Counters.—W. P. Anderson & N. A. Godel. (*Electronic Radio Engr*, Oct. & Nov. 1958, Vol. 35, Nos. 10 & 11, pp. 362–367 & 425–436.) The causes of low reliability of single-pulse counter chains are examined. A circuit with a multiphase output is described, which has a frequency range from d.c. to 500 kc/s and eliminates spurious pulses. Reset facilities and general design features including the use of transistor circuits are discussed.

621.374.32 **59**

Sensitive Single Channel Pulse-Height Analyser.—M. Simhi & M. Birk. (*Rev. sci. Instrum.*, Sept. 1958, Vol. 29, No. 9, pp. 766–768.) The analyser operates with input pulses of amplitude 4–40 mV and a dead time of about 6 μs. The stability over a period of several days is better than 1%.

621.374.4. : 621.314.6 **60**

Harmonic Generation with Ideal Rectifiers.—C. H. Page. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1738–1740.) The nth harmonic cannot be generated with an efficiency exceeding n^{-2}. At least 75 % of the power converted to d.c. and harmonics is d.c. dissipation; this cannot be reduced by an arrangement of selective circuits.

621.374.4 : 621.372.44 **61**

Harmonic Generation with Nonlinear Reactances.—K. K. N. Chang. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 455–464.) A simple theory of frequency multiplication using nonlinear reactances is derived and compared with the results obtained with the nonlinear capacitance of a Ge point-contact diode. The theory and the experimental results are in agreement at low input powers. Possible reasons for the discrepancies at high powers are mentioned.

621.375.2.018.756 **62**

Linear Amplifier for Negative Pulses.—A. S. Penfold. (*Rev. sci. Instrum.*, Sept. 1958, Vol. 29, No. 9, pp. 765–766.) A two-valve amplifier is described which has a gain of 100 and a rise time of 50 mμs for outputs up to 125 V.

621.375.2.029.3 **63**

Three-Valve Preamplifier.—C. Hardcastle. (*Mullard tech. Commun.*, Aug. 1958, Vol. 4, No. 32, pp. 39–42.) A description and performance specification of a circuit with inputs and crystal pickups, tape recorder and radio receiver. Simultaneous outputs for tape recording and for a power amplifier are provided. See also 2023 of 1958.

621.375.221.029.6 **64**

Broad-Band Amplifier for Radar and Scatter.—J. H. Phillips & E. Maxwell. (*Electronics*, 26th Sept. 1958, Vol. 31, No. 39, pp. 81–83.) The design of a low-noise two-stage amplifier and mixer for the range 400–450 Mc/s is described. Overall power gain is 29 dB with noise figures varying from 3·6 to 5·5 dB over the frequency band.

621.375.4 **65**

Design Considerations for Direct-Coupled Transistor Amplifiers.—J. E. Lindsay & H. J. Woll. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 433–454.) It is shown that: (*a*) the maximum obtainable signal/drift ratio of an amplifier is achieved by choosing the optimum source resistance, (*b*) the signal/drift capability of a transistor and the optimum source resistance are nearly the same for the three configurations: common emitter, common collector and common base, (*c*) negative feedback can change the optimum source resistance and generally degrades the signal/drift ratio, (*d*) improved signal/drift ratio can be obtained at reduced emitter currents.

621.375.4 **66**

Circuit Designed for High Voltage Gain from Transistors.—M. Price. (*Canad. Electronics Engng*, March 1958, Vol. 2, No. 3, pp. 20–21.) In the circuit described, a common-emitter stage feeds into a modified common-collector stage with low output impedance.

621.375.4.029.3 **67**

Transistor Audio Amplifier.—F. Butler. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 529–535.) Design principles are outlined for using multiple push-pull stages in series. The construction of a 10-W amplifier for battery operation and of a low-noise preamplifier is described.

621.375.4.029.3 **68**

15-W Public Address Amplifiers using OC16 Transistors.—P. Tharma. (*Mullard tech. Commun.*, Aug. 1958, Vol. 4, No. 32, pp. 30–32.) The amplifiers work from 14-W or 28-W supplies and may be fully driven by a low-impedance microphone.

621.375.4.029.33 **69**

A Transistor Video Amplifier having 80 Volts Output.—V. H. Grinich. (*Trans. Inst. Radio Engrs*, Sept. 1956, No. PGBTS-5, pp. 32–37. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1898.)

621.375.4.029.5 **70**

Transistorized I.F.-Strip Design.—R. E. Murphy & R. S. Mautner. (*Electronic Equipm. Engng*, June 1958, Vol. 6, No. 6, pp. 51–53.) The performance of tetrodes in grounded-emitter and grounded-base circuits is compared. A nine-stage 30-Mc/s i.f. amplifier using grounded-emitter tetrodes is described, with transformer coupling between stages providing for neutralizing and matching.

621.375.4.078 **71**

Temperature-Stable Transistor Circuit based on the Half-Supply-Voltage Principle.—B. G. Dammers, A. G. W. Uitjens & W. Ebbinge. (*Electronic Applic. Bull.*, Jan. 1958, Vol. 18, No. 1, pp. 1–11.) The principle is applied to the design of a transformerless a.f. output stage and is used in each stage of a six-transistor superheterodyne receiver. See 628 of 1958 (Johnson & Vermes).

621.375.9 : 538.569.4.029.6 **72**

Masers and Related Quantum-Mechanical Devices: Parts 1 & 2.—G. E. Weibel. (*Sylvania Technologist*, Oct. 1957, Vol. 10, No. 4, pp. 90–97 & Jan. 1958, Vol. 11, No. 1, pp. 26–43.) Theoretical aspects are considered leading to the computation of the dielectric constant of a gaseous medium in a form which can be directly applied to the solution of problems in waveguides and cavities.

621.375.9 : 538.569.4.029.6 **73**

Behaviour of a Two-Level Solid-State Maser.—I. R. Senitzky. (*Phys. Rev. Lett.*, 1st Sept. 1958, Vol. 1, No. 5, pp. 167–168.) It is shown, under simplifying assumptions, that when a system of spins is in resonance with the cavity, and when damping effects caused by relaxation phenomena and cavity losses are not excessive, modulation of the cavity power output is exactly what is to be expected on the basis of a dynamical analysis. See also 2677 of 1958 (Chester et al.).

621.375.9 : 538.569.4.029.6 **74**

Proposed Method for Tuning a Maser Cavity.—F. O. Vonbun. (*Rev. sci. Instrum.*, Sept. 1958, Vol. 29, No. 9, pp. 792–793.) A method is described for determining the small difference between the cavity resonance frequency and the frequency of the microwave spectral line.

621.375.9 : 621.3.011.23 **75**

Circuit Conditions for Parametric Amplification.—L. B. Valdes. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 129–141.) It is shown that signals may be amplified in a circuit which consists exclusively of passive elements, if a nonlinear or time-varying reactance is present. The circuit conditions which must be satisfied are established and the differences between parametric amplifiers and mixers or modulators are discussed.

621.375.9 : 621.3.011.23 **76**

Parametric Amplification and Frequency Mixing in Propagating Circuits.—P. K. Tien. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1347–1357.) An analysis is given of the properties of time-varying distributed reactance in propagating structures which may have positive or negative phase and group velocities. Power at different frequencies may be converted from one frequency to the others, and the amplitudes of the waves may vary exponentially or periodically, depending on the relation of the frequencies and of the phase constants between the propagating waves and the variable coupling reactance. Applications to wide-band frequency converters, frequency channel selectors, wide-band amplifiers, tunable narrow-band amplifiers, and oscillators are described.

621.375.9 : 621.3.011.23 **77**

Gain, Bandwidth, and Noise Characteristics of the Variable-Parameter Amplifier.—H. Heffner & G. Wade. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1321–1331.) Increased gain is achieved by raising the Q of the idling resonant circuit or the amount of variation in the variable coupling reactance. The bandwidth is inversely proportional to Q and to the voltage gain, and directly proportional to the ratio of idling resonance frequency to amplifying frequency. Very low noise figures should be attained by artificially cooling the idling resonant circuit. The use of the parametric principle as applied to frequency conversion is discussed.

621.375.9 : 621.385.029.63 : 537.533 **78**

A Low-Noise Electron-Beam Parametric Amplifier.—Adler & Hrbek. (See 321.)

621.375.9.029.6 : 621.3.011.23 **79**

The Mavar: A Low-Noise Microwave Amplifier.—S. Weber. (*Electronics*, 26th

Sept. 1958, Vol. 31, No. 39, pp. 65–71.) The operation of the Mavar (mixer amplification by variable reactance) requires a lossless nonlinear reactance exhibiting negative resistance under certain conditions. Low-noise amplification is achieved at room temperature. The performance and the operation of the main types of parametric amplifier are outlined.

GENERAL PHYSICS

534.1 + 538.56 80

Strongly Nonlinear Oscillations.—A. Papoulis. (*J. Math. Phys.*, July 1958, Vol. 37, No. 2, pp. 147–156.) Mathematical analysis of a special class of nonlinear oscillations in systems without or with losses.

536.7 : 530.17 : 621.372 81

A Note on Thermodynamics of the Harmonic Oscillator with Radiation Damping.—N. Saitô & J. Hori. (*J. phys. Soc. Japan*, July 1958, Vol. 13, No. 7, pp. 717–721.) It is shown that the impedance of an irreversible thermodynamical system must be positive and real; the argument is based on representation of the system by Langevin equations of the second order, corresponding to the construction of any passive network from elementary *RLC* circuits. The conclusion is illustrated by an examination of the properties of the harmonic oscillator.

537.525 82

Theory of the Cathode Sheath in a Low-Density Gas Discharge.—P. L. Auer, H. Hurwitz, Jr, & S. Tamor. (*Phys. Rev.*, 15th Aug. 1958, Vol. 111, No. 4, pp. 1017–1028.) The space-charge equations of a low-density gas discharge are derived and applied to the study of potential distributions in the cathode sheath. One conclusion is that a positive column will begin to evolve when the number of ionization mean free paths in the discharge gap is of the same order as the square root of the ratio of electron to positive-ion mass.

537.525 : 538.69 83

The Effect of a Uniform Magnetic Field on Electrodeless Discharge in a Tube and Measurement of Electronic Mobility: Part 1—Air.—S. N. Goswami. (*Indian J. Phys.*, Jan. 1958, Vol. 32, No. 1, pp. 35–41.)

537.533 84

Perveance and the Bennett Pinch Relation in Partially Neutralized Electron Beams.—J. D. Lawson. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 146–151.) The concept of perveance is extended to take account of relativistic conditions and the effects of space-charge neutralization. See also 1919 of 1958.

537.533 : 538.63 85

Magnetic Forces and Relativistic Speeds in Stationary Electron Beams.—J. M. Winwood. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 161–162.) Comment on 3037 of 1958 (Meltzer).

537.56 86

A Variational Calculation of the Equilibrium Properties of a Classical Plasma.—S. F. Edwards. (*Phil. Mag.*, Feb. 1958, Vol. 3, No. 26, pp. 119–124.)

537.56 87

Correlations in the Charge Density of a Classical Plasma.—S. F. Edwards. (*Phil. Mag.*, March 1958, Vol. 3, No. 27, pp. 302–306.)

537.56 88

Nonlinear Theory of Plasma Oscillations and Waves.—S. Amer. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 105–113.) An exact differential equation for density oscillations is found considering only Coulomb interactions between electrons. A number of properties of plasma oscillations are then derived including the form of stationary spherical waves in an isotropic plasma. See also 2707 of 1958 (Gold).

537.56 89

Thermal Diffusion in Ionized Gases.—S. Chapman. (*Proc. phys. Soc.*, 1st Sept. 1958, Vol. 72, No. 465, pp. 353–362.) The 'thermal diffusion factor' α, which is less than unity in a neutral gas mixture, is of order $Z+1$ in an ionized gas, where Ze is the ionic charge. In a mixed ionized gas containing also a small proportion of heavy ions the factor can be much greater and give increased heavy ion density in the hotter regions. This effect may be of importance in the solar corona.

538.122 90

Experimental Determination of the Field of a Permeable Alloy Cylinder, Placed in a Uniform Magnetic Field Parallel to its Axis of Revolution.—E. Selzer. (*Ann. Géophys.*, April–June 1956, Vol. 12, No. 2, pp. 144–146.)

538.22 91

Effective Parameters in Ferrimagnetic Resonance.—R. K. Wangsness. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 813–816.) The steady-state solution for the susceptibility tensor of a two-sublattice system has been found by using sublattice equations of motion which include complete Landau-Lifschitz relaxation terms with individual relaxation parameters and which describe relaxation toward the instantaneous total field acting on the sublattice.

538.3 : 52 92

Spherical Vortices in Magnetohydrodynamics.—C. Agostinelli. (*R.C. Accad. naz. Lincei*, Jan. 1958, Vol. 24, No. 1, pp. 35–42.)

538.566 : 535.43 93

Propagation of Waves through a Sheet of the Medium with a Cubic Periodic Structure.—P. Gosar. (*Nuovo Cim.*, 16th March 1958, Vol. 7, No. 6, pp. 742–763. In English.) An approximate solution to the wave equation is obtained for the case of normal and oblique incidence of the ingoing wave and a detailed investigation is made of the case for which the wave equation is separable.

538.566 : 621.396.677.85 94

Some Electromagnetic Transmission and Reflection Properties of a Strip Grating.—R. I. Primich. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 176–182. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 894.) See also 1119 of 1958.

538.569.4 95

A General Theory of Magnetic Resonance Saturation.—K. Tomita. (*Progr. theor. Phys.*, May 1958, Vol. 19, No. 5, pp. 541–580.) A general theory for describing a spin assembly due to a nearly resonant rotating magnetic field of arbitrary strength is given. Two different cases are treated dependent on the relation between the static field and the frequency of fluctuation of the environment.

538.569.4.029.6 96

A Microwave Electron-Spin-Resonance Spectrometer.—M. B. Palma-Vittorelli & M. U. Palma. (*Nuovo Cim.*, 1958, Vol. 7, Supplement No. 1, pp. 139–154.) A general description is given of equipment which operates in the X and K bands and has highly stable magnetic and microwave frequency systems.

538.569.4.029.6 : 621.375.9 97

Masers and Related Quantum-Mechanical Devices: Parts 1 & 2.—Weibel. (See 72.)

538.569.4.029.64 98

Paramagnetic Resonance in Copper Propionate Monohydrate.—H. Abe. (*J. phys. Soc. Japan*, Sept. 1958, Vol. 13, No. 9, pp. 987–997.) Results of an investigation at cm λ of absorption in single crystals.

538.691 99

The Influence of the Anomalous Magnetic Moment on the Spin Kinematics of Electrons in a Uniform Magnetic Field.—M. Carrassi. (*Nuovo Cim.*, 16th Feb. 1958, Vol. 7, No. 4, pp. 524–535. In English.) High-energy particles in a beam are considered.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.03 : 061.3 100

Proceedings of the Third Symposium on Cosmical Gas Dynamics held at the Smithsonian Astrophysical Observatory, Cambridge, Massachusetts, June 24th–29th, 1957.—(*Rev. mod. Phys.*, July 1958, Vol. 30, No. 3, pp. 905–1108.) 41 papers presented deal with velocity fields in the interstellar medium, cooling and condensation of interstellar matter, conditions at the ionization and shock fronts in collisions of gas clouds, and related problems.

523.164.32 **101**

Polarization Measurements of the Three Spectral Types of Solar Radio Burst.—M. Komesaroff. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, pp. 201–214.) A frequency-sweep technique was used in the range 40–240 Mc/s, most data being obtained between 40 and 140 Mc/s. Data for type-I and type-III bursts cover the period January–October 1955, and for type-II the years 1955 and 1956. Many type-III bursts were highly polarized, and there were strong indications that the polarization was due to the radiation source.

523.164.32 **102**

Evidence of Echoes in the Solar Corona from a New Type of Radio Burst.—J. A. Roberts. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, pp. 215–234.) The bursts contain two elements spaced by $1\frac{1}{2}$–2 sec, the second being a repetition of the first with a frequency increase with time of 2–8 Mc/s per sec. An explanation is derived in terms of echoes from lower levels of the solar corona.

523.164.4 **103**

Evidence on the Spatial Distribution of Radio Sources Derived from a Survey at a Frequency of 159 Mc/s.—D. O. Edge, P. A. G. Scheuer & J. R. Shakeshaft. (*Mon. Not. R. astr. Soc.*, 1958, Vol. 118, No. 2, pp. 183–196.) Some results from a survey of radio sources at 159 Mc/s are available for limited regions of the sky. These are discussed in relation to the spatial distribution of sources.

523.165 : 523.75 **104**

Short γ-Ray Burst from a Solar Flare.—L. Peterson & J. R. Winckler. (*Phys. Rev. Lett.*, 15th Sept. 1958, Vol. 1, No. 6, pp. 205–206.) Observations of a burst of γ radiation are compared with simultaneous records of solar radio emission at 3 and 27 cm λ.

