ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Combined Limiter and Discriminator

Temperature Measurement with Thermistors

Transmission-Line Low-Pass Filters

Amplifier Low-Frequency Compensation

Three shillings
and sixpence

MARCH 1958 Vol 35 *new series* No 3

Trade Mark

U.H.F. MEASURING EQUIPMENT

Type 1602-B U.H.F. Admittance Meter

No engineer concerned with impedance measurements from 41 Mc/s to 1500 Mc/s can afford to be without this *unique* Bridge.

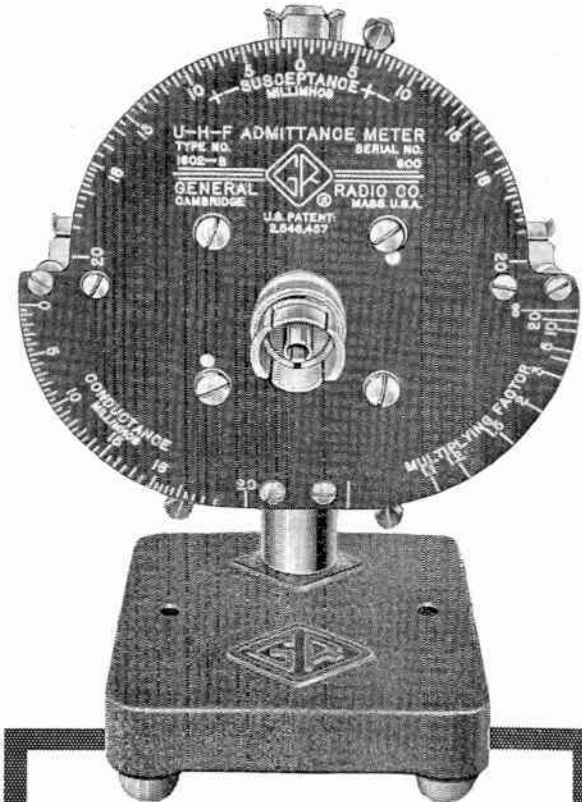
As a null instrument it can be used to measure the conductance and susceptance of an unknown impedance by direct reading of the scales. By connecting the unknown impedance through a 50 ohm line one or more odd quarter waves in length, the scales read directly in terms of resistance and reactance.

The Bridge can also be used as a comparator to indicate the degree of inequality between two admittances. In addition, as a direct reading device it can be used to determine the magnitude of the reflection coefficient of a coaxial feeder, or the magnitude of an unknown impedance, from the ratio of output voltages read on the detector meter. Balanced impedances can also be measured with the aid of the "G.R." Type 874-UB "Balun".

Owing to the *unique* coaxial form of the bridge arms and the use of the matched coaxial connectors "G.R." Type 874 throughout, any uncertainties regarding reflections (and thereby errors) at the vital points of connection are completely eliminated.

There are no sliding connections to cause intermittencies since the conductance, susceptance and multiplying arms merely control the rotation of small coupling loops within the coaxial arms of the bridge. A further *unique* feature is the independence with frequency of the susceptance readings.

Additional apparatus required consists of a suitable range Oscillator or a Signal Generator, and a sensitive, well shielded receiver as the detector. If the user does not already possess these, suitable instruments are available from the complete "GENERAL RADIO" range of measurement instruments, described in their 258-page current Catalogue "O", available on application.



BRIEF CHARACTERISTICS

FREQUENCY RANGE: 41 to 1500 Mc/s. This can be extended down to 10 Mc/s by the use of a correction factor, which is a function of frequency. (A Chart is provided).

ACCURACY: For both conductance and susceptance (up to 1000 Mc.): from 0 to 20 millimhos $\pm$ (3% + 0.2 millimho) from 20 to ∞ millimhos $\pm(3\sqrt{M}\% + 0.2$ millimhos) where M is the scale multiplying factor. Above 1000 Mc, errors increase slightly, and, at 1500 Mc, the basic figure of 3% in the expression above becomes 5%. For matching impedances to 50 ohms, the accuracy is 3% up to 1500 Mc.

DIMENSIONS: 7½" × 5½" × 5½".

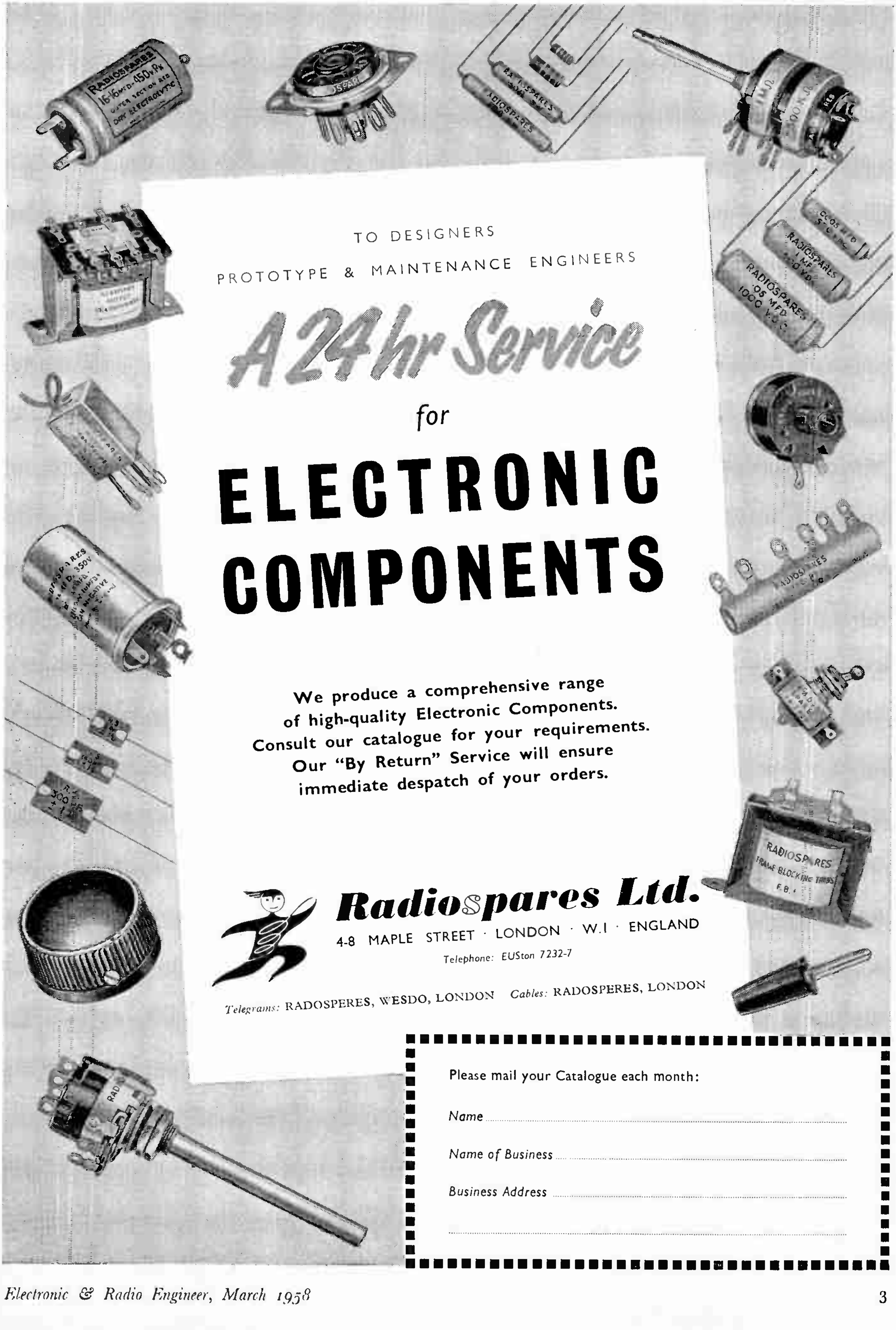
NET WEIGHT: 8¼ lbs.

REASONABLY PRICED: £177.0.0. net, delivered (U.K. only), complete with all basic accessories, all duties paid.

Claude Lyons Ltd.

76 OLDHALL STREET · LIVERPOOL 3 · LANCS · TELEPHONE: CENTRAL 4641
VALLEY WORKS · HODDESDON · HERTS · TELEPHONE: HODdesdon 3007-8-9