523.5 **105**

A Suggested Improvement to the C.W. Technique for Measurement of Meteor Velocities.—J. S. Mainstone, W. G. Elford & A. A. Weiss. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, pp. 277–278.)

523.5 : 538.082.74 **106**

A Search for Magnetic Effects from Meteors.—G. S. Hawkins. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 467–475.) No magnetic pulses could be detected as a result of meteors having visual magnitudes between $+5$ and -3. The sensitivity of the magnetometer used allowed pulses of 5×10^{-2} and later 3×10^{-3} γ to be recorded. See also 3797 of 1958.

523.746 **107**

Behaviour of the Present Sunspot Cycle.—S. W. Visser. (*Nature, Lond.*, 26th July 1958, Vol. 182, No. 4630, pp. 253–254.)

550.37/.38 **108**

Geomagnetic Dynamos.—A. Herzenberg. (*Phil. Trans. A*, 21st Aug. 1958, Vol. 250, No. 986, pp. 543–583.) A rigorous proof is given of the dynamo theory of geomagnetism by postulating a velocity pattern in a sphere filled with conducting fluid so that the arrangement acts as a dynamo producing an external magnetic field.

550.38 **109**

A Method for Analysing Values of the Scalar Magnetic Intensity.—A. J. Zmuda. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 477–490.) The method uses a series for the square of the scalar intensity, the terms of which are obtained from the spherical harmonics generally applied to each component of the intensity. Those magnetic characteristics normally given by analysis of the vector intensity are obtained.

550.38 : 551.510.536 **110**

Large-Amplitude Hydromagnetic Waves above the Ionosphere.—A. J. Dessler. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 507–511.) Arguments are presented for the existence of waves of amplitude about 100 times that of geomagnetic fluctuations at the earth's surface; it is assumed that the latter are generated at six to ten earth radii and carried to earth by the waves. It is suggested that the waves produce irregularities in electron density which cause radio-star scintillation and that they are also responsible for high-intensity particle radiation above about 1 000 km.

550.38 (98+99) **111**

Scientific Reports of the French Polar Expeditions S IV 2: Magnetic Activity in the Polar Regions.—P. N. Mayaud. (*Ann. Géophys.*, Jan.–March 1956, Vol. 12, No. 1, pp. 84–101.)

550.380.8 : 538.569.4 **112**

Measurement of the Earth's Magnetic Field with a Rubidium Vapour Magnetometer.—T. L. Skillman & P. L. Bender. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 513–515.) The instrument described makes use of the relationship between the earth's field and the absorption lines in Rb vapour caused by ground-state Zeeman transitions. An absolute accuracy of 2 γ is quoted. See also 760 of 1958 (Bell & Bloom).

550.385 : 523.75 **113**

Geomagnetic Activity and Eruptions.—P. Simon. (*Ann. Géophys.*, July–Sept. 1956, Vol. 12, No. 3, pp. 167–182.) Statistical analysis of observational data to determine the relation between various types of solar flare and increases in geomagnetic activity.

550.389.2 **114**

I.G.Y. Progress Report.—G. M. C. Stone. (*R.S.G.B. Bull.*, July 1958, Vol. 34, No. 1, pp. 15–17.) A general report of contributions made by members of the R.S.G.B. to the I.G.Y. program.

550.389.2 **115**

I.G.Y. V.H.F. Programme—Progress to Date.—C. E. Newton & G. M. C. Stone. (*R.S.G.B. Bull.*, July 1958, Vol. 34, No. 1, pp. 13–15.) The report, compiled from observations made by members of the R.S.G.B., deals with (*a*) an analysis of the origin of different air masses in relation to tropospheric propagation conditions, (*b*) fading, (*c*) auroral reflection propagation, and (*d*) solar noise.

550.389.2 : 621.396.11 **116**

Hit Rates of Radio Propagation Disturbance Warnings and S.W.I. Warnings.—K. Sinno. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 109–116.) In order to improve the hit rate, defined as the number of hits divided by the number of issued forecasts, it is suggested (*a*), that long-distance propagation characteristics should be studied to account for the fall in hit rate in winter, (*b*), that forecasting should be reduced to three or four days ahead, and (*c*), that although S.W.I. warnings are fairly good, there should be a further study of solar radio noise.

550.389.2 : 629.19 **117**

Gravitational Torque on a Satellite Vehicle.—R. E. Roberson. (*J. Franklin Inst.*, Jan. 1958, Vol. 265, No. 1, pp. 13–22.) See also 3803 of 1958.

550.389.2 : 629.19 **118**

Motion of the Nodal Line of the Second Russian Earth Satellite (1957β) and Flattening of the Earth.—E. Buchar. (*Nature, Lond.*, 19th July 1958, Vol. 182, No. 4629, pp. 198–199.)

550.389.2 : 629.19 **119**

Air Drag Effect on a Satellite Orbit described by Difference Equations in the Revolution Number.—R. E. Roberson. (*Quart. appl. Math.*, July 1958, Vol. 16, No. 2, pp. 131–136.) Difference equations are derived whose solutions express changes in orbital size and shape, and give the satellite's time behaviour as functions of the revolution number. The results are obtained from a first-order perturbation theory using a small air-drag parameter and assuming the air density function to be locally exponential. See also 3803 of 1958.

550.389.2 : 629.19 **120**

On the Reception of Radio Waves from Russian Earth Satellite I.—H. Uyeda, T. Ishida, H. Shibata & M. Mambo. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 135–141.) Observations of bearing and field strength are used to determine the orbit and to estimate regions of reception, taking into account the various modes of ionospheric propagation.

550.389.2 : 629.19 **121**

Radio Reflections from Satellite-Produced Ion Columns.—C. D. Hendricks, Jr, G. W. Swenson, Jr, & R. A. Schorn. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, p. 1763.) The records of signal strength of WWV transmission on 20 Mc/s measured at the University of Illinois were examined for periodic variations due to reflection from ion columns. The results do not indicate the presence of such columns for extremely high satellites. See also 1725 of 1958 (Kraus).

550.389.2 : 629.19 **122**

Satellite Tracking: Its Methods and Purpose.—E. G. C. Burt. (*Endeavour*, Oct. 1958, Vol. 17, No. 68, pp. 216–222.) Radio, radar and optical techniques are discussed.

550.389.2 : 629.19 **123**

Keeping Track of Earth Satellites.—C. J. Sletten, G. R. Forbes, Jr, & L. F. Shodin. (*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 81–83.) Details are given of an interferometer-type aerial array having 22 dipoles proximity-coupled to a two-wire transmission line. A receiver with inputs switched at 100 c/s is used to detect signals below receiver noise level. Calibration methods are described and some transit passage records, accurate to within 1 sec, are shown.

550.389.2 : 629.19 **124**

Telemetering Information from Satellites.—N. G. Hyde. (*R.S.G.B. Bull.*, July 1958, Vol. 34, No. 1, pp. 8–12.) The data given cover both U.S.S.R. and U.S. satellites. Channel assignments for the Lyman-α environmental satellites are tabulated and general notes on recording technique are included.

551.510.535 **125**

Pressure and Temperature Variation of the Electron-Ion Recombination Coefficient in Nitrogen.—E. P. Bialecke & A. A. Dougal. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 539–546.) The variations have been investigated in the pressure range 0·2—2 mm Hg and for electron temperatures between 92° and 300°K. At 1·3 mm Hg, α_{ei} varies from $8 \cdot 5 \times 10^{-7}$ cm^3 sec^{-1} at 300°K to $6 \cdot 7 \times 10^{-6}$ cm^{3}sec^{-1} at 92°K. Dissociative recombination is the most probable mechanism.

551.510.535 **126**

Computations of Electron Density Distributions in the Ionosphere making Full Allowance for the Geomagnetic Field.—R. A. Duncan. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 491–499.) Jackson's method (2381 of 1956) for computing the distributions from $h'(f)$ records has been slightly modified for use on an electronic computer. Each reduction takes about 20 secs. It is assumed that the electron density between two layers falls to 90 % of the lower layer value. Night time distributions for Brisbane, Australia, are found to be nearer to the Chapman than to the parabolic form.

551.510.535 **127**

Long Term Variations of the Sporadic E Layer in Japan.—I. Kasuya. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 117–125.) Statistical analysis of E_s-layer reflections at four observatories during 1947–1955 shows marked diurnal and seasonal variations in fE_s. No annual variation is observed nor any correlation with sunspot number. An analysis of the occurrence of reflections above certain limiting frequencies at one observatory yields similar results.

551.510.535 **128**

On the Influence of Electron-Ion Diffusion Exerted upon the Formation of the F_2 Layer.—T. Yonezawa. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 165–187.) The maximum electron density of the F_2 layer, and its height, $h_m F_2$, are obtained theoretically as functions of latitude and solar zenith distance. The theoretical height distributions derived are thinner than Chapman distributions and the effective recombination coefficient decreases with height at low altitudes but becomes constant at high altitudes. No better agreement between theory and observation is obtained when a vertical electron-ion drift velocity is considered. See also 3473 of 1957.

551.510.535 **129**

On the World Wide Distribution of f_0F_2.—H. Shibata. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 235–256.) The coefficients *a* and *b* in the relation $f_0F_2 = aR + b$, where $R =$ sunspot number, are separated into components which are constant and those which vary diurnally and annually (see 2410 of 1958). Examination of these components for 47 stations shows a geomagnetic latitude control of the constant component while there is no significant difference between control by geomagnetic or geographic latitude for the varying components.

551.510.535 : 523.7 : 621.396.11 **130**

Solar-Cycle Influence on the Lower Ionosphere and on V.H.F. Forward Scatter.—C. Ellyett & H. Leighton. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1711–1716.) Seven years' results at 49·8 Mc/s for the Cedar Rapids–Sterling path (1 243 km) have been analysed to determine the effect of the solar cycle. The monthly median signal intensity for the noon and afternoon period is found to follow the same trend as the sunspot number and the magnetic disturbance index.

551.510.535 : 523.78 **131**

Ionospheric Observations made at Freiburg during the Solar Eclipse of 30th June 1954.—R. Busch, E. Harnischmacher & K. Rawer. (*Ann. Géophys.*, Jan.–March 1956, Vol. 12, No. 1, pp. 1–15.)

551.510.535 : 525.624 **132**

Lunar Tides in E_{2s} at Brisbane.—A. D. Gazzard. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, p. 272.) See also 1717 of 1956 (Thomas & Svenson).

551.510.535 : 550.38 **133**

The Magnetic Field in the F_2 Layer at Dakar.—K. Suchy & P. Vila. (*Ann. Géophys.*, Oct.–Dec. 1956, Vol. 12, No. 4, pp. 277–282.) The gyrofrequency is calculated from observed values of f_0 and f_x, and its variations as a function of reflection height and time are analysed.

551.510.535 : 550.385 **134**

Anomalous Changes in the Ionosphere Related to a Severe Magnetic Storm, Oct. 28, 1951.—T. Obayashi. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 213–225.) Examination of world-wide geomagnetic and ionospheric data shows the existence of a short-lived electrojet stream with associated anomalous changes in F_2-layer electron density. Two possible mechanisms are suggested: (*a*) the formation of a new F_2 layer due to corpuscular streams, and (*b*) the vertical drift of electrons resulting from interaction of the geomagnetic field with currents in the F_2 layer returning from the main electrojet in the E region. See also 3139 of 1957 (Sato).

551.510.535 : 621.396.11 **135**

F_2 Layer Deduced from the 'Frequency Spectrum' of Radio-Propagation *Q* Figures at Long-Distance Routes.—T. Obayashi. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 227–234.) The hourly radio-propagation quality figures for WWV, received at many stations in Canada, were used as basic data. Assuming a suitable height for the F_2 layer, its critical frequency at the reflection point of a transmission path could be obtained. The values calculated compared fairly well with those obtained from vertical incidence soundings provided the correct mode of propagation had been used for the transmission path. This analysis was also carried out during magnetic storms where no vertical-incidence observations were obtainable.

551.524.1 **136**

Thermal Structures in the Lowest Layers of the Atmosphere.—R. J. Taylor. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, pp. 168–176.) Temperature variations recorded simultaneously at heights of 1·5, 4, 16, and 30 m were examined statistically; correlation between a pair of such records was greatest at a time-displacement depending on wind speed difference between the two heights. The evidence suggests organized thermal structures of great vertical extent.

551.594.5 **137**

The Geometry of Auroral Ionization.—R. S. Unwin. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 501–506.) "New evidence is presented which shows that v.h.f. radio echoes from auroral ionization are reflections from aspect-sensitive columns aligned with the earth's magnetic field."

551.594.5 **138**

Displacements of the Radiant Point during the Auroral Disturbance of September 22, 1957.—W. N. Abbott. (*Canad. J. Phys.*, June 1958, Vol. 36, No. 6, pp. 643–648.) Photographs of auroral corona enabled the position of the radiant point and its motion to be determined accurately. The motion was unrelated to the displacement of the magnetic zenith as determined from measurements on the ground.

551.594.5 : 523.164.3 **139**

Auroral Radiation at 500 Mc/s.—T. R. Hartz. (*Canad. J. Phys.*, June 1958, Vol. 36, No. 6, pp. 677–682.) Observations of r.f. noise emissions from aurorae were made during a type-A red auroral display on 21st–22nd October 1957. This unusual display was linked to a large flare on the sun some 30 h earlier. Theoretical considerations indicate that a large particle flux would produce such auroral radio emission.

551.594.5 : 621.396.11 **140**

A Comparison of Radio Echoes from the Aurora Australis and Aurora Borealis.—D. P. Harrison & C. D. Watkins. (*Nature, Lond.*, 5th July 1958, Vol. 182, No. 4627, pp. 43–44.) Observations of echoes on 4 m λ made at Halley Bay, Antarctica, for the period May–October 1957 are compared with contemporary observations made at Jodrell Bank. The great aurorae of the northern hemisphere are probably accompanied by greatly enhanced simultaneous activity of aurora australis.

551.594.5 : 621.396.11.029.6 **141**

V.H.F. Observations on the Aurora Australis.—T. J. Seed. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 517–526.) An investigation of radar echoes at 69 Mc/s is reported. The range exponent in the radar equation is calculated as $2\cdot15 \pm 0\cdot22$, and a maximum value of reflection coefficient was $1\cdot7 \times 10^{-7}$. Mechanisms of reflection are discussed, and only those involving column models are substantiated. See also 3466 of 1958 (Seed & Ellyett).

551.594.6 **142**

Some Properties of Lightning Impulses which Produce Whistlers.—R. A. Helliwell. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1760–1762.) Results obtained at Stanford, California, and Boulder, Colorado, show that a characteristic waveform, having an intense energy peak near 5 kc/s, is frequently associated with the impulse which produces a whistler. Whistler-producing discharges appear to be more frequent over sea than over land. Identification of the causative impulse should be based on waveform analysis as well as the time of occurrence.

551.594.6 **143**

Polarization of Atmospheric Pulses due to Successive Reflections from the Ionosphere.—B. A. P. Tantry & R. S. Srivastava. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 527-538.) Explanations are offered of the more complex patterns observed on a crossed-loop c.r.-tube direction finder for atmospherics. In addition to the normal straight-line display of an atmospheric pulse, elliptic traces or groups of straight lines within a small angle are produced. Elliptic traces are explained by polarization at ionospheric reflection, and groups by radiation from branch points in a nearly horizontal cloud-to-cloud lightning flash. See also 2420 of 1958 (Khastgir) and 3822 of 1958 (Horner & Khastgir).

551.594.6 : 523.746 **144**

Solar Activity and Whistler Dispersion.—G. McK. Allcock & M. G. Morgan. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 573–576.) The time delay between the occurrence of a local lightning flash and the beginning of the associated whistler is correlated with the monthly mean sunspot number. The correlation coefficient is greatest when the sunspot number is related to whistler measurements made 1·5 months later.

LOCATION AND AIDS TO NAVIGATION

621.396.93(94) **145**

A New Receiver for the Australian D.M.E. Beacon.—B. R. Johnson. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1957, Vol. 18, No. 11, pp. 423–430.) The receiver described operates at 206 Mc/s, has a triggering sensitivity of $3\,\mu$V and echo suppression by instantaneous a.g.c. Performance figures and the results of field trials are given. See also 3470 of 1958 (Stern).

621.396.933 **146**

Modified Transceivers Compute Distance.—H. Vantine, Jr, & E. C. Johnson. (*Electronics*, 12th Sept. 1958, Vol. 31, No. 37, pp. 94–98.) Two communications transceivers operating on a common frequency form an interrogator-responder system between an aircraft and the ground. The time delay between the transmitted and received pulses allows distances to be measured to within ± 0·1 mile. Circuit details are given.

621.396.96 : 551.501.8 **147**

Commercial Airborne Weather Radar.—A. W. Vose & F. V. Wilson. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 187–207.) Factors relating to the detection and penetration of atmospheric precipitation at microwave frequencies are discussed. Technical design features of both 5·6-cm weather penetration and 3·2-cm weather avoidance radars in present-day commercial aircraft are described.

621.396.96 : 621.396.822 **148**

A Proposed Technique for the Improvement of Range Determination with Noise Radar.—G. L. Turin. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1757–1758.) A proposal by Bourret [see 3833 of 1958 (Hochstadt)] is discussed and it is suggested that the true range cannot be marked unequivocally owing to the infinite average power in the processed noise component.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 **149**

The Identification of Luminescent Complexes of Cadmium Iodide and Lead Iodide.—G. Monod-Herzen & Nguyen Chung-Tu. (*C. R. Acad. Sci., Paris*, 14th April 1958, Vol. 246, No. 15, pp. 2247–2250.)

535.37 : 537.58 **150**

Cathodoluminescence of Strontium Oxide, Barium-Strontium Oxide, and Magnesium Oxide.—H. W. Gandy. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 764–771.) A study of the spectral and time decay properties of the cathodoluminescence of samples in the spectral range 220–600 mμ, for temperatures from 90°K to above room temperature.

535.376 **151**

Electroluminescence and its Applications.—J. N. Bowtell. (*J. Instn elect. Engrs*, Aug. 1957, Vol. 3, No. 32, pp. 454–459.) Characteristics of the more important coloured phosphors are given and the limitations in their application are discussed.

535.376 : 546.472.21 **152**

Effects of Impurities and Temperature on the Spectra of Electroluminescence.—R. Huzimura & T. Sidei. (*J. phys. Soc. Japan*, Sept. 1958, Vol. 13, No. 9, pp. 1064–1065.) Results are given of measurements on ZnS with impurities of Cu and Pb. See also 781 of 1956 (Alfrey & Taylor).

537.226/.228.1 **153**

Ferroelectric Crystals and Ceramics.—J. M. Herbert. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 168–192.) A comprehensive review of the structure, properties and applications, particularly of BaTiO$_3$. Over 30 references.

537.226/.228.1 **154**

Properties, Appraisal and Selection of Ferroelectric Materials.—B. Lewis. (*Brit. Commun. Electronics*, May 1958, Vol. 5, No. 5, pp. 332–337.)

537.226/.227 : 546.431.824-31 **155**

Some Experiments on the Motion of 180° Domain Walls in BaTiO$_3$.—R. C. Miller. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 736–739.)

537.227/.228.1 : 546.431.824-31 **156**

Primary Pyroelectricity in Barium Titanate Ceramics.—T. A. Perls, T. J. Diesel & W. I. Dobrov. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1297–1302.) It is established experimentally that the total pyroelectric effect in polycrystalline BaTiO$_3$ ceramic is opposite in direction to the piezoelectric contribution arising from volumetric changes caused by thermal expansion and contraction. There must therefore exist a primary pyroelectric effect which, at room temperature, is about twice the piezoelectric contribution.

537.227 **157**

Study of the Second-Order Ferroelectric Transition in Triglycine Sulphate.—S. Triebwasser. (*IBM J. Res. Developm.*, July 1958, Vol. 2, No. 3, pp. 212–217.) See also 1105 of 1957 (Matthias et al.).

537.227 : 546.431.824-31 **158**

Some Features of the Domain Structure of Ferroelectrics Shown by Electron-Microscope Observation.—G. V. Spivak, E. Igras & I. S. Zheludev. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Sept. 1958, Vol. 122, No. 1, pp. 54–57.) Investigation of BaTiO$_3$ single crystals, based on the different etch rates at the positive and negative ends of the electric dipole [see 2990 of 1955 (Hooten & Merz)]. Electron-microscope patterns obtained at magnifications up to 25 000 are shown.

537.227 : 546.431.824-31 **159**

Phenomenological Theory of Polarization Reversal in BaTiO$_3$ Single Crystals.—C. F. Pulvari & W. Kuebler. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1315–1321.) The theory, based on a study of polarization reversal, relates polarization P, switching field E, and switching time t, by an equation $P = f(E,t)$.

The switching resistance, the time and field dependence of the switching current at constant temperature, and the dependence of coercive field on applied field and frequency are readily obtained.

537.227 : 546.431.824-31 **160**

Ultra-Low-Velocity Component of Spontaneous Polarization in $BaTiO_3$ Single Crystal.—K. Husimi. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1379–1380.) The spontaneous polarization of $BaTiO_3$ consists of two components: the high-velocity component of 26 $\mu C/cm^2$ and the ultra-low-velocity component of 18 $\mu C/cm^2$. See also 3857 of 1958 (Husimi & Kataoka).

537.311.3 : [537.32 + 538.6 **161**

On the Galvanomagnetic, Thermomagnetic, and Thermoelectric Effects in Isotropic Metals and Semiconductors.—E. J. Moore. (*Aust. J. Phys.*, June 1958, Vol. 11, No. 2, pp. 235–254.) General expressions are obtained for metals and semiconductors which can be represented by the two-band model; the effects involve time-independent electric fields, magnetic fields, and thermal gradients. Emphasis is placed on experimentally and theoretically important coefficients. Formulae are given for isothermal galvanomagnetic effects in the presence of a h.f. electric field and a time-independent magnetic field.

537.311.31 : 539.23 **162**

Development of Electrical Resistance in Thin Films of Cobalt.—F. Savornin. (*C. R. Acad. Sci., Paris*, 14th April 1958, Vol. 246, No. 15, pp. 2230–2233.)

537.311.33 **163**

Intermetallic Semiconductors.—H. T. Minden. (*Sylvania Technologist*, Jan. 1958, Vol. 11, No. 1, pp. 13–25.) Methods of synthesis, purification and growth of compounds are described, properties are tabulated and some applications indicated.

537.311.33 **164**

Low-Field Mobility of Carriers in Nondegenerate Semiconductors.—M. S. Sodha & D. B. Agrawal. (*Canad. J. Phys.*, June 1958, Vol. 36, No. 6, pp. 707–710.) The theory of the low-field variation of carrier mobility is extended to include the effect of scattering by a single optical mode of lattice vibration. Numerical results obtained for *n*-type Ge indicate serious disagreement between theory and experiment.

537.311.33 **165**

Diffusion across a Semiconductor-Vapour Interface.—R. Bullough, R. C. Newman & J. Wakefield. (*Proc. phys. Soc.*, 1st Sept. 1958, Vol. 72, No. 465, pp. 369–379.) Theoretical solutions are derived for: (*a*) diffusion outwards of an impurity into a finite volume with or without simultaneous pumping; (*b*) diffusion inwards from a limited source of impurity, initially in the vapour phase, in a closed system. These results are discussed with reference to the diffusion of Group III and V elements across germanium and silicon surfaces.

537.311.33 : [546.28 + 546.289 **166**

Routine Crystal Orientation of Germanium and Silicon by High-Intensity Reflectograms.—G. H. Schwuttke. (*Sylvania Technologist*, Jan. 1958, Vol. 11, No. 1, pp. 2–5.) The equipment described measures tilt angle quickly with an accuracy of ±0·5° of arc.

537.311.33 : 546.28 **167**

Infrared Spectra of Heat Treatment Centres in Silicon.—H. J. Hrostowski & R. H. Kaiser. (*Phys. Rev. Lett.*, 15th Sept. 1958, Vol. 1, No. 6, pp. 199–200.)

537.311.33 : 546.28 **168**

The Preparation of Single-Crystal Ingots of Silicon by the Pulling Technique.—E. Billig & D. B. Gasson. (*J. sci. Instrum.*, Oct. 1958, Vol. 35, No. 10, pp. 360–365.) A description of the construction and operation of a laboratory furnace for preparing monocrystals of Si by a method which allows close observation of the growth front of the ingot during its preparation.

537.311.33 : 546.28 : 621.317.331 **169**

An Apparatus for Measuring the Electrical Resistivity of Silicon.—Heasell, Howard & Timmins. (See 214.)

537.311.33 : 546.289 **170**

Effect of Crystal Growth Variables on Electrical and Structural Properties of Germanium.—F. D. Rosi. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 349–387.) Examination of dislocation etch-pit distribution revealed plastic deformation by octahedral slip in crystals grown by the Czochralski technique. The conditions affecting the dislocation density are thoroughly investigated. Additions of *n*- and *p*-type impurities within their limits of solid solubility had no significant effect on the dislocation density.

537.311.33 : 546.289 **171**

Surface States on Germanium.—G. Wallis. (*Sylvania Technologist*, Jan. 1958, Vol. 11, No. 1, pp. 6–12.) The properties of slow and fast surface states on etched Ge are reviewed and some methods of investigating the states are presented.

537.311.33 : 546.289 **172**

Experiment Showing the Influence of Surfaces on 1/*f* Noise in Germanium.—J. J. Brophy. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1377–1378.) An attempt to measure the effect of changing the surface/volume ratio on 1/*f* noise was unsuccessful because of other surface effects. It was shown that etched surfaces are noisier than sanded surfaces by a factor of about 40.

537.311.33 : 546.289 **173**

Normal Modes of Germanium by Neutron Spectrometry.—B. N. Brockhouse & P. K. Iyengar. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 747–754.) The frequency wave-number of the lattice vibrations in Ge which propagate in the symmetric [100] and [111] directions were obtained by studying energy distributions of neutrons scattered by a Ge single crystal. A description of the methods of ascertaining the character of the phonons is given together with a discussion of the results.

537.311.33 : 546.289 **174**

Piezoresistance in Heavily Doped *n*-Type Germanium.—M. Pollak. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 798–802.) Piezoresistance has been measured as a function of temperature in *n*-type Ge specimens with donor concentrations between 6×10^{15} cm^{-3} and 3×10^{19} cm^{-3}. The results obtained are explained on the basis of the accepted multivalley model, provided that statistical degeneracy is taken into account. An analysis of the degeneracy observed in the data provides strong evidence for the four-ellipsoid model of the conduction band.

537.311.33 : 546.289 **175**

On the Hot-Electron Effect in *n*-Type Germanium.—R. Stratton. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 157–161.) The variation of the mobility with an applied electric field is calculated, assuming that interelectronic collisions are predominant. The scattering of electrons by acoustic and by nonpolar optical modes of lattice vibrations is considered without restricting the treatment to cases where the mean electron energy is greater than the phonon energy. Good agreement is obtained between the theory and experimental data. See also 1768 of 1958.

537.311.33 : 546.289 **176**

Infrared Absorption of Photogenerated Free Carriers in Germanium.—L. Huldt & T. Staflin. (*Phys. Rev. Lett.*, 1st Oct. 1958, Vol. 1, No. 7, pp. 236–237.) Preliminary account of a study of surface recombination and hole concentration in Ge produced by the absorption of radiation of different wavelength ranges.

537.311.33 : 546.289 **177**

Effect of Chemical Impurities on X-Ray Integrated Intensities in nearly Perfect Germanium.—B. W. Batterman. (*Phys. Rev. Lett.*, 1st Oct. 1958, Vol. 1, No. 7, pp. 228–229.)

537.311.33 : 546.289 : 537.322.4 **178**

Seebeck Effect Fluctuations in Germanium.—J. J. Brophy. (*Phys. Rev.*, 15th Aug. 1958, Vol. 111, No. 4, pp. 1050–1052.) Fluctuations in the thermoelectric power of Ge single crystals have been observed. The Seebeck noise-power spectrum varies inversely with frequency and may be quantitatively predicted from current-noise measurements. The results indicate that carrier fluctuations having a 1/*f* spectrum persist even in the absence of net d.c. flow.

537.311.33 : 546.682.86 **179**

Electrical Conductivity in *p*-Type InSb under Strong Electric Field.—Y. Kanai. (*J. phys. Soc. Japan*, Sept. 1958, Vol. 13, No. 9, pp. 1065–1066.) Results indicate the presence of holes with mobilities ten times greater than that of electrons.

537.311.33 : 546.682.86 : 537.228.1 **180**

Piezoelectric Effect in Indium Antimonide.—J. H. Wasilik & R. B. Flippen. (*Phys. Rev. Lett.*, 1st Oct. 1958, Vol. 1, No. 7, pp. 233–234.) Note on qualitative observations of piezoelectric resonances at the first and third harmonics of thickness vibrations in two InSb plates, the measurements being taken at 78°K and 4·2°K where the resistances of the samples were large.

537.311.33 : 546.817.221 **181**

Mobility of Electrons and Holes in PbS, PbSe, and PbTe between Room Temperature and 4·2°K.—R. S. Allgaier & W. W. Scanlon. (*Phys. Rev.*, 15th Aug. 1958, Vol. 111, No. 4, pp. 1029–1037.) Hall coefficient and resistivity measurements were made on 29 single crystals between room temperature and 4·2°K. Hall mobilities were calculated from the Hall coefficients and resistivity data and were found to increase rapidly with decreasing temperature. Between room temperature and about 50°K the mobility behaviour was essentially intrinsic and varied approximately as $T^{-2\cdot2}$. A possible explanation of the large mobilities at low temperatures is discussed.

537.311.33 : 546.873.241 **182**

Magnetothermal Resistance and Magnetothermoelectric Effects in Bismuth Telluride.—A. E. Bowley, R. Delves & H. J. Goldsmid. (*Proc. phys. Soc.*, 1st Sept. 1958, Vol. 72, No. 465, pp. 401–410.) These effects have been measured in *p*-type and *n*-type Bi_2Te_3 at 77°K. Scattering laws for the charge carriers have been derived from the results for *p*-type material, but no simple law fits the results from the magnetothermal resistance effect in *n*-type material.

537.311.33 : 546.873.241 **183**

Galvanomagnetic Effects in *p*-Type Bismuth Telluride.—J. R. Drabble. (*Proc. phys. Soc.*, 1st Sept. 1958, Vol. 72, No. 465, pp. 380–390.) The resistivity components, Hall coefficients, and low-field magnetoresistance coefficients have been measured on single-crystal specimens at 77°K. The results are consistent with a many-valley form of the valence band. See also 2151 of 1958 (Drabble et al.).

537.311.33 : 621.314.63 **184**

Solid-State Diffusion applied to Semiconductor Devices.—M. Darmony & B. Dreyfus-Alain. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 37–49.) Theoretical and practical aspects of diffusion are considered and three techniques described. Some data of a prototype Si solar battery are given.

537.311.33 : 621.314.632 **185**

On Electron-Hole Transition in Point-Contact Solid Rectifiers.—V. G. Mel'nik, I. G. Mel'nik & S. S. Gutin. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Aug. 1958, Vol. 121, No. 5, pp. 852–854.) Investigation of physical phenomena occurring in Ge and Si point-contact diodes under pulse conditions. Oscillograms of the thermo-e.m.f. on heating to 500–600°C and *V*/*I* characteristics indicate the presence of two regions of different types of conductivity forming a *p-n* junction.

537.311.33 : 621.314.7 **186**

On Avalanche Multiplication in Semiconductor Devices.—H. L. Armstrong. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 97–104.) The current multiplication and resulting characteristics are derived in general terms for the depletion region, and the conclusions applied to an intrinsic depletion region, an abrupt junction and a linear junction. Avalanche multiplication in an undepleted semiconductor carrying heavy currents should be similar to that observed for depletion regions.

537.311.33 : 621.314.7 **187**

Metallographic Aspects of Alloy Junctions.—A. S. Rose. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 423–432.) The presence of oxide films on the Ge and In surfaces of large-area junctions leads to non-planar alloying, irregular Ge regrowth and unwetted areas. These effects can be minimized by using Ge crystals having very low dislocation densities, and by thermal deoxidation.

538.22 **188**

Multispin Axis Structures for Antiferromagnets.—W. L. Roth. (*Phys. Rev.*, 1st Aug. 1958, Vol. 111, No. 3, pp. 772–781.) Neutron diffraction intensities have been computed for antiferromagnetic spin arrangements in the rock-salt type structure. By assuming the crystal consists of domains with a common magnetic axis the experimental results can be accounted for quantitatively. See also 3541 of 1958.

538.22 **189**

Nonmagnetic Ions in an Antiferromagnetic.—K. F. Niessen. (*Philips Res. Rep.*, Aug. 1958, Vol. 13, No. 4, pp. 327–334.) An experimental method is described for determining the distribution of a relatively small number of foreign nonmagnetic ions between the two sublattices of an antiferromagnetic.

538.22 **190**

Magnetic Susceptibility of Copper-Nickel and Silver-Palladium Alloys at Low Temperatures.—E. W. Pugh & F. M. Ryan. (*Phys. Rev.*, 15th Aug. 1958, Vol. 111, No. 4, pp. 1038–1042.)

538.221 **191**

Direct Observation of Spin-Wave Resonance.—M. H. Seavey, Jr, & P. E. Tannenwald. (*Phys. Rev. Lett.*, 1st Sept. 1958, Vol. 1, No. 5, pp. 168–169.) The direct observation of spin-wave resonances in a 5 600 Å film of permalloy is described, the exchange constant being determined from the location of the spin-wave peaks. See also 3544 of 1958 (Kittel).