CL42

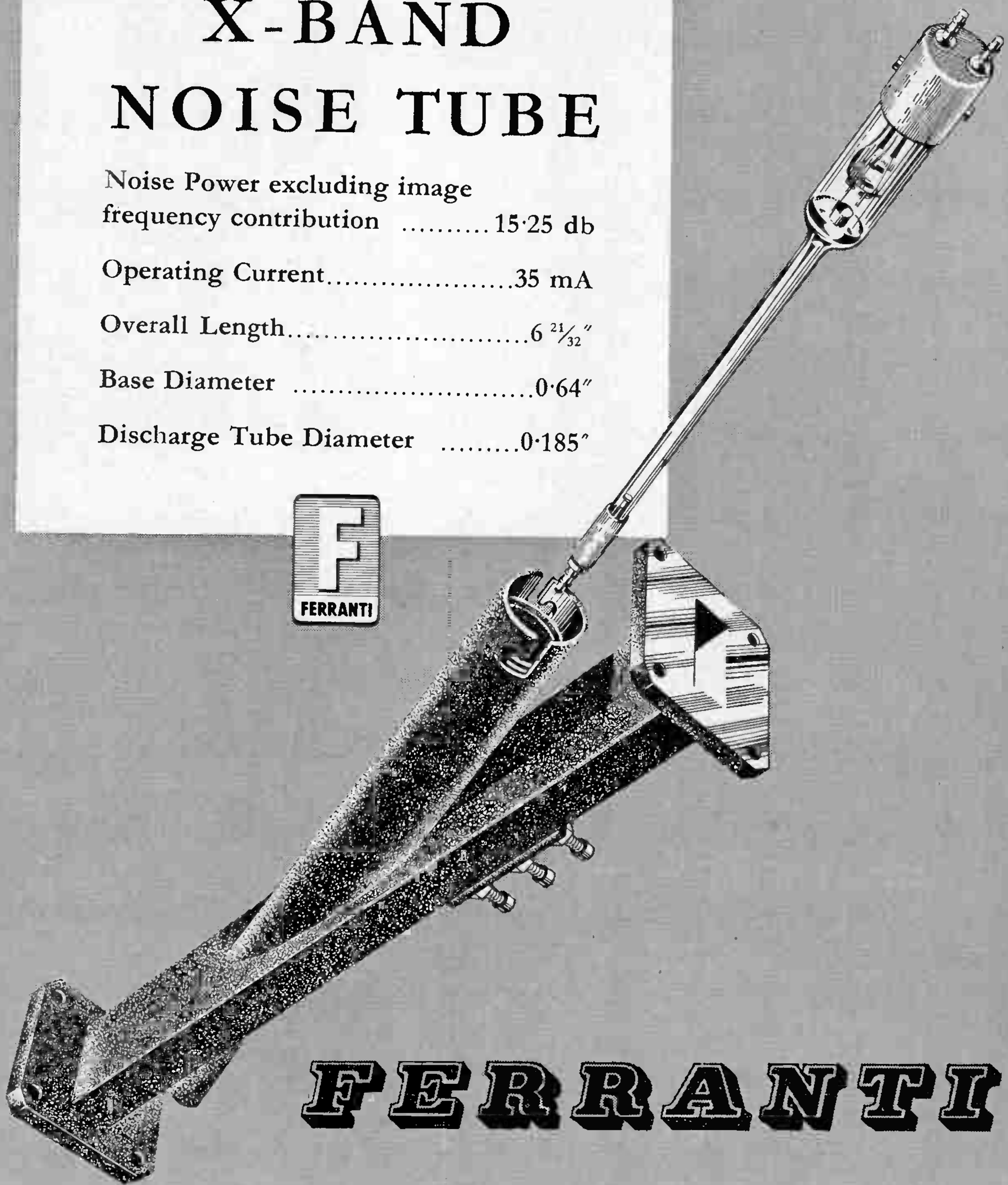

TO DESIGNERS
PROTOTYPE & MAINTENANCE ENGINEERS
A 24 hr Service
for
ELECTRONIC COMPONENTS
We produce a comprehensive range of high-quality Electronic Components. Consult our catalogue for your requirements. Our "By Return" Service will ensure immediate despatch of your orders.
Radiospares Ltd.
4-8 MAPLE STREET · LONDON · W.1 · ENGLAND
Telephone: EUSton 7232-7
Telegrams: RADOSPERES, WESDO, LONDON Cables: RADOSPERES, LONDON
Please mail your Catalogue each month:
Name
Name of Business
Business Address

Design Approved

X-BAND NOISE TUBE

Noise Power excluding image frequency contribution	15·25 db
Operating Current	35 mA
Overall Length	6 21/32″
Base Diameter	0·64″
Discharge Tube Diameter	0·185″

FERRANTI



FERRANTI

FERRANTI LTD. · FERRY ROAD EDINBURGH 5 Tel. Granton 89181

Visit our Stand No. M9 at the A.S.E.E. Exhibition, March 25th to 29th

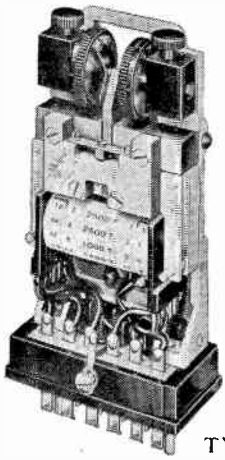
ES/T38

It doesn't matter whether you call it . . .

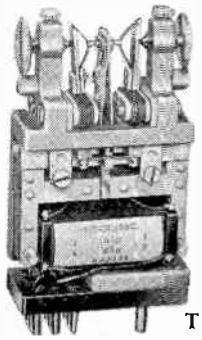
the **CARPENTER** Polarized Relay

or the Carpenter **POLARIZED** Relay

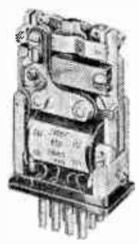
or the Carpenter Polarized **RELAY**



TYPE 3



TYPE 4



TYPE 5

TYPE 51

TYPE 6

5 Basic types are available each with several variations for special purposes.

it is the polarized relay, with the UNIQUE combination of superlative characteristics, that has solved, and is continuing to solve many problems in . . .

High speed switching · Control

Amplification · Impulse repetition

for : Industrial recording
Aircraft control and navigational equipment
Automatic machine control
Analogue computers
Temperature control
Servo mechanisms
Submarine cable repeaters
Burglar alarm and fire detection equipment
Nuclear operational equipment
Biological research
Theatre lighting "dimmer"
and colour mixing equipment
Teleprinter working
Automatic pilots
Remote control of Radio links
Theatre stage-curtain control
Long distance telephone dialling
V.F. Telegraphy
etc, etc, etc.

Therefore — if your project, whatever it may be, calls for a POLARIZED relay, with high sensitivity, high speed without contact bounce, freedom from positional error, and high reliability in a wide range of temperature variations, *you cannot do better* than use a CARPENTER POLARIZED RELAY.

Write or 'phone for technical data —

TMC TRADE MARK

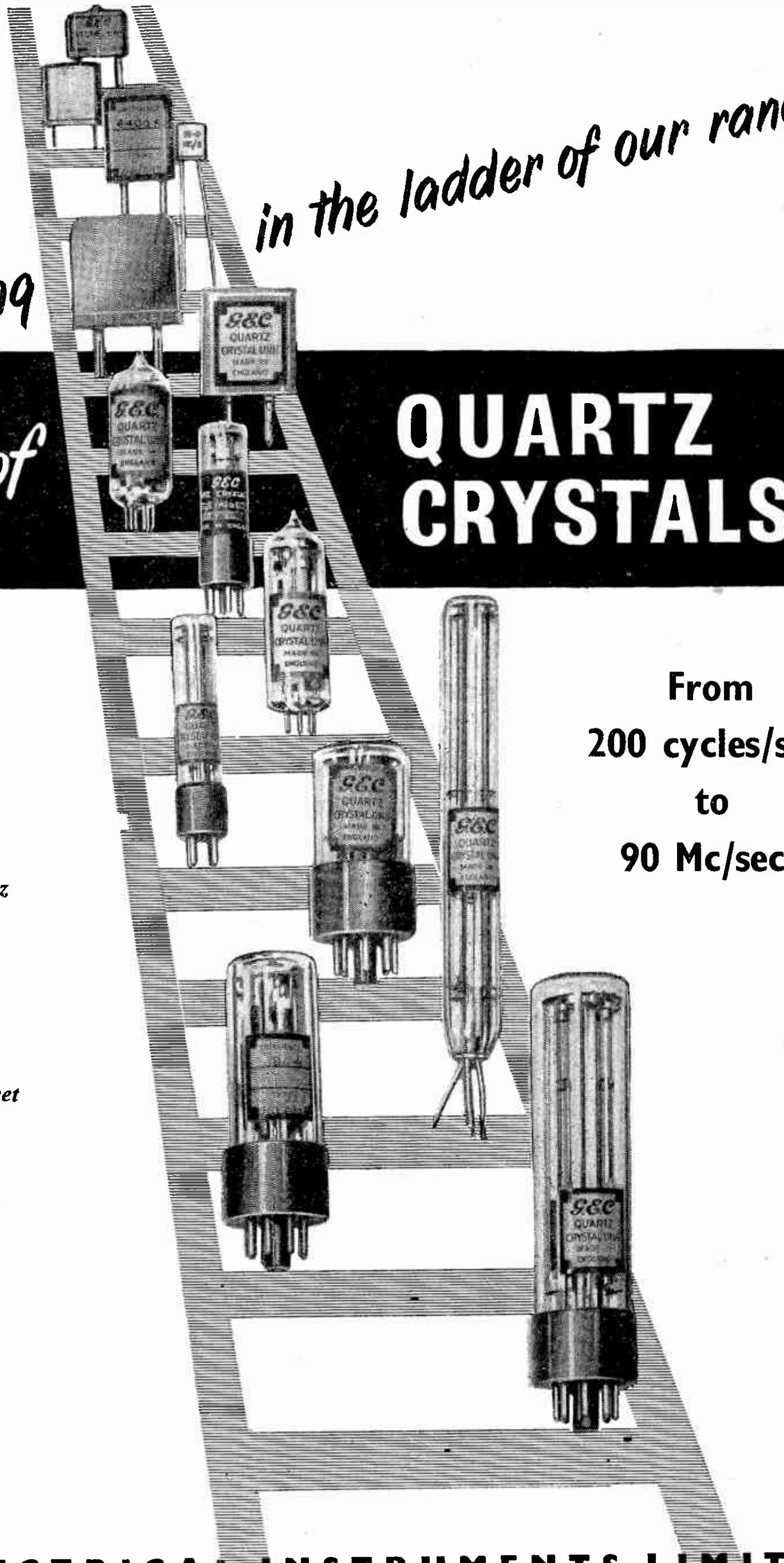
TELEPHONE MANUFACTURING CO. LTD.

DEPT. 407, HOLLINGSWORTH WORKS, DULWICH, LONDON, SE21

TELEPHONE: GIPSY HILL 2211

no rungs missing in the ladder of our range

of **QUARTZ CRYSTALS**



G.E.C.

From 200 cycles/sec. to 90 Mc/sec.

For long term stability and unfailing activity, G.E.C. Quartz Crystal Units provide the basis for reliable communications systems.

A complete range of units to meet D.E.F. 5271 and R.C.L.271 Inter-Services styles can be supplied.

SALFORD ELECTRICAL INSTRUMENTS LIMITED

(COMPONENTS GROUP)

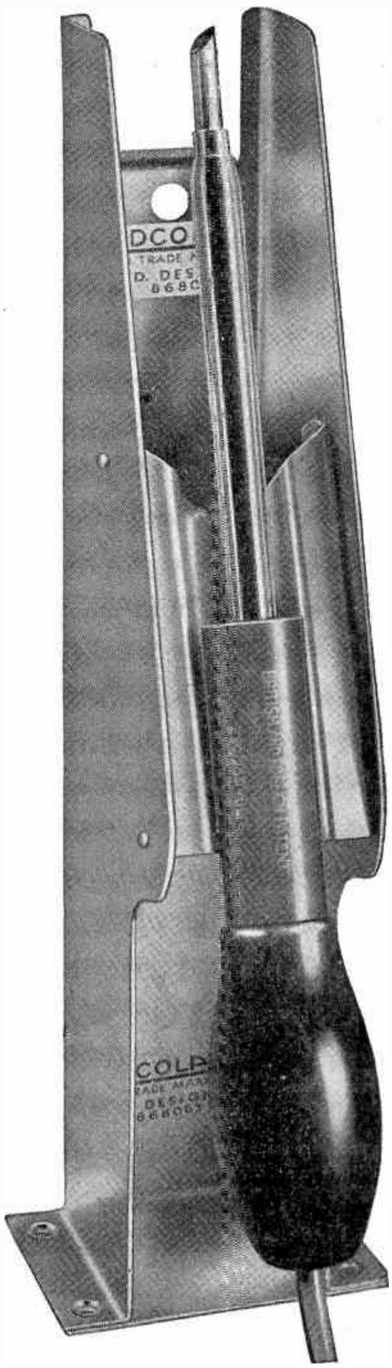
TIMES MILL · HEYWOOD · LANCASHIRE Tel: Heywood 6868

London Sales Office Tel: Temple Bar 4669

A SUBSIDIARY OF THE GENERAL ELECTRIC CO. LTD. OF ENGLAND

ADCOLA
PRODUCTS LIMITED
(Regd. Trade Mark)

Soldering Instruments and Equipment



Comprehensive Range of Models

P.V.C. Cable Strippers

Solder Dipping Pots

Supplied in ALL VOLT RANGES

A PRODUCT FOR PRODUCTION

RADIO, TV
RADAR
ELECTRONICS
TELECOMMUNICATIONS
ETC.