538.221 **192**

Ferromagnetic Domain Structure as Affected by the Uniaxial Anisotropy Induced in a 40 Per Cent Co-Ni Single Crystal.—M. Yamamoto, S. Taniguchi & K. Aoyagi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, Feb. 1958, Vol. 10, No. 1, pp. 20–33.) Changes in structure which occur after quenching lead to the conclusion that the perminvar-type magnetic properties are due to the stabilization of domain walls by the induced uniaxial anisotropy in face-centred cubic solid solutions with cubic anisotropy constants of any sign and body-centred cubic solid solutions with negative cubic anisotropy constants.

538.221 **193**

Temperature Dependence of Magnetic Properties of Silicon-Iron.—C. W. Chen. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1337–1343.) Permeability, coercive force, remanence, hysteresis and core losses of 3% singly oriented Si-Fe alloy were measured within the temperature range 30–700°C.

538.221 : 537.228.4 **194**

Direct Measurement of the Velocity of Propagation of a Ferromagnetic Domain Boundary in 'Perminvar'.—E. W. Lee & D. R. Callaby. (*Nature, Lond.*, 26th July 1958, Vol. 182, No. 4630, pp. 254–255.) Report of measurements made on thin rings of magnetically annealed perminvar, using the magneto-optic Kerr-effect technique described earlier [3415 of 1958 (Lee et al.)] with a double-slit light source.

538.221 : 537.312.62 **195**

On the Absence of Superconductivity in Ferromagnetics.—S. V. Vonsovskiĭ & M. S. Svirskiĭ. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Sept. 1958, Vol. 122, No. 2, pp. 204–207.) Theoretical investigation showing that superconductivity can only occur in ferromagnetic metals where electron interaction is very weak.

538.221 : [621.318.12 + 621.318.13] **196**
: 539.169

Radiation Effects in Magnetic Materials.—D. I. Gordon, R. S. Sery & R. E. Fischell. (*Nucleonics*, June 1958, Vol. 16, No. 6, pp. 73–77.) Percentage changes in magnetic properties are tabulated for seven commercial-type materials which had been irradiated.

538.221 : 621.318.13 **197**

Analytical Theory of the Behaviour of Ferromagnetic Materials.—G. Biorci & D. Pescetti. (*Nuovo Cim.*, 16th March 1958, Vol. 7, No. 6, pp. 829–842. In English.) Using a procedure based on the Preisach model of hysteresis [see e.g. 842 of 1957 (Feldtkeller & Wilde)], it is shown that every transformation in the *J-H* plane can be computed from the magnetization curve and saturation loop of the material. Experimental and theoretical results are in close agreement.

538.221 : 621.318.134 **198**

Microwave and Low-Frequency Oscillation due to Resonance Instabilities in Ferrites.—M. T. Weiss. (*Phys. Rev. Lett.*, 1st Oct. 1958, Vol. 1, No. 7, pp. 239–241.) Microwave and l.f. oscillations were observed in single-crystal yttrium iron garnet discs. An explanation of the effect is given.

538.221 : 621.318.134 **199**

Crystal Structure and Ferrimagnetism in $NiMnO_3$ and $CoMnO_3$.—W. H. Cloud. (*Phys. Rev.*, 15th Aug. 1958, Vol. 111, No. 4, pp. 1046–1049.) The position of the atoms in the unit cell has been determined and it is shown that there are two Ni-O-Mn configurations favourable to magnetic superchange interaction.

538.221 : 621.318.134 **200**

Observation of Domains in the Ferrimagnetic Garnets by Transmitted

Light.—J. F. Dillon, Jr. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1286–1291.) Light transmitted through sections of ferrimagnetic garnets about 0·005 cm thick undergoes a Faraday nonreciprocal rotation of the plane of polarization. This property enables the magnetic domain structure, at temperatures below the Curie point, to be studied with a polarizing microscope.

538.221 : 621.318.134 **201**

Magnetoacoustic Resonance in Yttrium Iron Garnet.—E. G. Spencer & R. C. LeCraw. (*Phys. Rev. Lett.*, 1st Oct. 1958, Vol. 1, No. 7, pp. 241–243.)

538.221 : 621.385.833 **202**

The Variation with Temperature of the Magnetic Leakage Field in Cobalt.—M. Blackman & F. Grünbaum. (*Proc. roy. Soc. A*, 24th June 1958, Vol. 245, No. 1242, pp. 408–416.) Both the prism and hexagonal faces of an unmagnetized crystal were examined in the temperature range 20°–380°C by a new method using a divergent electron beam. The prism face was also studied from + 20°C to − 170°C. See also 1483 of 1958.

538.23 : 538.221 **203**

On the Effect of Coupling between Grains of Ferromagnetic Material having Hysteresis Properties.—L. Néel. (*C. R. Acad. Sci., Paris*, 21st April 1958, Vol. 246, No. 16, pp. 2313–2319.) The properties of an assembly of couples, each formed by two interacting ferromagnetic grains, are examined. It is shown that in such a system two successive cycles of a field H do not necessarily give rise to the same magnetization.

538.23 : 538.221 **204**

Creep of Asymmetric Hysteresis Cycles as a Function of the Number of Cycles Described.—Nguyen Van Dang. (*C. R. Acad. Sci., Paris*, 21st April 1958, Vol. 246, No. 16, pp. 2357–2359.) It is shown experimentally that, whatever the initial condition, creep is approximately proportional to $(\log n)^{\frac{1}{2}}$ provided the number of cycles n is greater than 20. This confirms Néel's theoretical work (see 3109 and 3110 of 1957).

548.0 : [549.514.51 + 546.289-31 **205**

Infrared Studies on Polymorphs of Silicon Dioxide and Germanium Dioxide.—E. R. Lippincott, A. Van Valkenburg, C. E. Weir & E. N. Bunting. (*J. Res. nat. Bur. Stand.*, July 1958, Vol. 61, No. 1, RP 2885, pp. 61–70.)

549.514.51 : 621.372.412 **206**

Thermoelastic Loss in Quartz Vibrators.—H. Iwasaki. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 257–279.)

537.311.33 : [537.32 + 536.21 **207**

Semiconductor Thermoelements and Thermoelectric Cooling. [Book Review]—A. F. Ioffe. (Translated from the Russian.) Publishers: Infosearch, London, 1957, 184 pp., 42s. (*Nature, Lond.*, 12th July 1958, Vol. 182, No. 4628, pp. 72–73.) A combination of two earlier works, on the basic theory of thermojunctions and their application to refrigeration, including examples of practical units for power supply.

MATHEMATICS

517.5 : 53.087 **208**

Experiments in the Smoothing of Data.—M. Lotkin. (*Quart. appl. Math.*, July 1958, Vol. 16, No. 2, pp. 169–173.) The effects of function smoothing and argument smoothing, and a combination of the two, are investigated for a few particular simple functions.

517.948.34 **209**

Errors in the Solution of Integral Equations.—P. Moon & D. E. Spencer. (*J. Franklin Inst.*, July 1957, Vol. 264, No. 1, pp. 29–41.) A novel method of solving integral equations is presented. An analytic solution is obtained by exponential approximation of the kernel, and to this solution is added the solution of an auxiliary integral equation.

519.2 : 621.396.822 **210**

Generalized Rayleigh Processes.—K. S. Miller, R. I. Bernstein & L. E. Blumenson. (*Quart. appl. Math.*, July 1958, Vol. 16, No. 2, pp. 137–145.) The generalized Rayleigh process R is defined by $R = ({X_1}^2 + {X_2}^2 + \ldots\ldots + {X_N}^2)^{\frac{1}{2}}$, where X_1, X_2, $\ldots X_N$ are N independent Gaussian processes. The following properties of R are calculated: the first-order distribution function, the joint probability density, the correlation function, and the three-dimensional distribution.

MEASUREMENTS AND TEST GEAR

537.228.1 : 548.0].001.4 **211**

Methods of Measurement of the Parameters of Piezoelectric Vibrators.—E. A. Gerber & L. F. Koerner. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1731–1737.) The theory underlying the measurements specified in Standard 57 I.R.E. 14.S1 (1788 of 1957) is discussed. See also 1437 of 1951 (Koerner).

621.3.018.41(083.74) : 529.786 **212**

Comparison of Caesium Frequency Standards of Different Construction.—L. Essen, J. V. L. Parry, J. H. Holloway, W. A. Mainberger, F. H. Reder & G. M. R. Winkler. (*Nature, Lond.*, 5th July 1958, Vol. 182, No. 4627, pp. 41–42.) The frequency of two 'atomichrons', Nos. 111 and 117, and one experimental tube No. 857 *X* 1 operated with the electronics and servo system of No. 111, have been compared at Teddington with the N.P.L. standard. Final results of the comparison, with corrections applied for errors in the servo and electronics system and for asymmetry of resonance, show the mean frequency difference to be respectively, (2·2 ± 1·4), (3·2 ± 1·4) and (1·5 ± 1·4) in 10^{10}.

621.317.3 : 538.632 : 549.351.12 **213**

Measurement of the Hall Effect in Anisotropic Media by the Point Method.—M. Wintenberger. (*C. R. Acad. Sci., Paris*, 21st April 1958, Vol. 246, No. 16, pp. 2366–2369.) Valdes' four-probe method may be applied to the measurement of Hall effect e.g. in $CuFeS_2$. See also 2827 of 1957.

621.317.331 : 537.311.33 : 546.28 **214**

An Apparatus for Measuring the Electrical Resistivity of Silicon.—E. L. Heasell, N. R. Howard & E. W. Timmins. (*B.T.-H. Activ.*, July/Aug. 1958, Vol. 29, No. 4, pp. 147–149.) Equipment based on the four-point probe method [1502 of 1954 (Valdes)] is described.

621.317.335.3 **215**

The Measurement at Low Frequencies of the Dielectric Constant of Conducting Liquids.—V. I. Little. (*Proc. phys. Soc.*, 1st Sept. 1958, Vol. 72, No. 465, pp. 441–446.) "The dielectric constants of aqueous solutions of the chlorides of Li, Na, K, Rb and Cs have been measured over the concentration range 10^{-4} to 10^{-2} normal at 25°C, using a novel form of electrometer."

621.317.35 : 621.391 **216**

Uniform Transient Error.—E. L. R. Corliss. (*J. Res. nat. Bur. Stand.*, July 1958, Vol. 61, No. 1, RP 2879, pp. 25–30.) "Equations describing error in power-level measurements of transients can be used to compute the design of analysers so as to distribute transient error in a way compatible with experimental requirements. In addition, consideration of a limiting power-discrimination factor provides a measure of the largest number of band-pass filters that can be overlapped on adjacent channels to yield meaningful information about a rapidly changing signal." See also 3168 of 1954 (Chang).

621.317.373 **217**

Coincident Slicer Measures Phase Directly.—Y. P. Yu. (*Electronics*, 12th Sept. 1958, Vol. 31, No. 37, pp. 99–101.) The phase angle between two voltages can be read with an absolute accuracy of 1° and a relative accuracy of ¼°. The output meter reading is independent of signal amplitudes in the range 0·3–70 V and of supply voltage variations of ± 20 %.

621.317.42 **218**

An Electrodynamic Magnetic Field Gradiometer employing a Microvibration Technique.—Y. L. Yousef & H. Mikhail. (*J. sci. Instrum.*, Oct. 1958, Vol. 35, No. 10, pp. 375–377.) The instrument utilizes the alternating electrodynamic forces experienced by a small probe coil fed by a low-frequency voltage.

Magnetic field inhomogeneity can be measured to an accuracy of at least 10^{-8} G/cm.

621.317.616 **219**

An Amplitude/Frequency Response Display using a Ratio Method.—H. L. Mansford, K. M. I. Khan & D. T. A. Margetts. (*Electronic Engng*, Sept. & Oct. 1958, Vol. 30, Nos. 367 & 368, pp. 541–544 & 595–597.) A method is outlined for balancing out the errors due to amplitude variations of the test signal from a frequency-sweep signal generator operating over a wide video-frequency range. A domestic television receiver is modified by turning the deflection coils through 90° to show a vertical raster. Brightness modulation from 0·5 μs pulses gives spots on the raster which trace the curves to be displayed. Full circuit details are given.

621.317.73.029.6 : 621.372.8 **220**

An Automatic Swept-Frequency Impedance Meter.—J. A. C. Kinnear. (*Brit. Commun. Electronics*, May 1958, Vol. 5, No. 5, pp. 359–361.) The instrument measures the waveguide impedance in an appropriate reference plane and displays it on a c.r. tube screen.

621.317.74 : 621.372.8 **221**

A Production Testing Equipment for Microwave Components and Systems.—J. Welsh. (*Brit. Commun. Electronics*, June 1958, Vol. 5, No. 6, pp. 438–439.) The equipment described is based on the rotary standing-wave indicator [203 of 1955 (Cohn)], but has no moving parts. The frequency band is 3·2 cm ± 5 % and the indicated s.w.r. is within ± 0·02 of that measured using a slotted-line indicator.

621.317.74 : 621.396.65 : 621.376.3 **222**

A Portable Instrument for Measurements on Intermediate-Frequency Level on Frequency-Modulated Microwave Radio Links.—J. W. A. van der Scheer. (*Tijdschr. ned. Radiogenoot.*, Nov. 1957, Vol. 22, No. 6, pp. 359–373. In English.) Measurements are carried out in the i.f. band in the ranges 60–80 or 95–115 Mc/s and cover amplitude and group-delay variations as a function of frequency, discriminator linearity, and reflection coefficients.

621.317.784.029.64 **223**

A Double-Vane Torque-Operated Wattmeter for 7 000 Mc/s.—S. Okamura, S. Kanzaki, S. Kurokawa & G. Kondo. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 157–163.) An instrument developed from the basic design of Cullen [1082 of 1953 (Cullen & Stephenson)] is described and compared with a barretter-type wattmeter.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.376 **224**

The Sylvatron: A New Application of Electroluminescence.—K. H. Butler & F. Koury. (*Sylvania Technologist*, Oct. 1957, Vol. 10, No. 4, pp. 98–101.) Three types of display device are described which incorporate a ceramic type of electroluminescent lamp [see 3636 of 1954 (Butler et al.)].

535.376.07 **225**

ELF — a New Electroluminescent Display.—E. A. Sack. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1694–1699.) A flat-screen display unit is described which uses ferroelectric ceramics with an electroluminescent layer to provide high brightness, flexible storage and fast writing speeds. Luminances greater than 25 foot-lamberts and contrast ratios of 50 : 1 have been obtained. Model screens with ten elements per inch and a scanning system for μs switching rates have been constructed.

551.508.822 **226**

Methods for the Calculation of Radiation Errors of Radiosondes.—W. D. Reshetow. (*Z. Met.*, Jan./Feb. 1958, Vol. 12, Nos. 1/2, pp. 46–50.)

621.362 : 621.385.2 **227**

Thermionic Diodes as Energy Converters: an Addendum.—Feaster. (See 324.)

621.362 : 621.385.2 **228**

Thermionic Energy Converter.—Hernqvist, Kanefsky & Norman. (See 325.)

621.373.44 : 615.475 **229**

A Versatile Stimulator.—R. H. Kay, C. G. Phillips & R. H. Teal. (*Electronic Engng*, Oct. 1958, Vol. 30, No. 368, pp. 575–578.) Conventional techniques are used to obtain various delayed positive- or negative-going pulse trains primarily intended for neuro-physiological research. The pulse width is variable from 0·1 to 100 ms and the train duration from 10 ms to 10 s. Full circuit diagrams are given. See also 1206 of 1950 (Attree).

621.375.2.029.4 : 612.014.421 **230**

The Sensitivity of Low-Frequency Valve Amplifiers for Electromyography.—A. Nightingale. (*J. sci. Instrum.*, Oct. 1958, Vol. 35, No. 10, pp. 366–371.) The sources of noise in l.f. amplifiers are discussed and experimental results are given for various valve types for the frequency range 2 c/s–16 kc/s.

621.383.4 : 612.84 **231**

Photovoltaic Pile.—I. Levin. (*Nature, Lond.*, 5th July 1958, Vol. 182, No. 4627, pp. 44–45.) A multiple-point voltaic cell in which all the individual electrodes are connected in series has been made by modifying the retinal type of cell described earlier (2846 of 1958).

621.387.464 **232**

Scintillation Counting—1958.—(*Nucleonics*, June 1958, Vol. 16, No. 6, pp. 54–62.) Review based on papers presented at the 6th Scintillation Counter Symposium, Washington, D.C., 27th–28th January, 1958.

PROPAGATION OF WAVES

621.396.11 : 550.372 **233**

Further Studies of the Influence of a Ridge on the Low-Frequency Ground Wave.—J. R. Wait & A. Murphy. (*J. Res. nat. Bur. Stand.*, July 1958, Vol. 61, No. 1, RP 2884, pp. 57–60.) "Computations are presented in graphical form for the perturbation of a plane wave by a semicylindrical boss on an otherwise flat ground plane of perfect conductivity. The height of the ridge is comparable to the wavelength. This is an extension of earlier work [2571 of 1957] on the semi-elliptical boss."