(Illustrated)
Protective Shield
List No. 68

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

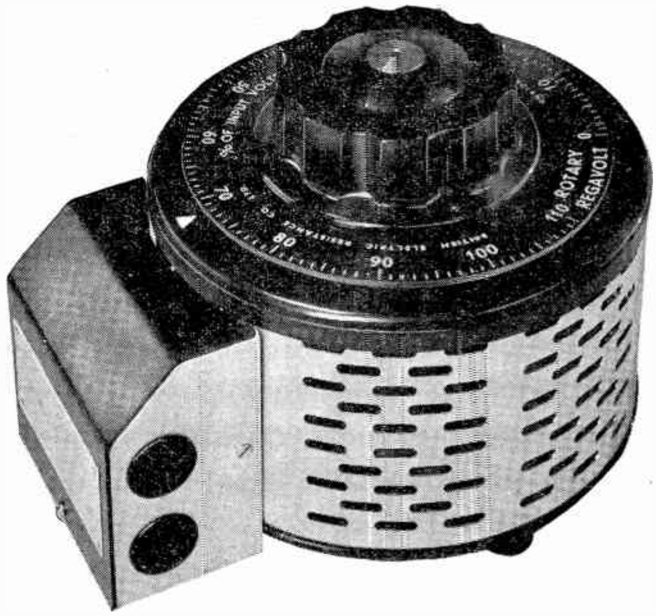
Traditional British Quality and Workmanship

ADCOLA PRODUCTS LIMITED

Head Office, Sales and Service:
GAUDEN ROAD, CLAPHAM HIGH STREET
LONDON, S.W.4 Telephones MACAULAY 3101 & 4272

One Rotary Regavolt* controls any A.C. load from zero to 270v

Designers no longer need special designs (or sets of adjustable resistances) to hold voltages steady. They can rely on Berco Rotary Regavolts for precise voltage regulation, especially where heavy currents are involved. That's why Regavolts are specified for industrial control . . . and for fine control on precision instruments too. Study the features below, and you'll agree the Rotary Regavolt **proves** *Berco know* **how** *!*



Ask for Lists 615A and 615B with full details of Regavolt range. Delivery from stock or in a matter of days.

1. Precision windings ensure even spacing of turns and smooth adjustment of voltage.
2. Windings locked in epoxy resin for permanence, turns can't ride up.
3. Careful design of brush and carrier prevents breakage.
4. Each model mounts F.O.B. or B.O.B.
5. Special self-aligning brush gear on larger models ensures constant contact resistance, constant losses, minimum heating.
6. Six standard sizes cover 250 VA to 2.5 kVA.
7. Universal fixing centres.
8. Multi-ganged assemblies for parallel, series connections, phase changing and 2 and 3 phase working.
9. Can be fitted with motor drive.

Berco have specialised in voltage and current control for 30 years. . .

BERCO KNOW HOW

See us at the Shows

- Electrical Engineers Exhibition (A.S.E.E. Ltd.) March 25th/29th Stand No. **H.22**
- R.E.C.M.F. Exhibition April 14th/17th Stand No. **75**
- I.E.A. Exhibition April 16th/25th Stand No. **943**

THE BRITISH ELECTRIC RESISTANCE CO., LTD.,
in association with
THE BRITISH POWER TRANSFORMER CO., LTD.,
QUEENSWAY · ENFIELD MIDDLESEX
Telephone: HOWard 2411 · *Telegrams:* Vitrohm, Enfield

BR 1431

A*

VISIT US AT THE

I E A

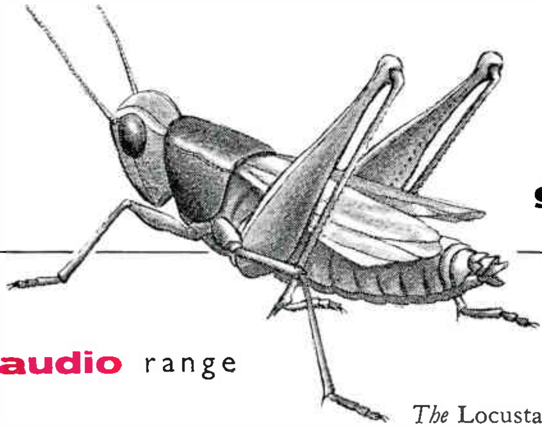
EXHIBITION

OLYMPIA

April 16th, to 25th

STAND NO.

608



if it's signal generation

in the **audio** range

The Locusta Viridissima *(Grasshoppers to you!) produce notes ranging from 6K into inaudibility. Signals with infinite acceleration, output is virtually square wave. As to dynes/cm. — it all depends — male or female — mating season or not!*

Advance

— to be sure!

TYPE H1
SIGNAL GENERATOR

Frequency Range
15 c/s to 50 kc/s

Output
sine wave
200 μV to 20 V
square wave
400 μV to 40 V

List price **£32** in U.K.

Leaflet R41

TYPE J1
SIGNAL GENERATOR

Frequency Range
15 c/s to 50 kc/s

Output
0·1mW to 1 W
(0·25V-25V)

Output Impedance
600 or 5 ohms

List price **£40** in U.K.

Leaflet R33

TYPE J2
SIGNAL GENERATOR

Frequency Range
15 c/s to 50 kc/s

Output
0·1mW to 1 W
(0·25V-25V)

Output Impedance
600 or 5 ohms

with output voltage meter

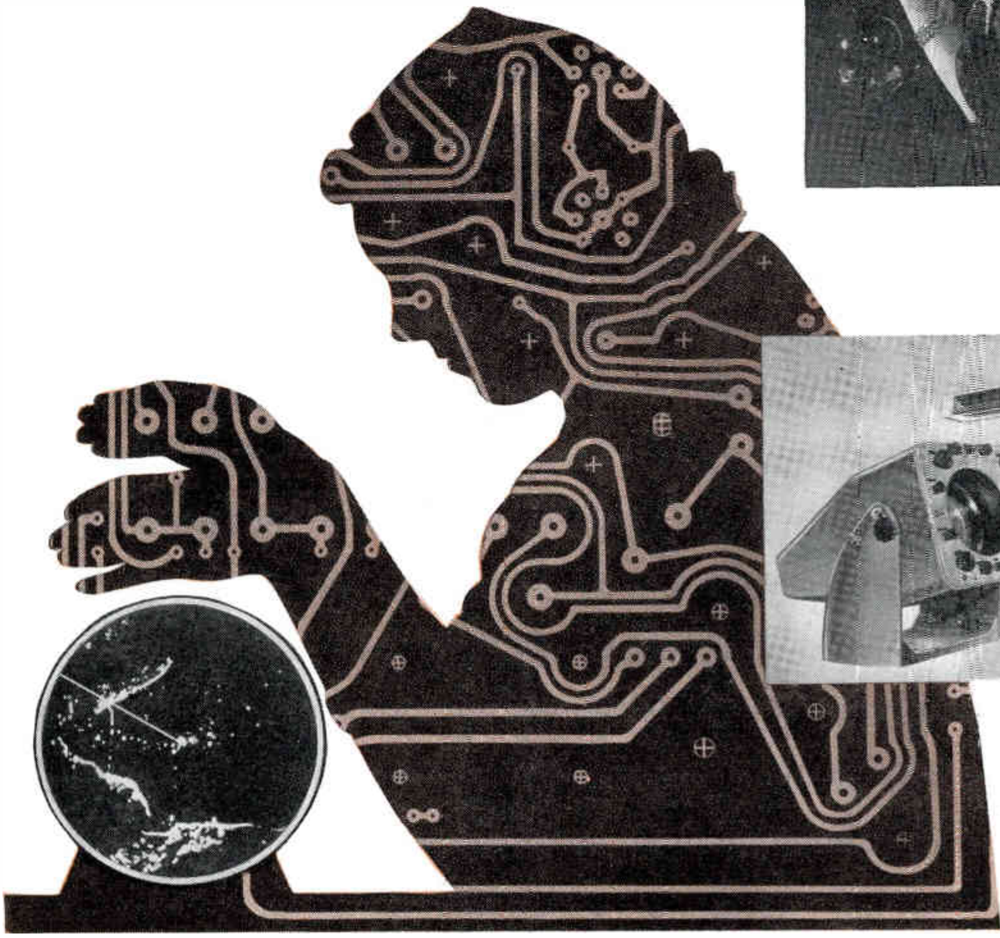
List price **£50** in U.K.

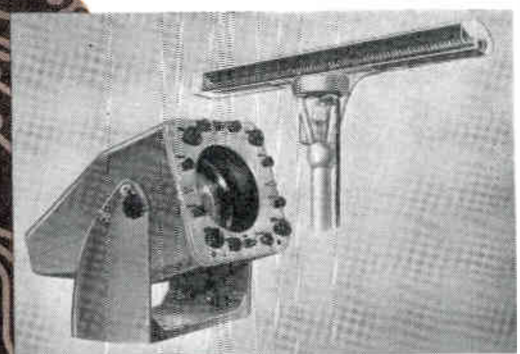
Leaflet R33

Advance COMPONENTS LIMITED · ROEBUCK ROAD · HAINAULT · ILFORD · ESSEX · TELEPHONE : HAINAULT 4444

G D.50

RADIO FORESIGHT–





See us at the Electrical Engineers Exhibition Stand No. K26

with BAKELITE Copper-Clad Laminated

TRADE MARK

The problem of fitting high performance Radar into a compact and lightweight unit suitable for almost any ship has been successfully solved in the Kelvin Hughes type 14 Marine Radar. An important contribution is the extensive use in the Power and Display units of printed circuits produced from BAKELITE Copper-Clad Laminated Materials.

Today printed circuits on BAKELITE Copper-Clad Laminated Materials, either rigid or flexible, are finding new applications throughout the Radio and Electronics industries. They allow more freedom and precision to the designer, reduce production time and costs for the manufacturer, and give the customer a lighter, more compact and reliable instrument.

Write today for a copy of "Copper-Clad BAKELITE Laminated for Printed Circuits".

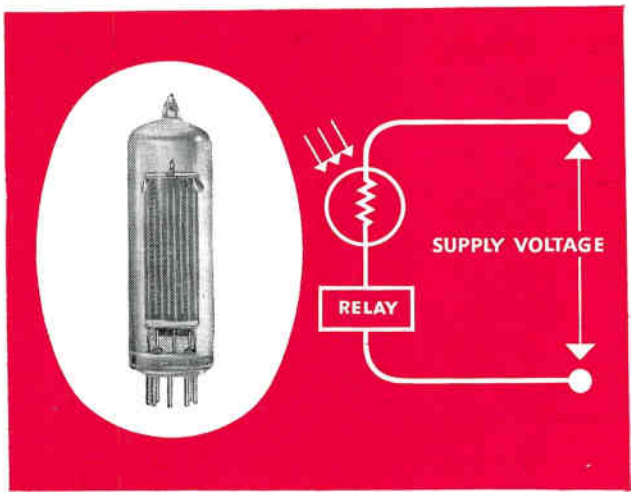
BAKELITE LIMITED

REGD.

12-18 GROSVENOR GARDENS LONDON SW1 · SLOane 0898

Bakelite Limited manufacture an extensive range of plastics materials and maintain a technical service unequalled in the industry. No matter what your plastics problems, this service is at your disposal. SLOane 0898 is the telephone number.

20,000 times more sensitive than an ordinary photocell



New cadmium sulphide technique revolutionizes photocell performance

A new type of photocell with an entirely new order of performance has been developed in the Mullard Solid State Physics Laboratories and is now about to go into production.

Employing a large area cadmium sulphide photosensitive element, the new cell provides a sensitivity approaching that of a photomultiplier . . . *and* a current handling capacity of tens of milliamps.

Technical information is in the course of preparation and you are invited to write to the address below.

Robust relays operated without amplifiers

At the extremely low illumination of 5 ft. candles, robust relays can be operated direct from 40 volt supplies.

Low dark current

In typical applications the ratio of light to dark current is in general more than 10,000:1.

Wide spectral response

The usable response extends through the entire visible spectrum to the near infra-red and peaks in the yellow/red region.

A.C. or D.C. operation

The cadmium sulphide type of photocell is non-polar and can be used with either a.c. or d.c. supplies.

Sturdy construction

Like many other solid state devices, the new type of cell is inherently rugged and suited for industrial use.

Wide range of applications

Flame failure detection, door opening, street lighting control, conveyor control, illumination control and smoke monitoring are only a few of the many possible applications.

Mullard

COMMUNICATIONS AND INDUSTRIAL VALVE DEPARTMENT

MULLARD LIMITED · MULLARD HOUSE
TORRINGTON PLACE · LONDON W.C.I · LANgham 6633

MVT 344

B T-H

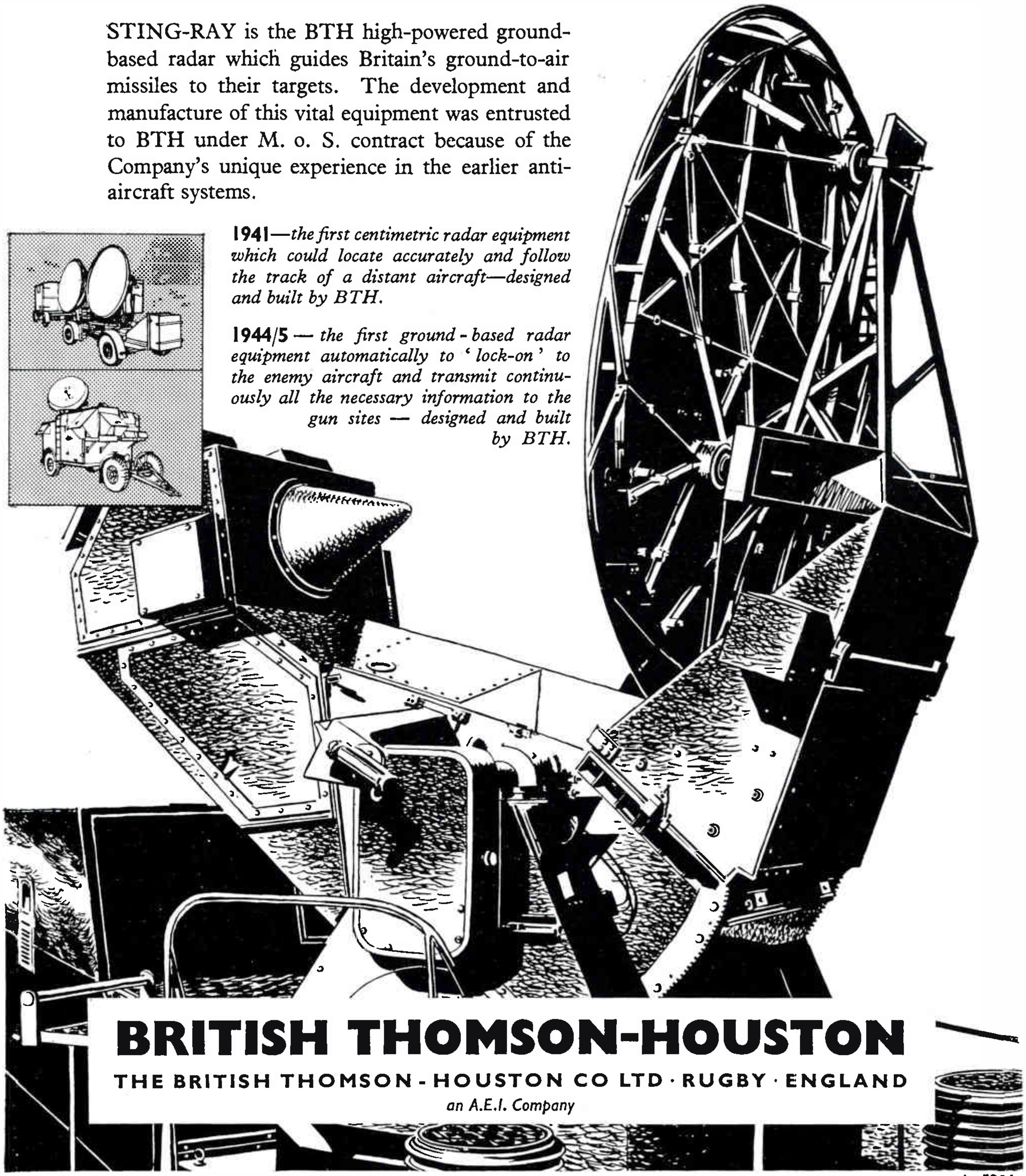
STING-RAY

AN INTEGRAL PART OF
BRITAIN'S GROUND-TO-AIR
GUIDED WEAPON SYSTEMS

STING-RAY is the BTH high-powered ground-based radar which guides Britain's ground-to-air missiles to their targets. The development and manufacture of this vital equipment was entrusted to BTH under M. o. S. contract because of the Company's unique experience in the earlier anti-aircraft systems.

1941—*the first centimetric radar equipment which could locate accurately and follow the track of a distant aircraft—designed and built by BTH.*

1944/5 — *the first ground - based radar equipment automatically to 'lock-on' to the enemy aircraft and transmit continuously all the necessary information to the gun sites — designed and built by BTH.*



BRITISH THOMSON-HOUSTON

THE BRITISH THOMSON - HOUSTON CO LTD · RUGBY · ENGLAND

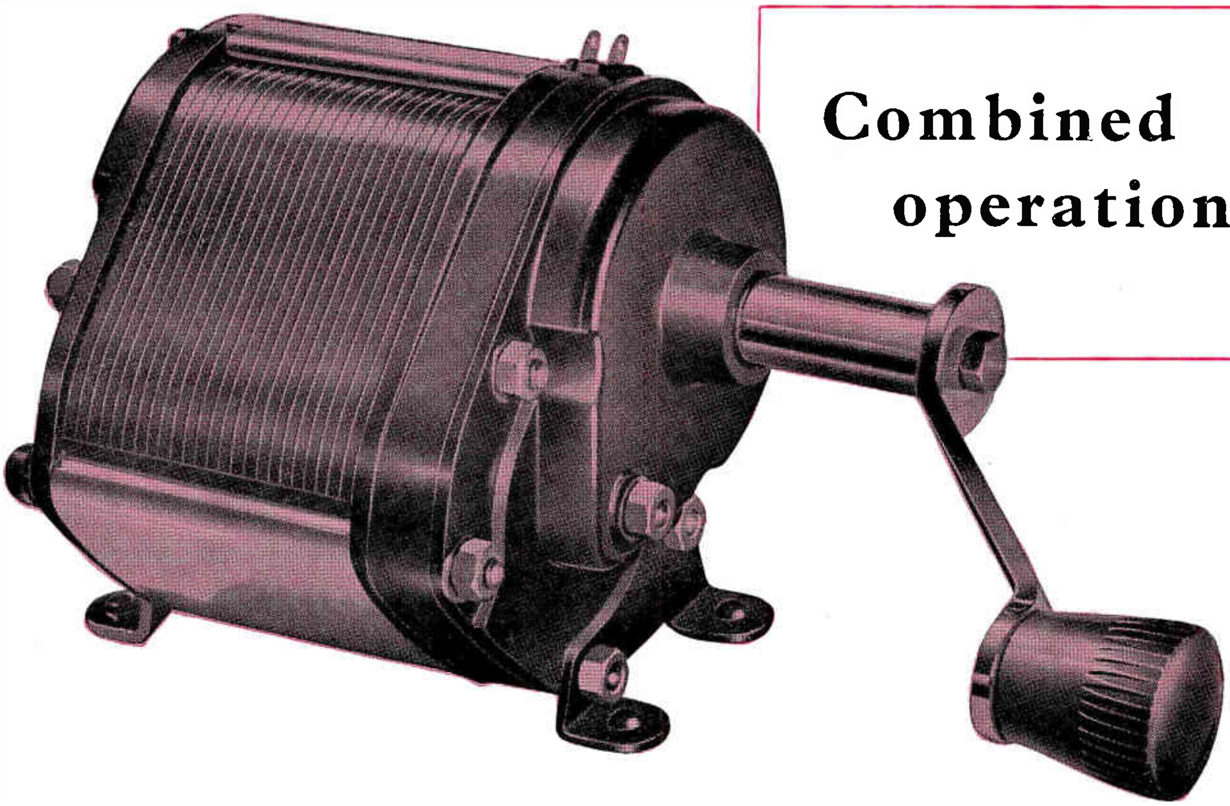
an A.E.I. Company

A. 5206

In a single application, Araldite epoxy resins combine such diverse functions as bonding, impregnating, insulating and providing surface finishes of remarkable protective value.

This rotating magnet generator, made by Ericsson Telephones Limited for telecommunication equipment, has a greater output than other generators of the same drive torque. It incorporates heavy-gauge iron sheet laminations bolted between diecast alloy end cheeks, and the coil is separately wound on to a moulded bobbin which, when assembled, forms an integral part of the lamination structure. Dimensional accuracy between the end cheeks must be combined with a relatively large dimensional tolerance on the thickness of the laminations, and Araldite Surface Coating Resins have proved indispensable for impregnating the coil, locking the laminations, and imparting an excellent surface finish to equipment which also conforms to tropical specifications.

Combined operations



Araldite epoxy resins are used

- ★ **for bonding metals, porcelain, glass, etc.**
- ★ **for casting high grade solid electrical 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 metal, wood and ceramic surfaces**

Araldite *epoxy resins*

Araldite is a registered trade name

Aero Research Limited

A Ciba Company. Duxford, Cambridge. Telephone: Sawston 2121

AP 321

Improved cable clamps on the new range of our Miniature Hermetically Sealed Co-axial Plugs and Sockets will withstand a pull of up to 35 lbs., and are suitable for use with Uniradio 70 cable.