621.396.11 : 551.510.52 **234**

Some Aspects of Tropospheric Radio Wave Propagation—A. P. Barsis. (*Trans. Inst. Radio Engrs*, Oct. 1956, No. PGBTS-6, pp. 1–10. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 112.)

621.396.11 : 551.510.52 **235**

Antenna-to-Medium Coupling Loss.—H. Staras. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 228–231. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 895.)

621.396.11 : 551.510.535 **236**

Radio Scattering Expressed in Terms of the Spectrum of Turbulent Fluctuation in the Ionosphere.—K. Tao. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 95–103.) A theoretical expression for the scattering cross-section is compared with the results of Bailey's experiment (2581 of 1952), and a cross-section $\sigma = 2 \times 10^{-15}$ cm^{-1}, is obtained, taking the size of large eddies to be 1 000 m and their velocity to be 30 m/sec. The frequency dependence of the scattering power is given. See also 3626 of 1957.

621.396.11 : 551.510.535 : 523.7 **237**

Solar-Cycle Influence on the Lower Ionosphere and on V.H.F. Forward Scatter.—Ellyett & Leighton. (See 130.)

621.396.11 : 621.372.2 **238**

Return Loss: Part 3.—Roddam. (See 12.)

621.396.11 : 621.396.674.3 **239**

The Transient Behaviour of the Electromagnetic Ground Wave on a Spherical Earth.—J. R. Wait. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 198–202. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 894.)

621.396.11.029.55 **240**

A Long-Distance Pulse-Propagation Experiment on 20·1 Megacycles.—R. Silberstein. (*J. geophys. Res.*, Sept. 1958, Vol. 63, No. 3, pp. 445–466.) The experiment was performed in October 1956 between Sterling, Virginia, and Maui, Hawaii (7 647 km) to study m.u.f. determination and mode structure. Oblique-incidence records were made at Boulder, Colorado (2 370 km) as well as back-scatter measurements at the transmitter and vertical-incidence records along the route. Results showed greatly differing mode structures from day to day; *M* and *N* reflections and layer tilts were important. The long path was very sensitive to ionospheric conditions.

621.396.11.029.6 **241**

An Experimental Investigation of the Diffraction of Electromagnetic Waves by a Dominating Ridge.—J. H. Crysdale, J. W. B. Day, W. S. Cook, M. E. Psutka & P. E. Robillard. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 203–210. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, pp. 894–895.)

621.396.11.029.6 **242**

The Dependence of Microwave Radio Signal Spectra on Ocean Roughness and Wave Spectra.—C. I. Beard & I. Katz. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 183–191. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 894.)

621.396.11.029.62/.63 **243**

An Experimental Investigation of the Diffraction at V.H.F. and U.H.F. by Mountain Ridges.—M. Hirai, Y. Fujii & H. Saito. (*J. Radio Res. Labs, Japan*, July 1958, Vol. 5, No. 21, pp. 189–211.) Field-strength measurements have been made simultaneously at frequencies of 159 Mc/s and 600 Mc/s over a distance of 280 km for a period of a few weeks. The results are analysed in terms of (*a*) variations in time and (*b*) spatial variation of transmission loss at right angles to the great circle. Diffraction losses over mountain ridges are also measured and compared with theoretical losses due to ideal knife edges.

621.396.11.029.62/.63 **244**

Results of Experiment of Long-Distance Overland Propagation of Ultra Short Waves.—M. Onoe, M. Hirai & S. Niwa. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 79–94.) Results of a path loss test over 364 km at frequencies of 160 and 600 Mc/s are discussed. The observed annual median basic transmission loss is about 13 dB greater than the estimated value for atmospheric scattering. The difference in loss between v.h.f. and u.h.f. is about 18 dB indicating a third-power frequency law. The loss is about 12 dB larger in winter than in summer. There is a negative correlation between loss and atmospheric temperature. When the path traverses an anticyclone the loss is small and fading rate is small.

621.396.11.029.63 : 551.510.52 **245**

A Study of 468-Megacycle Tropospheric Scatter Propagation over a 289-Mile Path.—J. B. Atwood, G. B. MacKimmie, D. G. Shipley & G. S. Wickizer. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 321–333.) Hourly distributions of field strength at the receiver were of Rayleigh type whilst the monthly cumulative distributions of the hourly medians followed a log-normal curve. The typical fading rate varied from 5 to 30 fades/min. A comparison between two receiving sites showed that the site which had better foreground clearance had an 8·8 dB stronger field.

621.396.11.029.65 **246**

Experimental Measurement of the Absorption of Millimetre Radio Waves over Extended Ranges.—C. W. Tolbert & A. W. Straiton. (*Trans. Inst. Radio Engrs*, April 1957, Vol. AP-5, No. 2, pp. 239–241.) Measurements were made using path lengths up to 61 miles at 8·6 mm λ, and up to 7·2 miles at 4·3 mm λ.

RECEPTION

621.376.23 : 621.396.822 **247**

Optimum Filter Functions for the Detection of Pulsed Signals in Noise.—H. S. Heaps. (*Canad. J. Phys.*, June 1958, Vol. 36, No. 6, pp. 692–703.) The filter characteristic which yields the optimum ratio of integrated signal power to noise power ovcr a number of pulses is derived. The transfer function is shown to be realizable, and for a rectangular pulse is closely approximated by a third-order low-pass Butterworth filter. See also 3729 of 1958 (Heaps & McKay).

621.376.23 : 621.396.822 **248**

Detection of Asymmetric-Sideband Signals in the Presence of Noise.—T. Murakami & R. W. Sonnenfeldt. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 388–417.) A 'video-to-noise error ratio' is proposed for a more adequate quantitative evaluation of detector performance. Theoretical curves of carrier-to-noise ratio prior to detection, against this ratio after detection, are derived and their validity proved by experimental results. These curves indicate that the video-to-noise error ratio can be improved by about 11 dB if a product or synchronous detector is substituted for the normal envelope detector. The effect of impulse noise on detected noise is considered for asymmetric sideband signals. Product detectors eliminate the rectified noise envelope produced by envelope detectors. The use of the asymmetric-sideband system means that the impulse noise output can be at a high frequency relative to the signal frequency and may be separated from that signal.

621.376.232.2 **249**

Analysis of Diode-Detector Circuits for Signals with Asymmetrical Sidebands.—A. van Weel. (*Philips Res. Rep.*, Aug. 1958, Vol. 13, No. 4, pp. 301–326.) "A theoretical analysis is given of a diode detector stage, for a modulated signal with asymmetrical sidebands, of which the carrier is detuned with respect to the resonance frequency of the i.f. circuit. The equivalent circuit is found to be a three-port consisting of the impedances for upper- and lower-sideband frequencies and video frequency in series. Measurements show different overall characteristics in the case of equal positive or negativc detuning of the carrier frequency, which effect is caused by the asymmetrical shape of the diode-current peak."

621.396.62 **250**

Instability in Radio Receivers.—D. R. Bowman. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 514–519.) Methods of locating and eliminating self-oscillation due to various forms of coupling in r.f. and i.f. amplifiers are given.

621.396.62 : 621.376.3 **251**

F.M. Tuner uses Four Transistors.—H. Cooke. (*Electronics*, 1st Aug. 1958, Vol. 31, No. 31, pp. 72–73.) Only one variable tuning element is used in the single-transistor frequency-changer. The ratio detector has 700 kc/s peak separation, and an emitter-follower circuit provides audio amplification.

621.396.62 : 629.113 **252**

Car Radio Design.—F. Grimm. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 541–543.)

621.396.62.004.6 **253**

What Goes Wrong?—W. Oliver. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 522–523.) A classified analysis of 600 sound-broadcast receiver repairs.

621.396.62.029.62 **254**

The ARR3 Sonobuoy Receiver.—R. V. Taylor. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 544–549.) Conversion details for v.h.f./f.m. broadcast reception.

621.396.812.029.63 **255**

Diurnal Variation of Hourly Median Field Strength of 1 940-Mc/s Signal in the Foothills of Central Himalayas.—R. Vikramsingh, M. N. Rao, S. Singh & S. Uda. (*J. Instn Telecommun. Engrs, India*, June 1958, Vol. 4, No. 3, pp. 147–156.) Typical chart records of field strength, with corresponding ground level meteorological data for the period April 1957–March 1958, are presented.

621.396.812.3.029.6 **256**

Diurnal Variation in Intensity of Fading of V.H.F. Wave.—K. Hirao. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 105–107.) The day/night ratio of fading intensity decreases with the height of the point of intersection of the lines of sight from both the transmitter and receiver according to a ½- to ⅓-power law. The ratio is unity at about 1 000 m which appears to be the limiting height for atmospheric turbulence. See also 924 of 1957.

STATIONS AND COMMUNICATION SYSTEMS

621.391 **257**

An Error-Correcting Encoder and Decoder of High Efficiency.—J. H. Green, Jr, & R. L. San Soucie. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1741–

1744.) The applicability of regenerative shift-register sequences to error-correcting codes is demonstrated.

621.391 : 621.396.82 **258**
On the 'Minimum Loss Operation Time' for Short-Wave Communication: Part 1.—H. Shibata. (*J. Radio Res. Labs, Japan*, April 1958, Vol. 5, No. 20, pp. 143–151.) Theoretical treatment for the determination of optimum operation time on a statistical basis.

621.395.4 : 621.318.5 **259**
A High-Speed Data Signalling System.—E. A. Irland. (*Bell Lab. Rec.*, Oct. 1958, Vol. 36, No. 10, pp. 376–380.) In the system described, which is used in the SAGE (semi-automatic air-ground environment) project, binary digits are transmitted over voice-bandwidth telephone channels at rates of up to 1 600 bits/s. See also 3973 of 1958 (Michael).

621.396.2 : 551.510.52 **260**
Tropospheric Scatter Tests in Great Britain.—A. J. Wheeldon. (*Brit. Commun. Electronics*, June 1958, Vol. 5, No. 6, pp. 428–431.) An outline is given of the equipment used in the circuit operating between Start Point and Chelmsford. The performance of the system is assessed and future plans are discussed.

621.396.2 : 621.396.11 **261**
Difficulties Facing Long-Distance H.F. Communications in the Approaching Years.—R. J. Hitchcock. (*Brit. Commun. Electronics*, May 1958, Vol. 5, No. 5, pp. 340–344.) With several decades of low sunspot maxima, combined with high interference levels, circuit performance between the U.K. and the Commonwealth will deteriorate. By installing a cable link from Great Britain to the equatorial belt, and subsequent radio transmission, a good performance index is obtainable, taking advantage of high ionization levels in the equatorial region.

621.396.41 **262**
Mobile Radio System Provides 920 Channels.—F. Brauer & D. Kammer. (*Electronics*, 10th Oct. 1958, Vol. 31, No. 41, pp. 96–99.) Channels are spaced 50 kc/s apart in the frequency range 30–76 Mc/s. Local-oscillator frequencies are controlled by reference to crystals using three successive interpolations. Transmitter frequency is derived from the receiver local oscillator.

621.396.41 : 621.396.933.42 **263**
Single-Sideband, Present and Future.—R. Jeremy. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1957, Vol. 18, No. 10, pp. 363–371.) S.s.b. and independent-sideband systems are described, and problems regarding the use of such techniques for ground/air communications are discussed. See also 3983 of 1958 (Grisdale).

621.396.65 : 621.396.43 **264**
Radio Relay Systems and the C.C.I.F.—W. A. E. Quilter. (*Point to Point Telecommun.*, Oct. 1956, Vol. 1, No. 1, pp. 20–31.) C.C.I.F. recommendations for international trunk cable systems are examined. They cannot always be economically applied to radio relay systems, particularly with respect to noise levels. The hypothetical reference circuit recommended by the 1956 C.C.I.R. Conference in Warsaw is described. See also 2232 of 1958 (Mansfeld).

621.396.71 (492) **265**
Brief Survey of 30 Years' Activity in Short-Wave Broadcasting in the Netherlands.—A. J. Duivenstijn. (*Philips Telecommun. Rev.*, April 1958, Vol. 19, No. 3, pp. 97–103.)

621.396.712 (492) **266**
The New Netherlands World Broadcasting Centre at Lopik-Radio.—G. Radstake. (*Philips Telecommun. Rev.*, April 1958, Vol. 19, No. 3, pp. 104–111.) Transmitters and directional aerial arrays covering the frequency range 5·9–26·1 Mc/s are briefly described. See also 293 below.

SUBSIDIARY APPARATUS

621.314.5 : 621.314.7 **267**
Designing Transistor D.C. to A.C. Converters.—S. Schenkerman. (*Electronics*, 26th Sept. 1958, Vol. 31, No. 39, pp. 78–80.) Nomograms are presented for the design of the basic two-transistor saturable-core symmetrical circuit.

621.314.63 **268**
Silicon Power Rectifiers.—M. Sassier. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 139–151.) A technique is described for the construction of alloy-junction rectifiers.

621.314.63 **269**
Germanium and Silicon Industrial Rectifiers.—J. Lecorguillier. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 153–170.) A review of methods of construction and applications.

621.316.72 : 621.316.8 **270**
The Theory of Ballast Tubes or Barretters.—R. O. Jenkins. (*Brit. J. appl. Phys.*, Oct. 1958, Vol. 9, No. 10, pp. 391–394.) Results are given which explain the observed behaviour of various low-voltage barretters. A better combination than iron wire in hydrogen is unlikely to be found.

621.316.722.078.3 : 621.385.2 **271**
Stabilization by Zener Diodes.—J. Pereli. (*Wireless World*, Nov. 1958, Vol. 64, No. 11, pp. 537–538.) Elementary design principles are discussed.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 **272**
Closed-Circuit Television.—(*Overseas Engr*, Nov. 1958, Vol. 32, No. 371, pp. 116–118.) Industrial applications are reviewed.

621.397.24 **273**
Some Problems concerning the Choice of Cable Circuits for Television Transmission.—A. P. Bolle. (*Tijdschr. ned. Radiogenoot.*, Nov. 1957, Vol. 22, No. 6, pp. 321–327. In English.)

621.397.24 **274**
Experiments in Television over Telephone Cable Facilities.—C. R. Kraus. (*J. Franklin Inst.*, Jan. 1958, Vol. 265, No. 1, pp. 1–12.) The limitations of telephone line systems are examined generally and two short-distance demonstration circuits, using 250 kc/s bandwidths, are described.

621.397.24 : 621.396.65 **275**
Long Cable Links for Television.—D. W. Harling. (*Brit. Commun. Electronics*, June 1958, Vol. 5, No. 6, pp. 416–421.) A survey of techniques used for the transmission of video signals to switching, control and transmitting centres.

621.397.33 **276**
A New Narrow-Band Image Transmission System: Parts 1 & 2.—C. MacDonald. (*QST*, Aug. & Sept. 1958, Vol. 42, Nos. 8 & 9, pp. 11–15 . . 142, & 31–36 . . 148.) In the system described a c.r. tube flying-spot scanner and a photomultiplier are used to develop a 120-line picture once every six seconds from a photographic negative. No special equipment is needed for transmission or reception.

621.397.33 **277**
Stop-Go Scanning Saves Spectrum Space.—H. E. Haynes & D. T. Hoger. (*Electronics*, 26th Sept. 1958, Vol. 31, No. 39, pp. 84–88.) A method is described for increasing spectrum utilization for facsimile-type transmission, where graduations between white and black are not important, by using scanning that is halted at black/white transitions. Scanning velocity over the non-transition portions is determined by available signal/noise ratio of the channel.

621.397.5 : 535.623 **278**
Technical Standards for Colour Television.—J. W. Wentworth. (*Trans. Inst. Radio Engrs*, Sept. 1956, No. PGBTS-5, pp. 25–31.) A discussion of desirable characteristics with reference to F.C.C. standards.

621.397.61 : 535.623 **279**
The Vitascan Live Flying-Spot Colour Scanner.—J. H. Haines & G. R. Tingley. (*Trans. Inst. Radio Engrs*, Oct. 1956, No. PGBTS-6, pp. 11–25. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 112.) See also 1926 of 1957 (Mate).

621.397.62 : 535.623 : 621.385.832 **280**
Design and Development of the 21CYP22 21-Inch Glass Colour Picture Tube.—C. P. Smith, A. M. Morrell & R. C. Demmy. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 334–348.) The tube gives a brighter and more contrasted picture than its metal predecessor. Improvements are obtained by the use of a graded-hole shadow mask, an improved gun and pole-piece assembly, and increased filtering in the faceplate glass.

621.397.62 : 621.385.004.6 281

Unusual Tube Effects cause Circuit Troubles.—Babcock. (See 311.)

621.397.621.2 : 621.316.92 282

Protection Device for Stabilized Line Timebase Circuit.—B. G. Dammers, A. G. W. Uitjens & H. Heyligers. (*Electronic Applic. Bull.*, Jan. 1958, Vol. 18, No. 1, pp. 12–14.) A positive-going pulse, applied to the line output valve control-grid during fly-back period, prevents excessive rise of e.h.t. upon failure of the stabilizing circuit. See 3316 of 1957.