They have been developed and Type Approved to Inter-Service Requirement under R.C.S. 322 and are freely interchangeable with the types which they supersede.

The new types are hermetically sealed and 100% production tested. Sealing caps are available for uncoupled units.

MAGNETIC DEVICES LIMITED

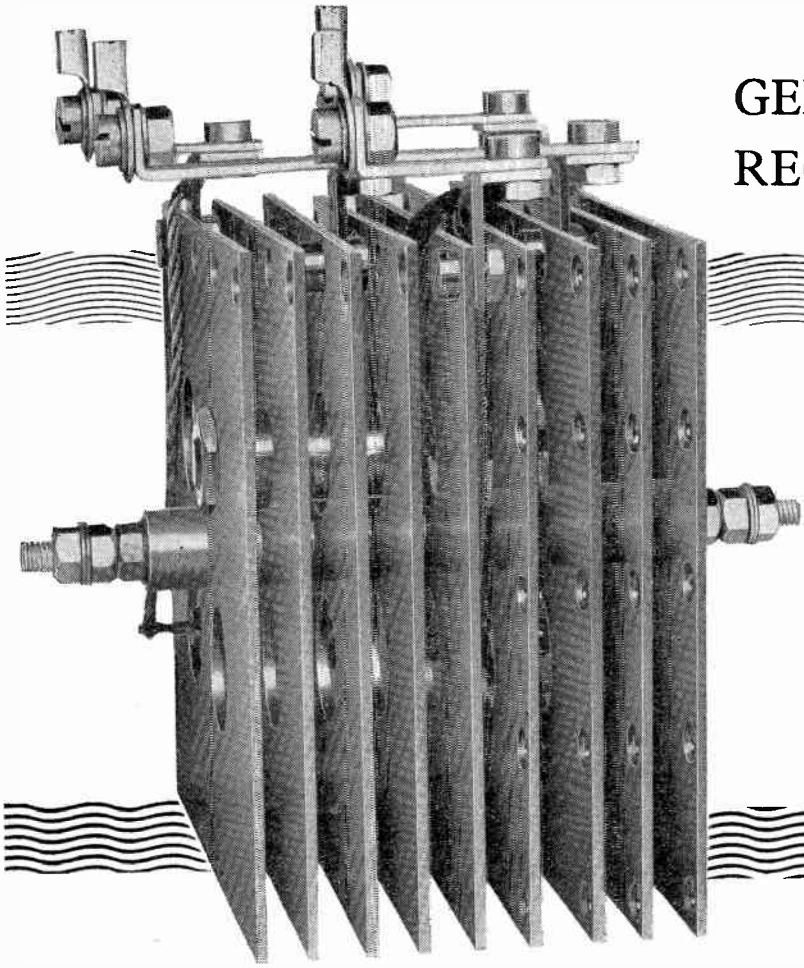
MAGNETIC DEVICES LIMITED

A. I. D. & A. R. B.—Approved

Exning Road, Newmarket, Suffolk

Phone: Newmarket 3181/2/3. Grams: Magnetic Newmarket.

MD 23



GERMANIUM RECTIFIERS

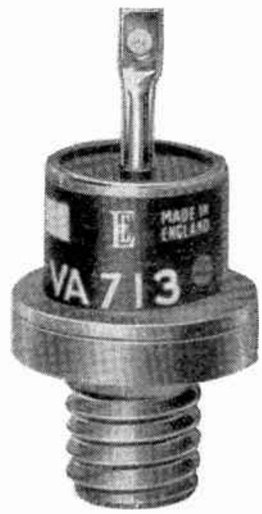
Due largely to advanced crystal perfection techniques and to the generous use of large area junctions, The English Electric Valve Co. Ltd. are able to offer Germanium Power Rectifiers and Stacks with high ratings at 55°C ambient temperature. Standard Stacks cater for all normal single and three-phase circuits.

Stacks are also designed on request to meet individual requirements.

Just a few of the possibilities of one E.E.V. Germanium Rectifier, the VA713, are listed in the table below.

Some Typical Stacked Assemblies

Stack No.	Circuit	R.M.S. Input Voltage per leg (Volts)	D.C. Output (Volts)	D.C. Output Current (Amperes) 55°C.	D.C. Output Current (Amperes) 35°C.	No. of VA713 diodes	Approx. overall dimensions (inches)
13E1	Single phase, half wave	56	25	4·25	13	1	$4 \times 4\frac{3}{8} \times 2\frac{1}{2}$
13F1		43	19	4·25	13	1	
13G1		28	12	4·25	13	1	
13E2	Single phase, full wave	28	25	8·5	26	2	$4 \times 4\frac{3}{8} \times 3\frac{3}{8}$
13F2		21	19	8·5	26	2	
13G2		14	12	8·5	26	2	
13E3	Single phase, full wave bridge	56	50	8·5	26	4	$4 \times 4\frac{3}{8} \times 4\frac{3}{8}$
13F3		43	38	8·5	26	4	
13G3		28	24	8·5	26	4	
13E4	Three phase, half wave	33	38	13	39	3	$4 \times 4\frac{3}{8} \times 4\frac{3}{8}$
13F4		24	29	13	39	3	
13G4		16	19	13	39	3	
13E5	Three phase, full wave	33	75	13	39	6	$4 \times 4\frac{5}{8} \times 6\frac{1}{4}$
13F5		24	56	13	39	6	
13G5		16	38	13	39	6	
13E6	Six phase, half wave	28	38	26	78	6	$4 \times 4\frac{5}{8} \times 7\frac{1}{4}$
13F6		21	29	26	78	6	
13G6		14	19	26	78	6	



'ENGLISH ELECTRIC'

ENGLISH ELECTRIC VALVE CO. LTD.

Chelmsford, England
Telephone: Chelmsford 3491

AP/99

NEW...
COMPACT...

series 320 relay

This relay is of compact design and extremely small for the duty it performs, the contact arrangement is three pole double throw and connections to contact and coil are conveniently brought out at one end of the relay.

Contacts and terminations are housed in a high grade Bakelite moulding and the design of the moulding gives increased creepage path and excellent insulation. Guards are fitted to avoid flashover between contacts.

Maximum Working Voltage: 440 volts A.C. 50 cycles.
V.A. Rating: 5 V.A.
Contacts rated up to 10 amperes at 250 volts A.C. or 30 volts D.C.

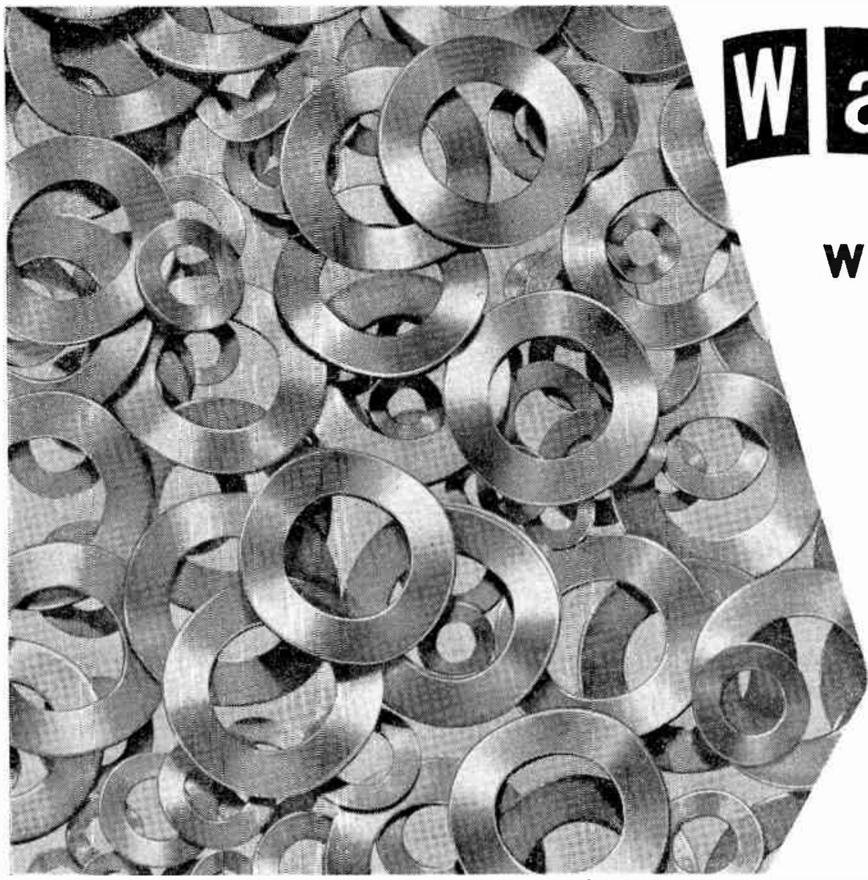
Series 325 *Relay*. The only difference is that this is a D.C. Relay and the maximum working voltage is 250 volts D.C.
Wattage: 2 watts.
Contacts rated up to 10 amperes at 250 volts A.C. or 30 volts D.C.

MAGNETIC DEVICES LIMITED

Magnetic Devices LTD.

A.I.D. AND A.R.B. APPROVED

MAGNETIC DEVICES LIMITED, EXNING ROAD, NEWMARKET, SUFFOLK
Telephone: Newmarket 3181/2/3 Grams: Magnetic Newmarket



Wavey

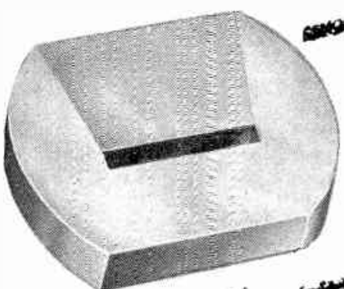
WASHERS

Beryllium Copper "Wavey" Washers. For Electronic Components. Range 8BA to $\frac{1}{2}$. Corrosion resisting, non-damaging to components, higher resistance of "set," higher recovery of load, permanent spring action. They are fabricated from Beryllium copper, heat treated to VPN.350 minimum, Electro Plated to DTD.924, Cadmium Plated to DTD.904 and Passivated to DTD.923. Samples and price lists on request.