621.397.621.2 : 621.385.032.263 283

A New High-Transconductance Electron Gun for Kinescopes.—Schwartz. (See 322.)

621.397.621.2 : 621.385.832 284

TV Picture Tubes employing 110-Degree Deflection.—W. A. Dickinson & W. D. Schuster. (*Sylvania Technologist*, Oct. 1957, Vol. 10, No. 4, pp. 111–114.) Development problems concerning both c.r. tube design and associated scanning circuits are outlined.

621.397.7 285

Remote-Control Carrier Systems in Two-Way Closed-Circuit Educational TV.—J. R. Martin, G. W. Warnick & R. N. Vandeland. (*Elect. Engng, N.Y.*, April 1958, Vol. 77, No. 4, pp. 304–306.)

621.397.7 : 621.376.79 286

Design Improvements in High-Wattage Tungsten Filament Lamps for Motion-Picture and Television Studios.—L. G. Leighton & A. Makulec. (*J. Soc. Mot. Pict. Telev. Engrs*, Aug. 1958, Vol. 67, No. 8, pp. 530–533.) The life of high-wattage tungsten filament lamps is improved by the introduction into the bulb of a collector grid which reduces lamp blackening so that light output is maintained until filament failure occurs.

621.397.8 287

Measurement of Service Area for Television Broadcasting.—R. S. Kirby. (*Trans. Inst. Radio Engrs*, Feb. 1957, No. PGBTS-7, pp. 23–30. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4. p. 573.)

621.397.8 : 621.317.328 288

Measurement of Television Field Strength.—H. T. Head. (*Elect. Engng, N.Y.*, April 1958, Vol. 77, No. 4, pp. 298–302.) The preparation by T.A.S.O. [see e.g. 3277 of 1958 (Bowie)] of new propagation curves and data on the basis of extensive field-strength measurements is outlined.

621.397.826 289

The Use of Vertical Polarization to Solve U.H.F. Television 'Ghosting' Problems in a Shadowed Valley.—D. W. Peterson. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 208–215.) Field tests were made at a frequency of 640 Mc/s.

TRANSMISSION

621.396.61 : 621.375.2.026.445 290

A 60-kW Radio-Frequency Power Amplifier.—J. A. Gassner. (*Philips Telecommun. Rev.*, April 1958, Vol. 19, No. 3, pp. 130–134.) A grounded-grid push-pull circuit is used in the equipment described, which is designed for class-B operation in the frequency range 6–27 Mc/s.

621.396.61 : 621.396.4 291

Linear-Amplifier Transmitters.—W. J. Morcom. (*Point to Point Telecommun.*, Oct. 1956, Vol. 1, No. 1, pp. 7–19.) The transmitters described are used for independent-sideband operation in the frequency range 4–27·5 Mc/s.

621.396.61 : 621.396.664 292

Automatic Supervision of F.M. Transmitters.—J. A. van der Vorm Lucardie. (*Philips Telecommun. Rev.*, April 1958, Vol. 19, No. 3, pp. 135–141.) Remote control and automatic change-over equipment is described.

621.396.712 293

Type SOZ 294/00, High-Efficiency, 100-kW Short-Wave Broadcast Transmitter.—H. E. Eckhardt & A. G. Robeer. (*Philips Telecommun. Rev.*, April 1958, Vol. 19, No. 3, pp. 113–129.) Description of the equipment, with particular reference to the r.f. power stage which is of the grounded-cathode type with anode modulation. Spot frequencies cover the range 5·9–26·1 Mc/s and all valves are air cooled.

VALVES AND THERMIONICS

621.314.63 + 621.314.7 294

Some General Remarks on the Development of Techniques for Producing Semiconductor Devices and on the Improvement of their Characteristics.—J. M. Mercier. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 7–13.)

621.314.63 295

Theoretical Study of the Temperature Reached by a *p-n* Junction Operating as a Rectifier Element: Practical Consequences.—R. M. Henry. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 15–35.) Methods of measuring two essential constants in the formulae derived are given, from which the working limits of Ge and Si diodes can be calculated.

621.314.63 296

Point-Contact Germanium and Silicon Diodes.—R. M. Henry. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 63–81.) Parameters, characteristics and the manufacturing procedure are described.

621.314.63 297

Silicon Junction Diodes.—F. Provost. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 83–94.) A general report on the construction and properties of Si diodes including a note on Zener diodes.

621.314.63 298

Study on Annealing of Radiation-Induced 1/*f* Noise in Ge *p-n* Junction.—K. Komatsubara & U. Hashimoto. (*J. phys. Soc. Japan*, Sept. 1958, Vol. 13, No. 9, p. 1062.) A rapid variation of noise level and common-emitter current gain is observed, followed by a slow variation of reverse saturation current.

621.314.7 299

Evaluating the Effects of Temperature on Junction Transistors.—W. Bye. (*Brit. Commun. Electronics*, June 1958, Vol. 5, No. 6, pp. 440–442.) Methods of evaluating junction temperature, and its correlation with the α cut-off frequency are outlined, and test equipment is described.

621.314.7 300

Silicon Transistors.—R. M. Henry. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 115–125.) Alloy and grown-junction Si transistors are compared theoretically. The preparation and mounting of the latter type is described and characteristics are given.

621.314.7 301

Silicon Transistors by means of Alternate Doping.—B. Dreyfus-Alain & M. Darmony. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 51–62.) The construction of *n-p-n* type Si transistors, using the 'grown junction' principle, is described, and theoretical aspects of the problem are discussed.

621.314.7 302

Germanium Power Transistors.—R. Dubois. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 127–137.) Applications and performance limits are discussed.

621.314.7 303

The Tecnetron Principle.—(*Electronic Ind.*, July 1958, Vol. 17, No. 7, pp. 120–122.) Simple explanations with diagrams of the principles and application of the device are given. See also 3657 of 1958 (Aisberg) and 4005 of 1958 (Teszner).

621.314.7 : 621.372.5 304

A Four-Pole Analysis for Transistors.—Alcock. (See 48.)

621.314.7 : 621.375.4 305

Design of a Transistor for an I.F. Amplifier.—E. Martin. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1957, No. 27, pp. 95–113.) The development and manufacture of a transistor for use at 500 kc/s in receiver i.f. stages is described.

621.314.7 : 621.397.62 306

Bilateral Conductivity in Power Transistors.—I. G. Maloff. (*Electronic Ind.*, July 1958, Vol. 17, No. 7, pp. 82 . . . 122.) Positive and negative portions of collector current curves at constant base voltage are similar in most junction *p-n-p* power transistors, which are therefore useful in horizontal-deflection output stages in television receivers.

621.314.7.002.2 **307**

Precision Evaporation and Alloying.—R. J. Gnaedinger, Jr. (*Bell Lab. Rec.*, Oct. 1958, Vol. 36, No. 10, pp. 364–367.) The production of diffused-base *p-n-p* transistors depends on alloying by precision evaporation. The technique of evaporation at very low pressures is described.

621.383 : 546.289 **308**

The Germanium Photo-tetrode.—F. A. Stahl & G. Dermit. (*Electronic Ind.*, July 1958, Vol. 17, No. 7, pp. 64–66.) Details are given of the response of a particular *p-n-p* Ge transistor to light incident on the base region.

621.383.5 : 621.314.63 **309**

The Use of Germanium Junction Cells for Photoelectric Control Circuits.—R. F. Peart, T. B. Rymer & D. H. Tomlin. (*J. sci. Instrum.*, Oct. 1958, Vol. 35, No. 10, p. 383.)

621.385.001.2 : 623.451.8-519 **310**

Tube Developments for Guided-Missile Applications.—R. W. Slinkman. (*Sylvania Technologist*, Oct. 1957, Vol. 10, No. 4, pp. 102–105.) Design problems and their solution are discussed.

621.385.004.6 : 621.397.62 **311**

Unusual Tube Effects cause Circuit Troubles.—W. E. Babcock. (*Electronics*, 12th Sept. 1958, Vol. 31, No. 37, pp. 90–93.) A short survey of some less familiar faults in electronic valves and their effect on television receivers. Methods of eliminating the faults are described.

621.385.029.6 **312**

Water-Cooling of Low-Power Klystrons used in the Laboratory.—E. Niesen, R. W. Beatty & W. J. Anson. (*Rev. sci. Instrum.*, Sept. 1958, Vol. 29, No. 9, pp. 791–792.)

621.385.029.6 **313**

Beam Noise in Crossed Electric and Magnetic Fields.—R. P. Little, H. M. Ruppel & S. T. Smith. (*J. appl. Phys.*, Sept. 1958, Vol. 29, No. 9, pp. 1376–1377.) The large amount of noise which originates in the gun region may be reduced by operating the cathode under temperature-limited rather than space-charge-limited conditions, or by injecting the electrons into the crossed-field region with a substantial velocity. An explanation of the effect is put forward.

621.385.029.6 **314**

Effect of Collector Potential on the Efficiency of Travelling-Wave Tubes.—H. J. Wolkstein. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 259–282.) Methods are described for increasing the overall efficiencies of travelling-wave valves. These methods allow the collector voltage to be reduced, reducing the collector dissipation, without spoiling the r.f. performance. Secondary electron emission is increased but this effect can be minimized. A method is given for estimating the minimum potential to which an axially symmetric collector electrode can be depressed without reduction of beam current.

621.385.029.6 **315**

Distribution of Leakage Flux around a TWT-Focusing Magnet—a Graphic Analysis.—M. S. Glass. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1751–1756.)

621.385.029.6 **316**

Dispersion of Electron Velocities and Pass Band in Travelling-Wave Amplifiers.—R. Warnecke, O. Doehler & B. Epsztein. (*C. R. Acad. Sci., Paris*, 14th April 1958, Vol. 246, No. 15, pp. 2239–2242.) Experiment shows that the bandwidth of M-type valves is increased above the value predicted from simple theory because of dispersion in the electron beam. It is hoped to obtain similar results in high-power O-type valves by artificial means, notably using two electron beams from cathodes at different potentials.

621.385.029.6 **317**

On the Increase of Bandwidth in O-Type Travelling-Wave Valves.—J. Arnaud & R. Warnecke. (*C. R. Acad. Sci., Paris*, 21st April 1958, Vol. 246, No. 16, pp. 2359–2362.) Calculations are made of the increase in bandwidth obtained at the expense of gain in an O-type valve using two coupled or noncoupled beams. See 316 above.

621.385.029.6 **318**

Electron Trajectories in the Gun of an M-Type Carcinotron.—D. H. Davies & K. F. Sander. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 114–128.) The method uses a modified version of the automatic tracer described by Pizer et al. (3833 of 1956). The beam shape is investigated and optimum operating voltages suggested for the given geometry.

621.385.029.6 **319**

The Helitron Oscillator.—D. A. Watkins. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1700–1705.) A new type of voltage-tuned microwave oscillator requiring no magnetic field is described. Electron focusing is accomplished by balancing centrifugal force against a radial electric force and r.f. field interaction is both radial and angular. The electron beam travels in a helical path interacting with the r.f. field which is provided by an internal circuit structure. Experimental results show continuous voltage tuning from 1·2 to 2·4 kMc/s with a low starting current. Power outputs up to 10 mW have been obtained with low harmonic content. Theoretical studies to explain the operation of the device are not yet completely successful. See also 3436 of 1955 (Heffner & Watkins).

621.385.029.6 : 537.58 **320**

Current and Velocity Fluctuations at the Potential Minimum.—F. N. H. Robinson. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 152–156.) "The fluctuations in temperature-limited emission are shown to be uncorrelated. Space-charge smoothing leaves the velocity fluctuations unchanged but at low frequencies even moderate degrees of smoothing lead to complete correlation of the two fluctuations."

621.385.029.63 : 537.533 : 621.375.9 **321**

A Low-Noise Electron-Beam Parametric Amplifier.—R. Adler & G. Hrbek. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1756–1757.) A new electrode structure is described which provides parametric amplification of the fast wave on an electron beam. A noise figure of 1·3 dB was obtained. See also 2934 of 1958 (Adler).

621.385.032.263 : 621.397.621.2 **322**

A New High-Transconductance Electron Gun for Kinescopes.—J. W. Schwartz. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 232–243.) The characteristics of a gun employing two mesh-covered apertures, which act as space-charge grid and control grid, are presented. Peak currents of 800–1 600 μA for 2–5 V drive are obtained, and resolution is equivalent to conventional guns.

621.385.032.269 **323**

On the Determination of the Electrodes Required to Produce a Given Electric Field Distribution along a Prescribed Curve.—P. T. Kirstein. (*Proc. Inst. Radio Engrs*, Oct. 1958, Vol. 46, No. 10, pp. 1716–1722.)

621.385.2 : 621.362 **324**

Thermionic Diodes as Energy Converters: an Addendum.—G. R. Feaster. (*J. Electronics Control*, Aug. 1958, Vol. 5, No. 2, pp. 142–145.) The work of Moss (1983 of 1957) is extended to include diodes whose anode work function is less than their cathode work function, the space charge being negligible or neutralized. Output potentials of the order of 1 V may be obtained when the diodes are loaded for maximum power output.

621.385.2 : 621.362 **325**

Thermionic Energy Converter.—K. G. Hernqvist, M. Kanefsky & F. H. Norman. (*RCA Rev.*, June 1958, Vol. 19, No. 2, pp. 244–258.) The operation of a thermionic diode with Cs vapour is analysed and experimental results are given. Conversion efficiencies of about 10 % have been measured. See also 4024 of 1958 (Hatsopoulos & Kaye).

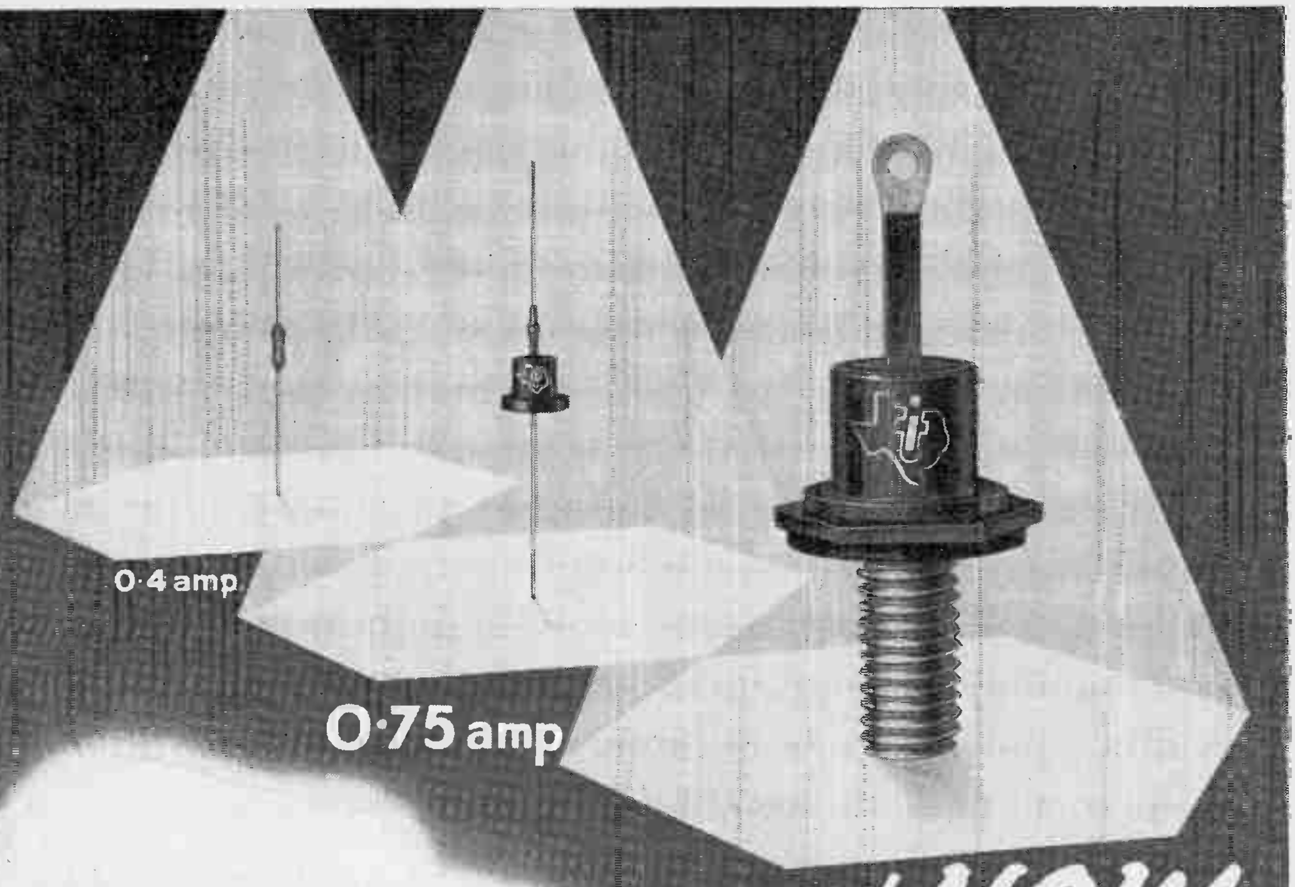
621.385.832 : 621.397.621.2 **326**

TV Picture Tubes Employing 110-Degree Deflection. — Dickinson & Schuster. (See 284.)

MISCELLANEOUS

621.3 : 537 : 55 **327**

Advances in Electronics and Electron Physics, Vol. 9. [Book Review]—L. Marton (Ed.). Publishers: Academic Press, New York, and Academic Books, London, 1957, 347 pp., $9. (*Nature, Lond.*, 21st June 1958, Vol. 181, No. 4625, pp. 1688–1689.) This volume is devoted to geophysics and includes chapters on the aurora borealis, negative ions, meteors, cosmic rays and radio wave propagation.



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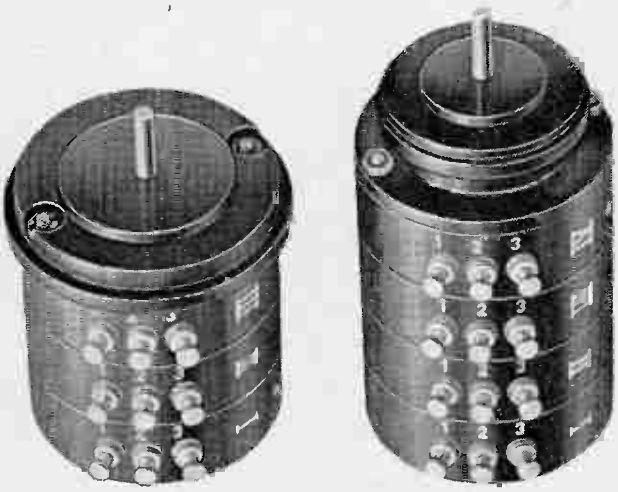

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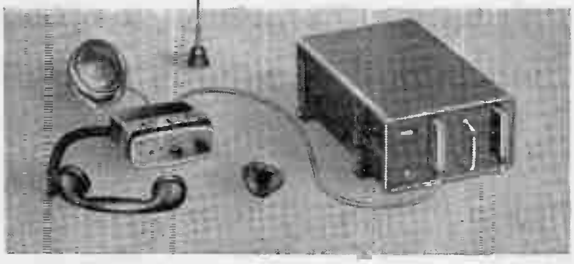

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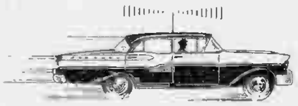

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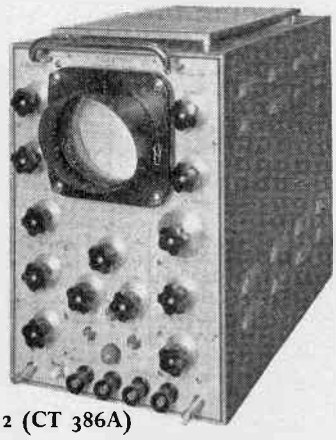

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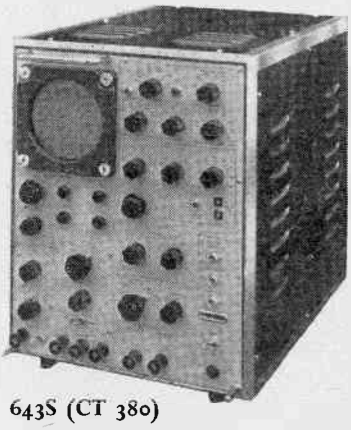

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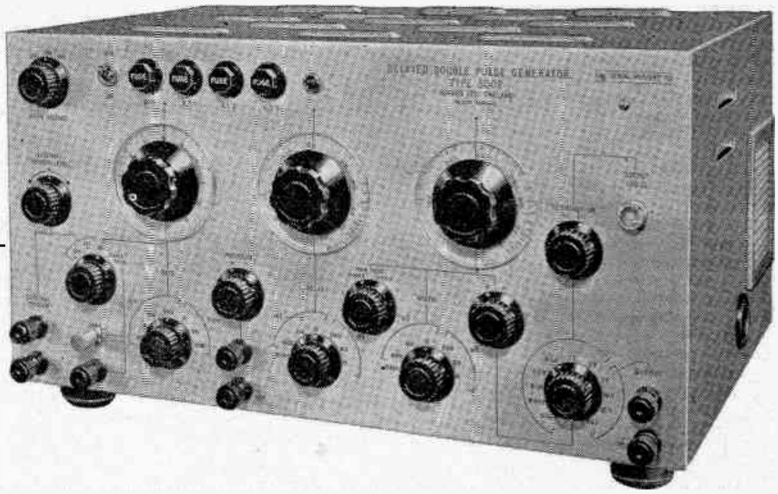

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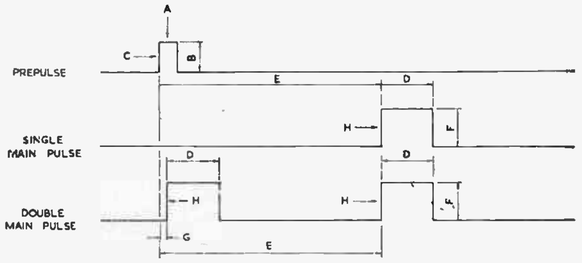

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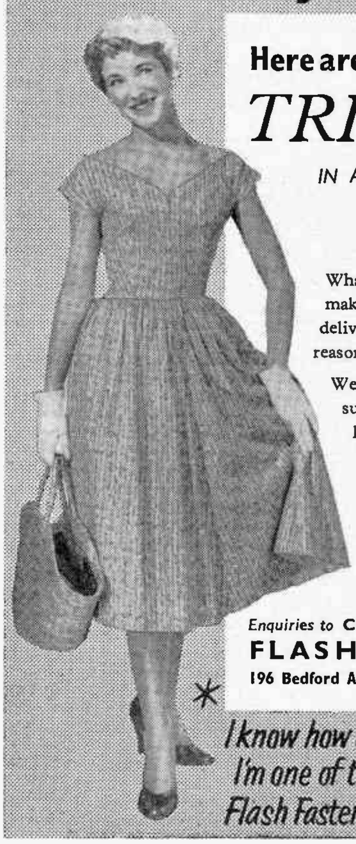

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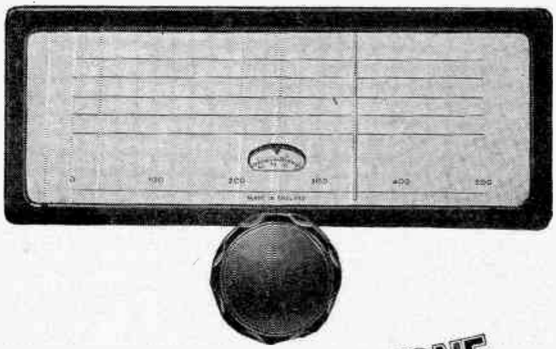

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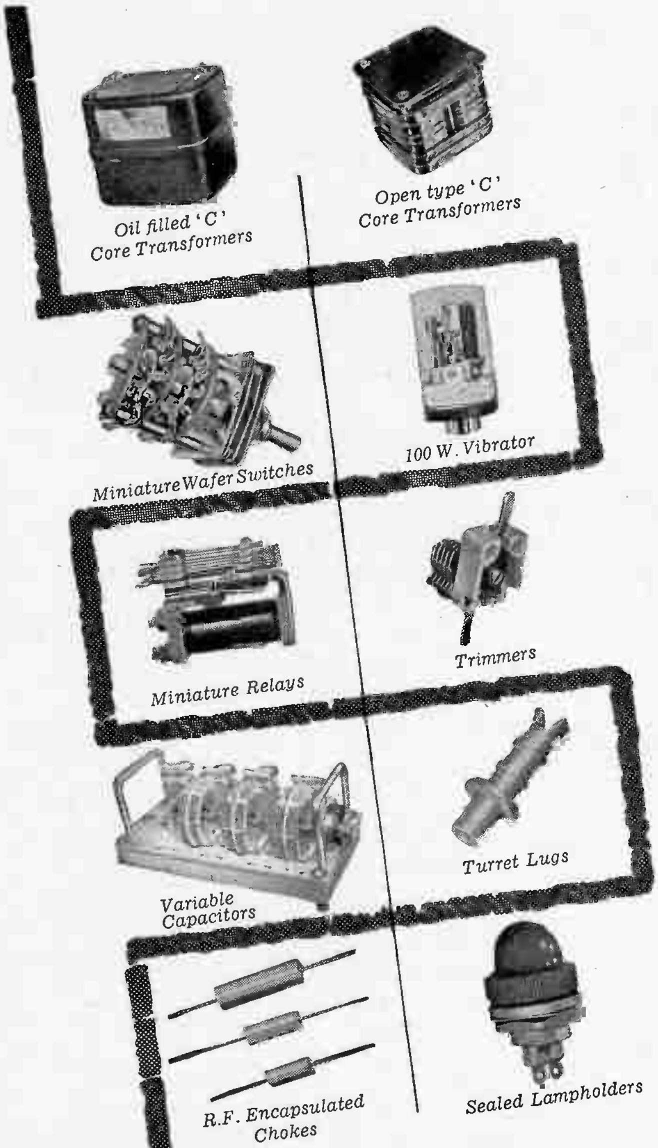

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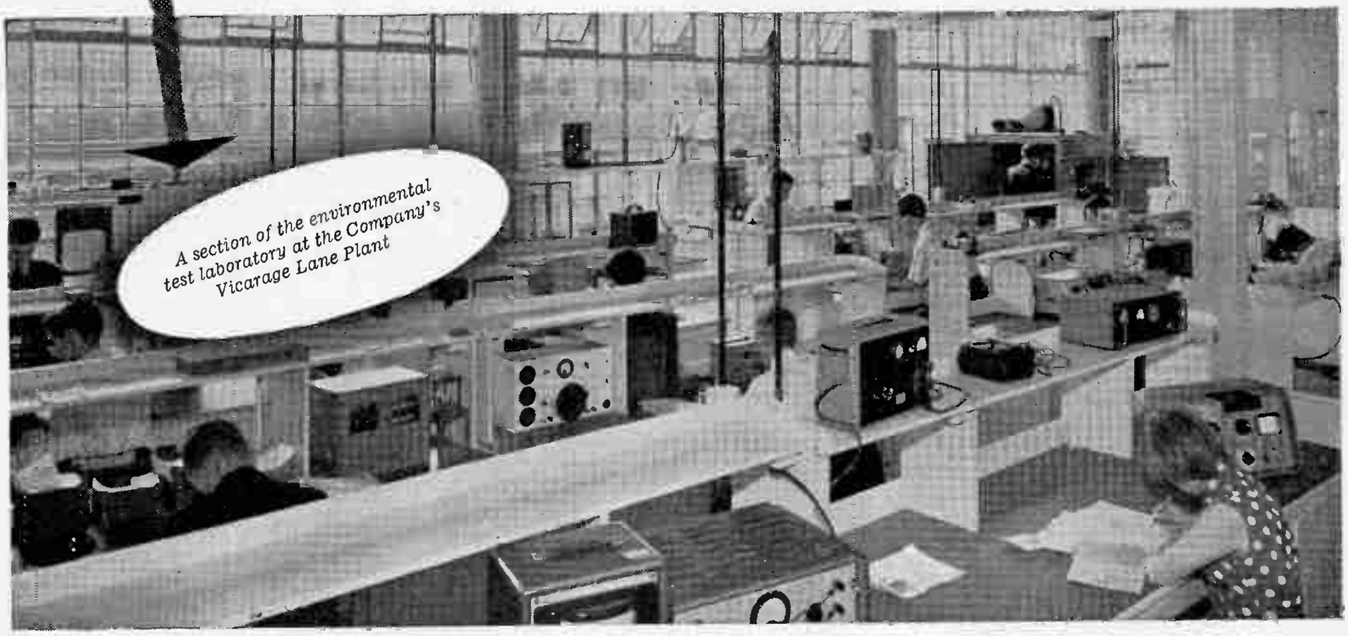

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WRITE to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote reference shown against the post applied for. [1281

GOVERNMENT OF SIERRA LEONE

Telecommunications Engineer, Post and Telecommunications Department

TO be responsible for supervising, maintaining, installing and developing telecommunications services generally under the direction of the Postmaster-General and the Engineer-in-Chief.

CONTRACT appointment in salary range £1,300 to £1,924 p.a. Gratuity of 15% of emoluments payable on satisfactory completion of contract. Free passages for officer and wife and up to three children under 18. Generous home leave. Quarters, if available, at moderate rent. Income tax at local rates. Free medical attention.

CANDIDATES under 45 years of age should be either A.M.I.E.E. or hold a university degree in electrical engineering. They must have a sound knowledge of the theory of telecommunications engineering generally and extensive practical experience of its application and be capable of controlling and training subordinate staff.

WRITE Director of Recruitment, Colonial Office, London, S.W.1, giving briefly age, qualifications and experience, quoting BCD 108/15/03. [1279

LABORATORY TECHNICIAN
GILLETTE RAZORS AND BLADES

(A Division of Gillette Industries Ltd.)

LABORATORY Technician, male or female, for the Electronic Section of our Research Laboratory. Age range 21-35. Education to O.N.C. level or equivalent in Electrical Engineering or Applied Physics, with practical experience of and an interest in the research and development of Electronic Equipment. Successful applicant will work initially at Isleworth but must be prepared to move to Reading in mid-1959.

APPLICANTS must give details of age, education, qualifications, experience and present salary to:

THE PERSONNEL MANAGER
(Ref. No. 28/58)
GILLETTE RAZORS AND BLADES
GREAT WEST ROAD
ISLEWORTH, MIDDLESEX [1280

UNIVERSITY COLLEGE, LONDON (Gower Street, W.C.1) requires Senior Workshop Technician combining knowledge of electronics with experience of machining for development and maintenance of apparatus in Psychology Dept. Salary up to £735 p.a. Application forms from Secretary, quoting Psychology/1. [1275

SITUATIONS VACANT

A NEW JOB IN THE NEW YEAR

Expansion of work on vital Defence projects
has created a number of attractive vacancies for
ELECTRONIC and **ELECTRO-MECHANICAL ENGINEERS**
at
COTTAGE LABORATORIES LIMITED
PORTSMOUTH ROAD, COBHAM, SURREY

THE work is interesting and unique and offers scope for both creative thinking and practical engineering to prototype stage using and developing the latest techniques in the electronic and electro-mechanical fields.

THESE positions are permanent and pensionable and carry starting salaries of up to £1,400 a year, depending on experience and qualifications. There are also opportunities for junior engineers.

WORKING conditions are ideal and the equipment and laboratories are well kept and modern in every detail. The company is situated between Esher and Cobham within its own gardens, in country surroundings, but with public transport routes passing the entrance. There are all the usual amenities of an up-to-date and progressive company.

WRITE or telephone for an appointment with the Chief Engineer (Cobham 3191). [1271

UNIVERSITY OF AUCKLAND
NEW ZEALAND
ELECTRONIC TECHNICIAN

APPLICATIONS are invited for the position of Electronic Technician in the Department of Physics. Applicants should possess Higher National Certificate or equivalent and be capable of assisting in the design and servicing of a wide variety of apparatus, especially nuclear. Experience with pulse and digital techniques or applications of transistors would be an advantage, and in industrial research or similar training preferred.

SALARY within scale rising from £875—£1,005 per annum, by annual increments, according to qualifications and experience. Superannuation rights available.

CONSIDERATION will be given to an assisted passage under the Immigration Scheme.

CONDITIONS of Appointment available from the Secretary, Association of Universities of the British Commonwealth, 36 Gordon Square, London, W.C.1, or from the Registrar, University of Auckland, P.O. Box 2553, Auckland, New Zealand.

APPLICATIONS close in New Zealand and London on the **15th February, 1959.** [1286

BELL PUNCH COMPANY LTD., OF UXBRIDGE

have vacancies for well experienced
automatic telephone
CIRCUIT DESIGNERS
and
DESIGN DRAUGHTSMAN

CAPABLE of adapting themselves to other interesting types of circuitry.

PERMANENT, well-paid and progressive posts are available to men with the right qualifications and experience.

APPLICANTS should write giving details of their education, training and experience in the work described above, to the Personnel Department or telephone Uxbridge 8211 Extension 35 to arrange a personal interview. [1282

INDEPENDENT TELEVISION AUTHORITY

APPLICATIONS are invited for the post of Engineer in the Headquarters Lines Group based in the London area.

APPLICANTS, who should hold a degree, H.N.C. or equivalent qualification in electrical engineering, should be familiar with transmission techniques and methods of performance, testing long vision and sound circuits.

SOME operational experience in television broadcasting would be an advantage.

THE salary scale for this appointment is £825—£1,160 per annum. Contributory pension scheme. Applications should be submitted to the Personnel Officer, 14 Princes Gate, London, S.W.7. [1272

SITUATIONS VACANT

ELECTRONIC PROCESS PLANNING ENGINEER

A well-established engineering company situated in West Middlesex has an opening for a qualified Process Planning Engineer at present engaged in the electronics engineering industry.