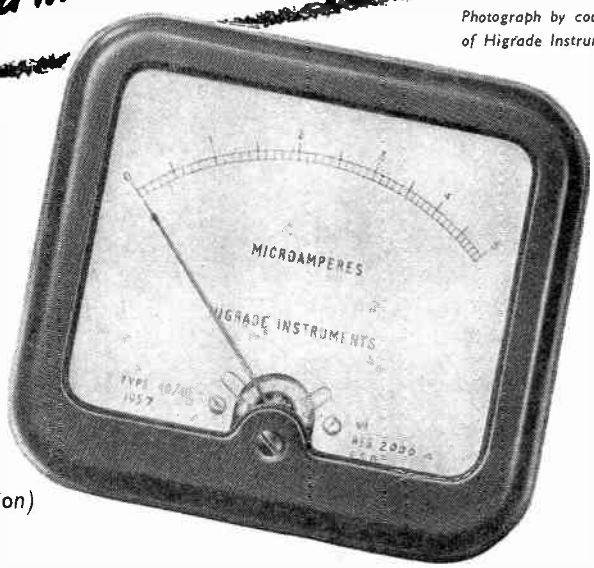
LEAVE IT TO Lewis OF REDDITCH
SPRINGS, SPRING CLIPS
PRESSWORK, WIRE FORMS
VOLUTE SPRINGS

THE LEWIS SPRING CO. LTD. Resilient Works, Redditch Tel. Redditch 720 PBX
London Office: 122 High Holborn, W.C.1 · Tel. Holborn 7470 & 7479

MUREX 'SINCOMAX' MAGNETS are used in 'HIGRADE' INSTRUMENTS



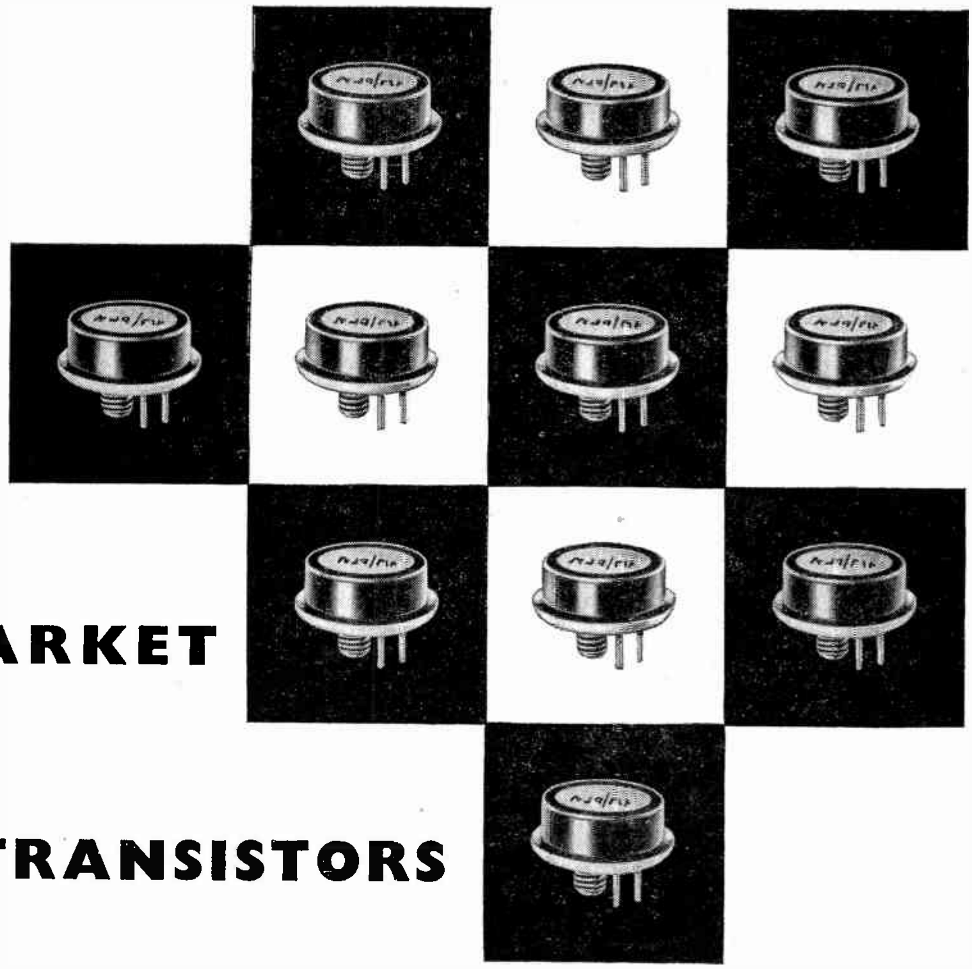
Actual size of Magnet



Photograph by courtesy of Higrade Instruments

The instrument illustrated is 0/5 µA. F.S.D. with terminal resistance of 2,000 Ω. Another example of the use of Murex "Sincomax" magnets where a high energy product with high magnetic stability is essential.
Murex "Sincomax" magnets, with an alloyed bond between magnet and soft iron, continue to give accurate and reliable service in this and many other applications.

MUREX LTD. *(Powder Metallurgy Division)*
RAINHAM · ESSEX · *Rainham, Essex 3322*
Telex 28632 Telegrams: Murex, Rainham-Dagenham Telex
London Sales Office: CENTRAL HOUSE, UPPER WOBURN PLACE, W.C.1. *Euston 8265*



ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

HUGH S. POCOCK M.I.E.E. *Managing Editor*
W. T. COCKING M.I.E.E. *Editor*
Professor G. W. O. HOWE D.Sc., LL.D., M.I.E.E., Hon. M. Brit. I.R.E. *Technical Consultant*

CONTENTS

VOLUME 35 NUMBER 3 MARCH 1958

Three Steps to Victory **79** Editorial
Temperature Measurement with Thermistors **80** *by J. C. Anderson, M.Sc.*
Combined Limiter and Discriminator **85** *by J. W. Head, M.A., and C. G. Mayo, M.A., B.Sc.*
The Fringe of the Field **89** *by Quantum*
Amplifier Low-Frequency Compensation **92** *by J. E. Flood, Ph.D., and J. E. Halder, B.Sc.*
Mathematical Tools **100** *by Computer*
Transmission-Line Low-Pass Filters **103** *by F. Charman, B.E.M.*
ZETA **112**
Correspondence **113**
New Books **115**
Standard-Frequency Transmissions **116**
New Products **117**
Abstracts and References **A33–A48**

Established 1923
Published on the fifth of each month

ANNUAL SUBSCRIPTION
(including Annual Index to Abstracts and References)
Home and Overseas £2 9s 0d Canada and U.S.A. $7.50
Second-class mail privileges authorized at New York, N.Y.

Published by

ILIFFE & SONS LTD. DORSET HOUSE STAMFORD STREET LONDON S.E.1
Telephone · Waterloo 3333 Telegrams · Wirenger, Sedist, London

Branch Offices

COVENTRY *8-10 Corporation Street—Telephone · Coventry 5210* : BIRMINGHAM · *King Edward House, New Street, 2—Telephone · Midland 7191*
MANCHESTER · *260 Deansgate, 3—Telephone · Blackfriars 4412 and Deansgate 3595* : GLASGOW · *26b Renfield Street, C.2—Telephone · Central 1265*

VOLUME 35 NUMBER 3

MARCH 1958

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Three Steps to Victory

IT is generally recognized that radar permitted us to win the Battle of Britain by enabling us to deploy fighters where they were wanted when they were wanted. By thus changing the course of the war, it can be said that radar led to final victory. It is also generally known that Britain had a major lead over all other countries in radar.

In his book "Three Steps to Victory" (Odhams Press, Ltd., price 30s.), Sir Robert Watson-Watt makes it clear that our lead was not purely a technical one. Most countries had been developing radar apparatus before the war. Our lead was brought about by a combination of things. First, there was the decision well before the war to construct Chain stations for early warning in spite of the fact that, in their initial concept, they would provide only range measurement. Secondly, there was the fact that these stations were built and in operation by the outbreak of the war. Thirdly, an operational method of utilizing the information provided by the stations was worked out and operators were trained.

Other countries awaited a higher standard of technical perfection before bringing radar into operational use and, apparently, did not so soon solve the problems of making proper use of the information which the apparatus provided. The result was that we had useful radar before anyone else, and were able to maintain a lead of perhaps two years over the enemy.

Sir Robert claims that this was all due to his cult of the third best—"Give them the third best to go on with; the second best comes too late, the best never comes". This slogan has obviously much truth in it, for it is better to have something that works when needed, even if imperfectly, than nothing at all.

"Three Steps to Victory" is an account, partly autobiographical, of the development of radar and of the part which Sir Robert played in that development and in its first conception. It is a very long book (480 pp.) and it is difficult reading; the author does not tell a straight story but jumps about in time and place and is very repetitive. This is a pity, for it contains much information and there are some lessons to be learnt from it about the handling of new projects. The author seems to have applied his slogan to the writing but, in this case, should we really have had to wait too long for the second best?

Temperature Measurement With Thermistors

SOME PRACTICAL INSTRUMENTS

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

SUMMARY. *Characteristics of thermistors are discussed, and three temperature-measuring devices described:—an industrial thermometer for* 0–100° *C using a Wheatstone bridge, a medical thermometer covering* 85–105° *F incorporating a balanced transistor amplifier and a high-sensitivity device using a two-stage transistor amplifier. Some observations on thermistor stability are included.*

The thermistor is a direct result of the intensive work carried out on the class of materials known as semiconductors, and makes use of their extremely high temperature coefficients of resistance. The actual material used as an element varies with manufacturers; pure germanium has been employed in Germany while, on the British and American markets, the commonest type is a mixture of suitable metallic oxides, such as those of copper, manganese, cobalt and iron. The advantage of the latter type is the higher melting point of the material allowing a higher maximum operating temperature.

The common characteristic of all semiconductors is a large and negative temperature coefficient of resistance governed by the intrinsic conductivity of the material, which arises from the generation of hole-electron pairs by thermal agitation. This is related to temperature by an exponential law which is usually quoted in the form

$$R_T = R_0 \exp{(b/T)} \qquad \ldots \ldots \ldots \ldots \qquad (1)$$

where R_T = resistance at temperature $T°$ K.
R_0 = resistance at absolute zero.
b = a constant for a given thermistor.

The temperature coefficient of resistance is given, by definition, by

$$\alpha = \frac{1}{R}\,\frac{dR}{dT}$$

whence, from (1)

$$\alpha = b/T^2 \qquad \ldots \ldots \ldots \ldots \qquad (2)$$

From this it is seen that α decreases with increasing temperature.

This law will be modified in practice when a current is passed through the thermistor, due to internally-generated heat. The effects of such heating are readily seen by plotting a graph of current against voltage for a thermistor as in Fig. 1. It will be seen that there is a maximum possible voltage for any given thermistor. The effect of this on temperature measurements is discussed later.

Types of Thermistor

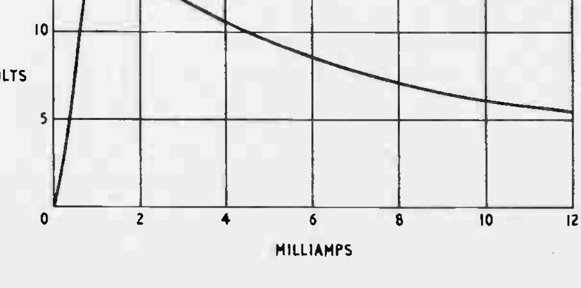
There are two main types of thermistor currently available. The first of these is a ceramic type, in which the appropriate mixture of oxides is sintered with alumina into the desired shape. These are generally rods or discs, having electrical connections made by sputtering on appropriate electrodes and then soldering connecting wires to them. These are not the most convenient type for temperature measurement, but can be used when the body whose temperature to be measured has a large thermal capacity. The Mullard Varite types VA 1005 and VA 1013 are shown in Fig. 2, and the Standard Telephones types K and KB are shown in Fig. 3.