APPLICANTS must be well experienced in modern production methods as related to large quantity output, in the manufacture of component parts as used in radio, television and computers; must be accustomed to establishing production procedure in detail, the design of necessary equipment, jigs and tools, and supplying the initial instruction to production shops.

APPLICANTS should provide full information regarding technical education, training and industrial experience, and state names of previous employers and positions held, by letter. The salary envisaged would range from £800 to £1,000 per annum, according to qualifications and experience, with excellent prospects for advancement.

Box No. 0910. [1277

Research and Development
in a
Unique and Challenging Field
ELECTRONIC ENGINEERS
(Senior and Junior)
are required by
COTTAGE LABORATORIES LIMITED
PORTSMOUTH ROAD, COBHAM, SURREY

for the expansion of work on vital defence projects of a specialised and highly interesting nature.

THE positions are permanent and pensionable with excellent prospects of advancement. Starting salaries up to £1,400.

FIRST-CLASS working conditions, up to the minute equipment, five-day week and all the amenities of a well-established and progressive company.

WRITE to the Chief Engineer at the above address or telephone Cobham 3191. All applications treated in absolute confidence.

[1283

MINISTRY OF SUPPLY

ARMAMENT RESEARCH AND DEVELOPMENT ESTABLISHMENT, Fort Halstead, Sevenoaks, Kent, requires Physicists, Electronic, Electrical and Mechanical Engineers for work on guided weapons and fire control systems; Physical Chemists for explosives phenomenology; Metallurgists for problems in physical metallurgy. Appointments in grade (a) Senior Scientific Officer (min. age 26, at least three years' post grad. research experience) or (b) Scientific Officer. 1st or 2nd hons. degree or equiv. Salary ranges (a) £1,130—£1,330, (b) £595—£1,050 (male, in provinces). Rates for women same by 1961. Superannuation under F.S.S.U. Opportunity for those under 31 to compete for pensionable posts. Candidates should indicate fields of work in which interested. Forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1 (quote A454/8A). [1278

INSPECTOR of Lights required by Sarawak Government Marine Dept. either on probation to pensionable establishment or on contract with gratuity of 12½% of final monthly salary for each completed month of service. Salary scale (including Inducement pay) equivalent to £1,176 to £1,932 a year. Child allowance between £72 and £123 a year. Education allowance up to £280 a year. Outfit allowance £60. Free passages. Liberal leave on full salary. Candidates preferably under 35 and A.M.I.E.E. or A.M.I.Mech.E. and have a good all-round knowledge of electricity, especially in connection with the operation of lighthouse flashers (Chance/Londex), Reed and Klaxon Motors, Echo Sounding Gear, otherwise straightforward small unit diesel electric generation, current conversion, etc., a knowledge and experience of small diesel engines, working knowledge V.H.F. radio equipment and have a good head for heights up to 130 feet. Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2A/50186/EO. [1276

SITUATIONS VACANT

ELECTRO-ENCEPHALOGRAPHY RECORDIST, GRADE I

APPLICATIONS are invited from experienced technicians. The department also serves other hospitals in the district and a new research unit for neuro-surgery which opens shortly. Whitley Council salary scales and conditions apply. Apply, giving qualifications, experience, and quote two referees to Medical Superintendent, Parkside Hospital, Macclesfield. [1242

VACANCIES exist in a Government establishment in North Buckinghamshire for Principal Scientific and Senior Scientific Officers.
APPLICANTS should possess an Honours Degree in Physics or Engineering, or equivalent qualifications, and have at least three years' experience of working on Radio or Electronic equipment.
GRADING according to qualifications and experience.
SALARY, Principal Scientific Officer £1,370 rising to £1,950; Senior Scientific Officer £1,130 rising to £1,330.
CANDIDATES must be British subjects or citizens of the Irish Republic born within the Commonwealth, or in the Irish Republic of parents born within those territories.
CLOSING date for applications 10 days from the appearance of this advertisement.
WRITE giving age, qualifications and experience to Box No. 0824. [1274

UNIVERSITY COLLEGE LONDON (Gower Street, W.C.1) requires Electronics Assistant for variety of development and servicing work. Experienced in layout and wiring of circuits. Salary £450-£735. Application forms from Secretary, quoting Chemistry/18. [1270

THE post of Chief Designer at Painton's & Co. Ltd, Northampton, will shortly become vacant. Applications are invited from properly qualified men who have had several years' experience designing components for the Electronic Industry and capable of taking charge of a small drawing office. Apply to the Personnel Officer, Painton & Co. Ltd., Bembridge Drive, Kingsthorpe, Northampton. [1285

BRITISH PETROLEUM COMPANY LIMITED has a vacancy for an Electronic Engineer for work in the Instrument Group at its Research Station at Sunbury-on-Thames. Applications are invited from men with a University Degree in Physics or Electrical Engineering or equivalent. Some experience in the design and development of electronic instruments preferable. Work will include development of instruments for the control and testing of internal combustion engines, for process control, and for chemical and physical measurements. Age under 35. Salary according to age, qualifications and experience. Non-contributory Pension Fund. Assisted House Purchase Scheme. Removal expenses and settling-in allowance payable in certain cases. Luncheon club. Write giving full details, quoting H.3303A, to Box 5955, c/o 191 Gresham House, E.C.2. [1284

TUITION

BOROUGH POLYTECHNIC
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Department of Electrical Engineering and Physics
A COURSE OF SEVEN LECTURES ON
PULSE COMMUNICATION SYSTEMS

WILL be given by specialists from industry, the Scientific Civil Service and Universities, at 7 p.m. on Fridays, commencing on January 23rd. Further particulars and application forms from The Secretary, Borough Polytechnic, Borough Road, London, S.E.1. [1273

BOOKS, ETC.

"ADVANCED Theory of Waveguides." By L. Lewin. Sets out various methods of treating problems arising in work on this complex subject. The author has selected for discussion a number of topics as representative of the field in which the centimetere-wave engineer is engaged, many of the examples being concerned with the rectangular waveguide. 30s. net from all booksellers. By post 31s. from Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1.

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"RADIO Designer's Handbook" (4th Edition). Editor: F. Langford-Smith, B.Sc., B.E., Senior Member I.R.E.(U.S.A.), A.M.I.E.(Aust.). A comprehensive reference book, the work of ten authors and twenty-three collaborating engineers, containing a vast amount of data in a readily accessible form. The book is intended especially for those interested in the design and application of radio receivers or audio amplifiers. Television, radio, transmission and industrial electronics have been excluded in order to limit the work to a reasonable size. 50s. net from all booksellers. By post 52s. 3d. from Iliffe & Sons, Ltd., Dorset House, Stamford Street, S.E.1.

"RADIO Laboratory Handbook" (6th Edition). By M. G. Scroggie, B.Sc., M.I.E.E. This well-known practical work describing laboratory equipment and methods of operation has been entirely rewritten and greatly enlarged. Among the subjects considered are layout and furnishing of premises, methods of measurements, sources of power and signal measurements. There is a comprehensive reference section and many photographs, diagrams, graphs and tables. 25s. net from all booksellers. By post 26s. 9d. from Iliffe & Sons, Ltd., Dorset House, Stamford Street, S.E.1.

"TELEVISION Explained." By W. E. Miller, M.A.(Cantab), M.Brit.I.R.E. Revised by E. A. W. Spreadbury, M.Brit.I.R.E. The sixth edition of a book which assumes a knowledge of the ordinary sound radio receiver but no previous knowledge of television circuits. It is non-mathematical, written in simple language, and comprehensively illustrated by many diagrams and photographs. It will prove of great assistance to all students of television, to radio service engineers who wish to embark upon television work and want to understand the principles and circuits involved, and to knowledgeable owners of television receivers who would like to understand the working of their set. 12s. 6d. net from all booksellers. By post 13s. 5d. from Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1.

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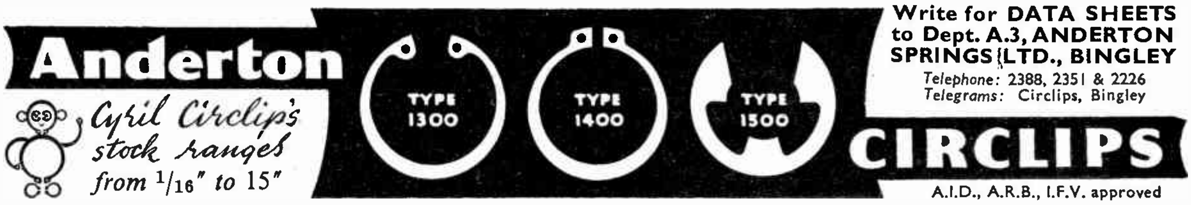

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Introducing another outstanding Ediswan Mazda valve, type 30PL1

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 30PL1, a triode-output tetrode for a.c./d.c. mains television receivers.

The tetrode may be used as the frame output valve with the triode operating as the saw tooth generator, or the tetrode may be used as the sound output valve with the triode as audio amplifier.

Maximum overall length (mm)	67.5
„ diameter	22.2
„ seated height	60.5
Heater volts	13
Heater current (amps)	0.3

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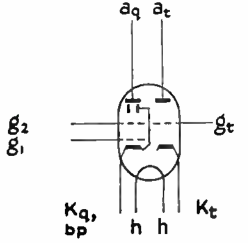

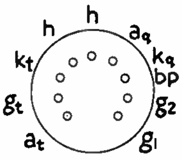

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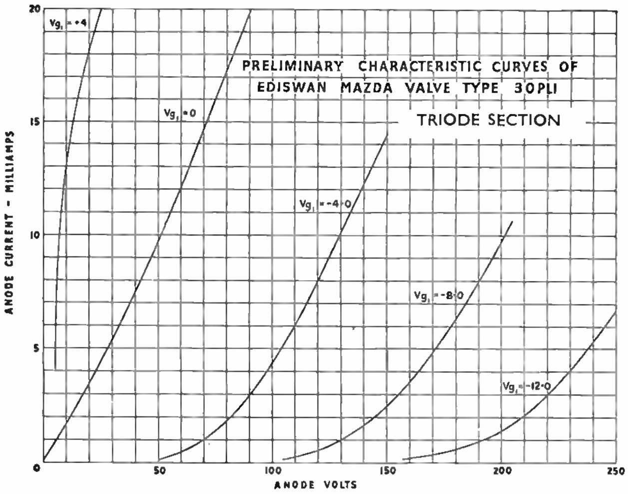

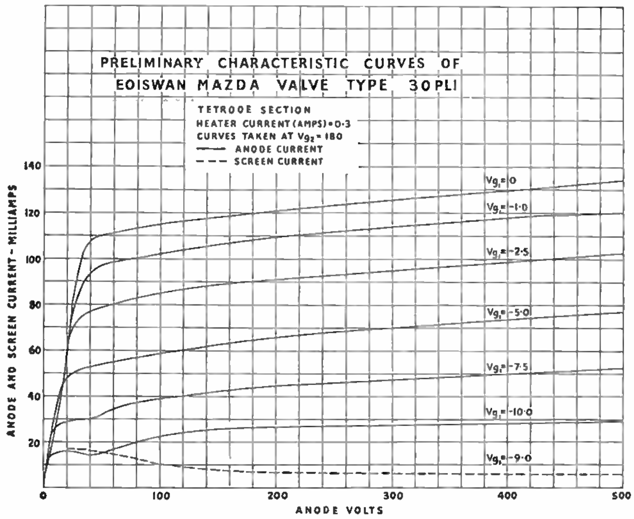

MAXIMUM DESIGN CENTRE RATINGS

	Tetrode	Triode
Anode Volts	250	250
Screen Volts	250	
Anode Dissipation (watts)	5.5*	2*
Screen Dissipation (continuous) watts	1.5	
Screen Dissipation (Speech & Music) watts ...	2.2	
Heater to Cathode Volts (r.m.s.)	150†	

**Total anode dissipation of the two sections must not exceed 6 watts.*
†Measured with respect to the higher potential heater pin.

TRIODE CHARACTERISTICS

Anode Volts	200
Anode Current (mA)	10
Grid Volts	—8
Mutual Conductance (mA/V)	3.4
Amplification Factor	16

TETRODE CHARACTERISTICS AND OPERATING CONDITIONS

Class A Audio Output Stage

	Fixed Bias	Self Bias
Anode Supply Volts (measured anode to earth)	170	180
Screen Supply Volts (measured screen to earth)	180	190
Grid Bias Volts Applied	—9.6	0
Cathode Self Bias Resistance (ohms)	0	270
Quiescent Anode Current (mA) ...	28	28
„ Screen „ „ ...	6.5	6.5
Mutual Conductance (mA/V) ...	6.0	6.0
Anode Load Resistance (ohms) ...	6000* 5300†	7000* 6200†
Power Output (watts)	2.0* 2.35†	1.85* 2.2†
Grid Input Swing (Volts r.m.s.) ...	3.3* 3.9 †	3.1* 3.8†

**For 5% third harmonic, and second harmonic not exceeding 5%.*
†For 7% third harmonic, and second harmonic not exceeding 7%.

FRAME TIME BASE OUTPUT STAGE

Allowance must be made in circuit design, for valve spread and deterioration during life, in addition to component variation. Values of total tetrode peak anode current available, for a new average valve, and at the assumed end of life point on any valve, are as follows:—

	V_{g_2}	V_{g_1}	V_a	I_a (mA)
Average New Valve	170	—1	50	88
	190	—1	55	104
Assumed End of Life Conditions ...	170	—1	50	57
	190	—1	55	67

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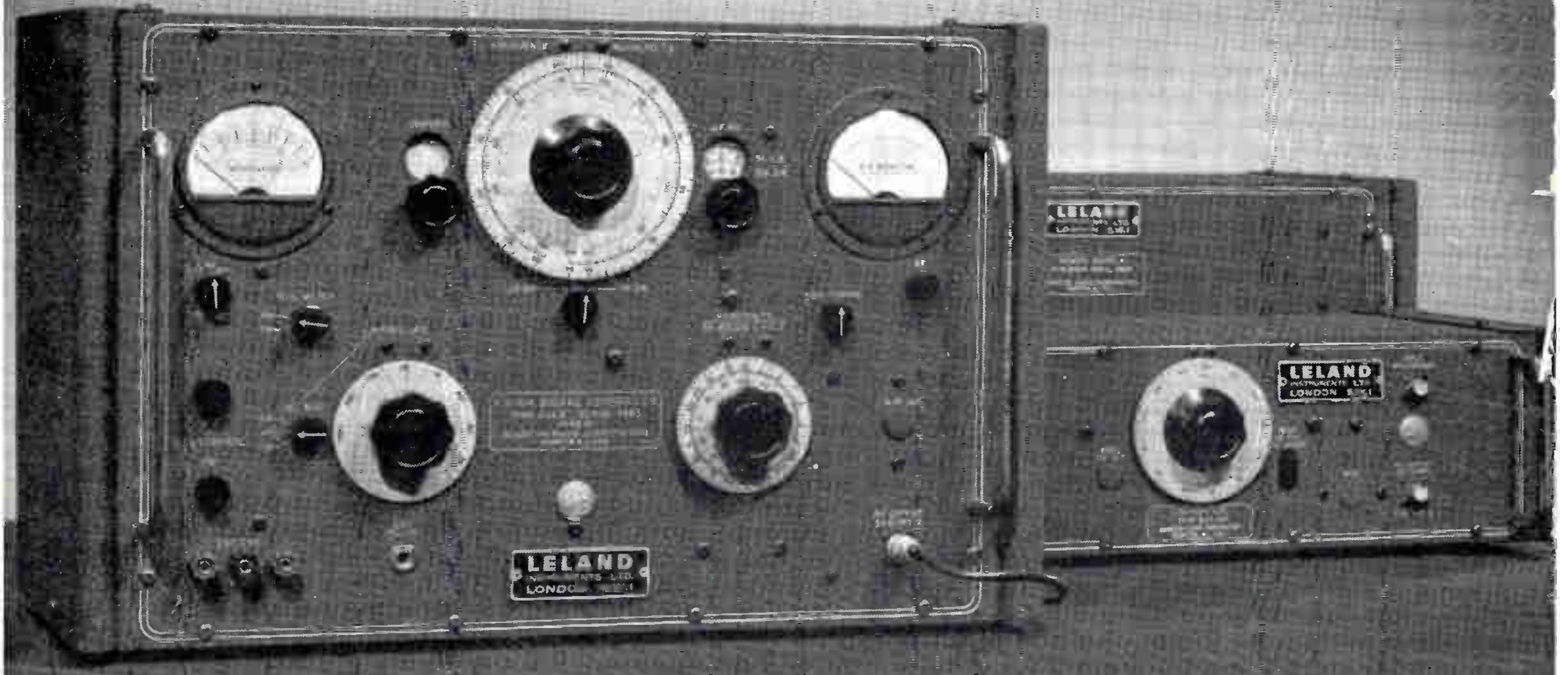
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