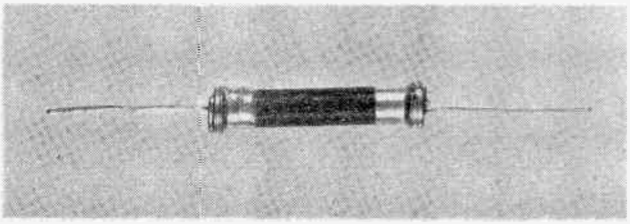
The second type, more generally applicable to temperature measurement, is the bead construction. This is prepared by sintering the oxide mixture into a small bead of about 0·02 in. diameter, which is integrally formed on two parallel platinum wires. The spacing of the wires, chemical constitution and heat treatment of the bead material determine the cold resistance of the bead. It may be mounted in a sealed glass bulb or on a metal disc, or may even be used unmounted in certain applications. The S.T. & C. types M and F are shown in Fig. 4, and Mullard types 2000 and 2400 are shown in Fig. 5.

Temperature Measurement

The most obvious circuit in which to apply the thermistor is that of the Wheatstone bridge of which the

Fig. 1. *Voltage–current curve, F-type thermistor*





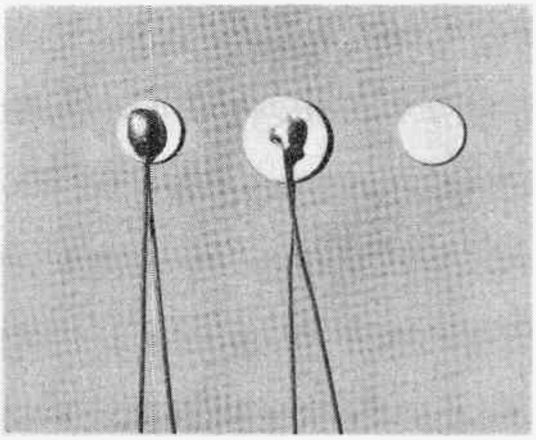


Fig. 2. *Mullard 'Varite' thermistors, types VA 1005 (above) and VA 1013 (below)*

thermistor forms one arm. This, however, is subject to a number of disadvantages which will now be considered.

If a direct-reading instrument is required, the meter will be recording the out-of-balance current of the bridge and, for any given combination of resistors, this current is proportional to the battery voltage. The sensitivity of the instrument will thus be determined by the sensitivity of the meter, and the value of the battery e.m.f. Using a given meter, a high sensitivity will only be achieved with a high battery voltage, and this increases the current in the thermistor, with consequent increase in the heat generated in it. This heat is a source of error since, for a given amount of internally-generated heat, the temperature rise of the thermistor will depend on the thermal conductivity of its surroundings, being different if the element is immersed in, say, air or liquid. The limit to permissible heat dissipation in the thermistor is therefore set by the accuracy required, the temperature rise due to it being necessarily less than the desired limit of accuracy. It will thus be seen that high sensitivity and high accuracy are mutually incompatible in the bridge circuit.

It will also be realised that the sensitivity will fall off as the battery ages. This can be compensated by the use of a battery-standardizing arrangement such as that shown in Fig. 6, or by the use of constant-e.m.f. batteries of the Reubens-Mallory type.

Linearization of Response

Since the temperature coefficient of resistance of the thermistor is inversely proportional to the square of the temperature, the scale of a direct-reading instrument will be markedly non-linear. Some improvement may be effected by shunting the thermistor with a resistor of suitable value. If we denote its resistance by R_s, the modified temperature coefficient of the combination is given by

$$\alpha = \frac{-b}{T^2} \; \frac{1}{1+R_T/R_s} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

where R_T is the resistor of the thermistor at the temperature T. It will be seen that, for instance, when $R_T = R_s$ the temperature coefficient is halved. As the temperature rises the coefficient approaches more nearly to its normal value due to the decrease of R_T. The resulting improvement in linearity is thus seen to be at the expense of sensitivity. Fig. 7 shows the relation between resistance and temperature for an F type (S. T. & C.) thermistor, and the effect of shunting it with a fixed resistor, over the range 0–100°C.

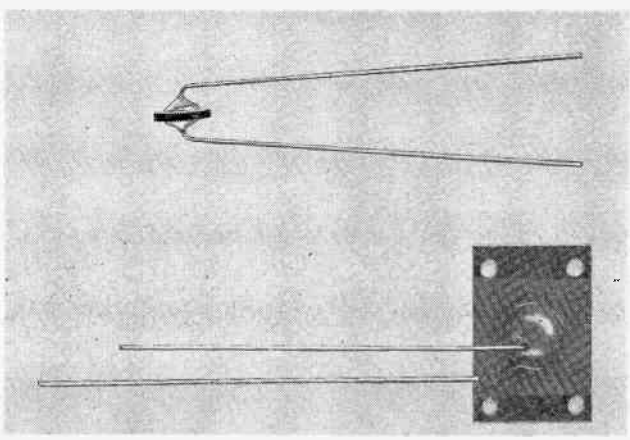
Industrial Wheatstone Bridge Circuit

The circuit for a direct-reading thermometer over the range 0–100°C is shown in Fig. 6. The shunt resistor was chosen to be equal to the resistance of the thermistor at 40°C. The calibration scale of the 500 micro-amp. meter used is shown in Fig. 8. The standing current from the battery was approximately 2 milliamps at 0°C, rising to 3·3 mA at 100°C. The current in the thermistor itself at 0°C was approximately 0·5 mA, giving a dissipation of 1 mW, and this rose to 2·5 mA at 100°C, corresponding to a dissipation of 1·6 mW.

The 'set-zero' control was such that, with new batteries, the full resistance was in circuit, producing an effective bridge voltage of 7·5 V. This allowed for a fall of about 17% in battery e.m.f. before replacement was necessary. The zero was adjusted by switching in a preset resistor, equal in value to that of the thermistor arm of the bridge at 100°C, and adjusting the control for full-scale reading on the meter. The use of full-scale rather than zero for this purpose gives a higher accuracy in setting, as the maximum current drain occurs under this condition.

No error was detectable due to self-heating effects within the limits of accuracy of the meter, reckoned at 0·2° C when calibrated directly in degrees Centigrade. It was, however, found that the thermistor-mounting

Fig. 3. *S.T.C. thermistors, types K (above) and KB (below). In the latter, the block of thermistor material is soldered to a metal plate*



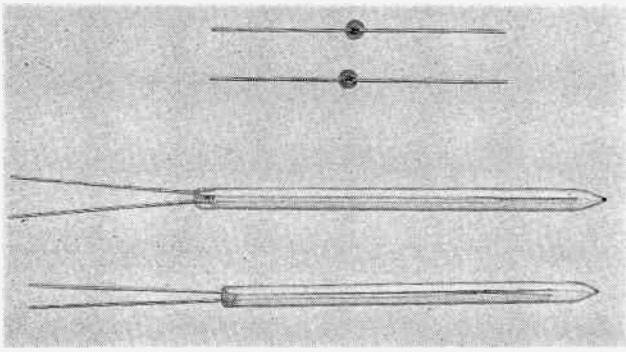


Fig. 4. *S.T.C. bead thermistors, types M (above) and F (below). In type M, the bead is fixed to a flat copper disc*

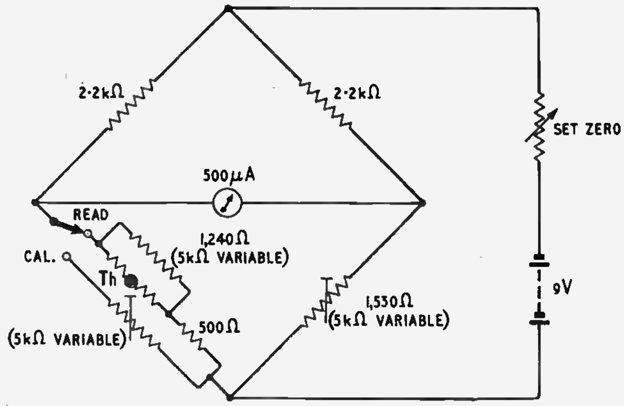


Fig. 6. *Industrial bridge instrument*

tube stored heat which flowed back into the thermistor mounted in the tip. This is illustrated in the cooling curve of Fig. 9, which was plotted after immersion of the thermistor in boiling water for 5 minutes.

Replacement of the thermistor, if broken, required re-adjustment of the pre-set shunt resistor and of the pre-set zero. These adjustments were found to be adequate to cover variations between thermistors, which are supplied at 20% tolerance by the manufacturers.

Transistorized Medical Instrument

To meet the need for a portable self-contained instrument of high accuracy and long life, a transistorized circuit, given in Fig. 10 was developed. This circuit has been patented by the author.

The range covered was 85–105° F with an accuracy of 0·1° F. Mallory cells were used in order to obtain stability of supply voltage, and their life with normal use proved to be in excess of one year. The standing current through the thermistor is of the order of 150 micro-amps, giving a dissipation of approximately 30 micro-watts, which remains virtually constant. This is due to the high current amplification of the transistor circuit, since the variation in current through the thermistor to produce full-scale deflection of the meter is of the order of 10 μA. Errors due to self-heating effects may, therefore, be considered to be negligible.

The circuit is designed to minimize variation of sensitivity due to changes in ambient temperature affecting the transistors. The effect of an ambient temperature change from 32 to 100° F was found to be to introduce a full-scale error of 0·2° F. Over the range

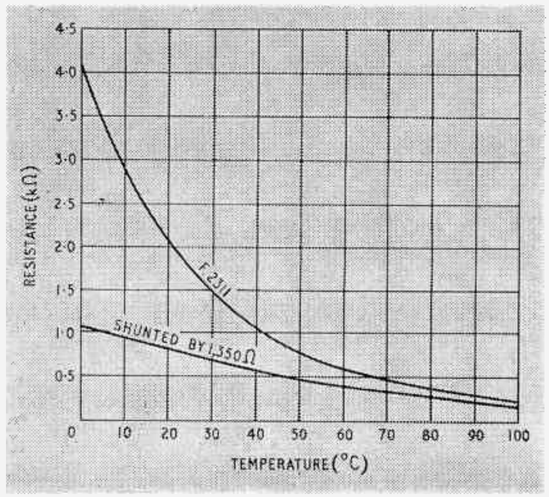


Fig. 7. *Effect of shunting F-type thermistor*

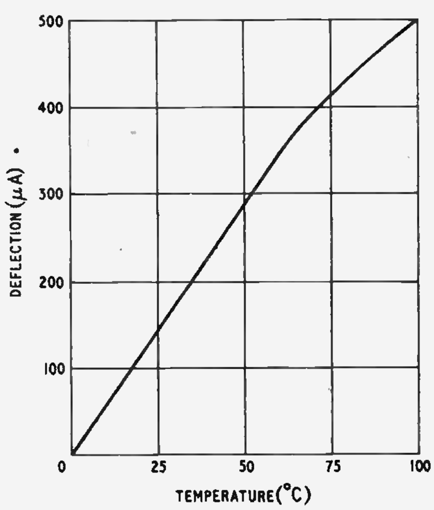
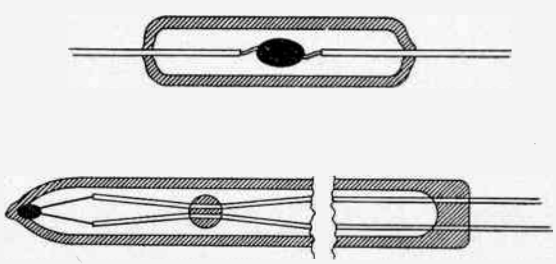


Fig. 8. *Calibration scale of industrial bridge*

Fig. 5. *Mullard bead thermistors, VA 2000 (above) and VA 2400 (below)*



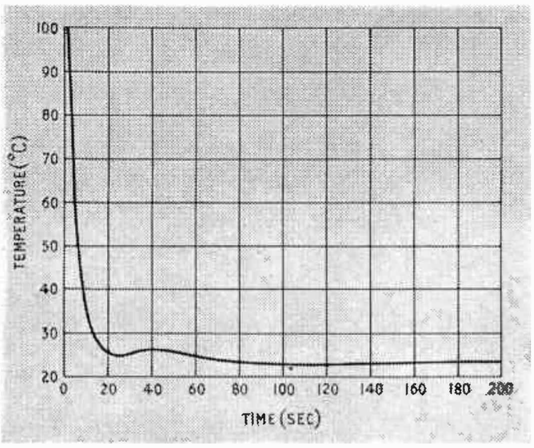


Fig. 9. *Cooling curve for F 2311 after 5 minutes at* 100° C

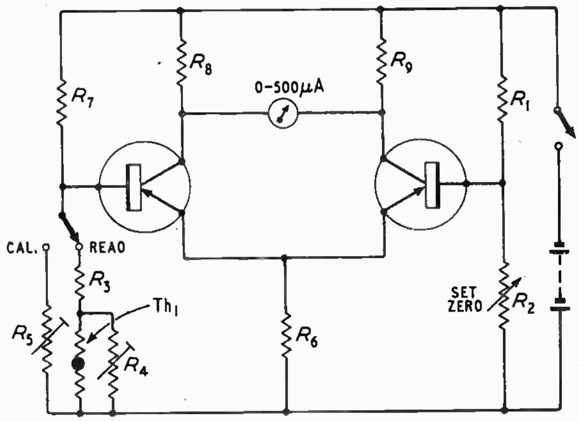


Fig. 10. *Transistorized medical instrument*

of normal room temperatures 55 to 70° F, no detectable error was found.

A differential type of circuit may readily be produced from that of Fig. 10 simply by replacing the resistor R_2 by a thermistor maintained at a standard temperature, while a number of test thermistors may be switched in turn in place of Th_1. To allow the same calibration to apply to each thermistor, it is necessary to match them by means of series and shunt resistors R_3 and R_4 of the appropriate values.

A photograph of a five-inch instrument produced for hospital use as a skin thermometer and in connection with hyperthermia operations (refrigeration of the patient), is shown in Fig. 11.

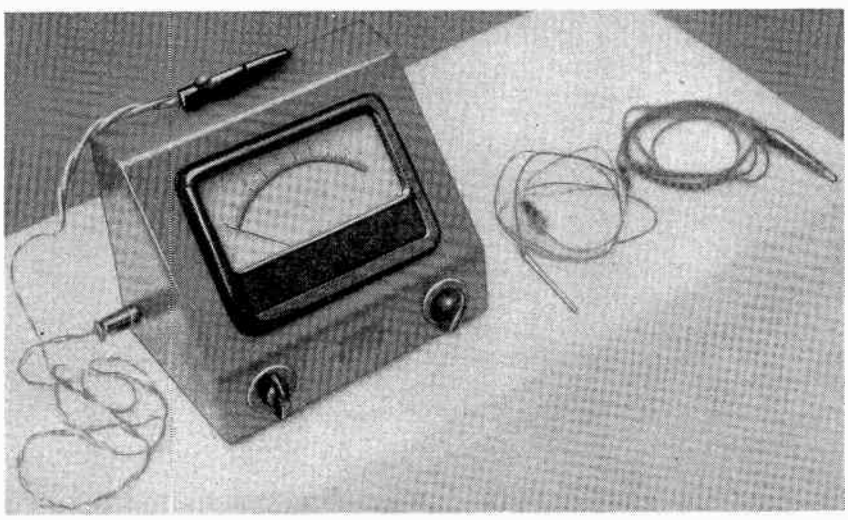
High-Sensitivity Circuit

This instrument made use of the temperature-stabilized transistor circuit due to Shea, shown in Fig. 12. This was an experimental instrument, designed to measure to 0·01° C accurately, with a full-scale deflection corresponding to 1° C. Absolute measurement of temperature was not required, and the set-zero control was simply arranged to set the meter to zero for any given temperature of the thermistor, a rise of up to 1 degree from this temperature then being measurable to ± 0·005 degree.

For purposes of calibration, and estimation of accuracy, an N.P.L.-standardized 2-foot mercury-in-glass thermometer was used to calibrate a thermocouple. The cold junction was maintained at a constant temperature in melting ice and the response was assumed to be sufficiently linear to allow sub-division of the potentiometer scale down to 0·002° C. A standard 100° C was provided by boiling distilled water, correction being made for atmospheric pressure. A double vacuum flask was used to contain water at intermediate temperatures the temperature being monitored by the thermocouple.

The accuracy of the meter used in the high-sensitivity instrument was 1%, and the overall accuracy of the instrument was found to vary between 0·01 degree and 0·04 degree, depending on the condition of the thermistor. This was investigated, and it was found that if the thermistor was maintained at 100° C for a long period (3 hours) and then allowed to return to 20° C (at which the meter had been zeroed) there was no detectable error. If, however, the thermistor was 'cycled' by heating it to 100° C and then allowing it to cool to 20° C at two-minute intervals for a period of 20 minutes, the

Fig. 11. *Five-inch portable instrument showing alternative probes*



resistance at 20° C was found to have changed by 2% —revealed by the necessity of re-setting the zero. More important than this, however, was that the constant b of equation (1) was found to have changed. This was revealed as a change in overall sensitivity of the instrument, and the change in b value was calculated at between 3% and 4%. After allowing the thermistor to rest for a period of three hours, it exhibited a partial recovery, the calculated b value being within 1% of the original.

The general conclusion from these investigations was that the thermistor can be relied upon for temperature measurements down to 0·005° C providing it is not cycled.

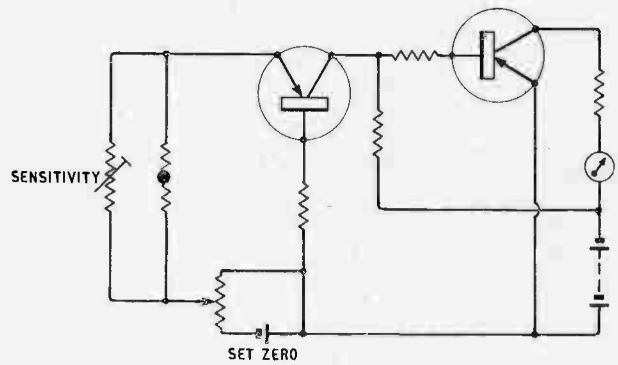


Fig. 12. *High-sensitivity circuit*

Double-Coil Instrument

An alternative method of overcoming the voltage-dependence of the bridge-type circuit is to employ a double-coil instrument movement of the megger type. This is already in use by car manufacturers for dashboard indication of radiator temperature. In British cars now fitting this system the thermistor used is a bead type, immersed in the radiator, connected in series with one of the meter coils and the battery. The other coil is supplied from the battery via a suitable resistance which is chosen initially to give correct temperature indication with the thermistor used. The movement is usually biased to read hot in the event of thermistor failure.

Valve-Type Instruments

Where a permanent installation is required, a valve circuit will generally be the most suitable. The prime requirement of design is that the standing current in the thermistor should be as small as possible, and that it should change as little as possible. The author has used a number of different circuits of conventional design, the most successful of which has been the 'long-tailed pair' with the signal provided from a Wheatstone bridge containing the thermistor. From experience, however, it is strongly felt that, unless extremes of ambient temperature are to be encountered, the transistor circuit is much more economical, effective and satisfactory for this type of work.

Conclusion

The thermistor-thermometer offers tremendous advantages in terms of rapidity of response, and ability to read surface temperatures and the like. It also permits continuous monitoring of temperature from a remote point, the distance not being limited by the resistance of the leads, since this will always be small compared with the resistance of the thermistor. For highly accurate work, however, the usefulness of the thermistor is in some doubt.

In all types of instrument, breakage of the thermistor requires recalibration owing to the inability of the manufacturers to supply elements to a tolerance closer than 20%. It is considered that there is a strong case for a considerable reduction in the tolerance limits in the manufacture of thermistors intended specifically for temperature measurement, even at increased cost. This has already been done in Germany. If this demand can be met, there undoubtedly exists a very wide field of application in both industry and medicine for the thermistor as a temperature-measuring device.

The author wishes to express his appreciation to Messrs. Standard Telephones and Cables' Industrial Supplies Division, and to Messrs. Mullard Ltd., for their frequent and ready assistance in all matters pertaining to the preparation of this article.

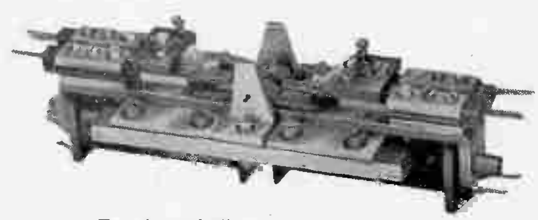
REFERENCE

[1] R. F. Shea, "Transistor Circuits", Van Nostrand, 1955.

A NEW TYPE OF CHOPPER

The photograph shows an experimental vibrator constructed largely from standard relay parts. It was shown at the recent Open Days of the British Scientific Instrument Research Association, being intended for use as the interrupting device in contact-converter or 'chopper' types of d.c. amplifier.

Contrary to normal practice the vibrator is unpolarized: instead, the coils are fed in quadrature (Fig. 1) and the armature is thereby forced to vibrate at twice the energizing frequency. Direct inductive pick-up in the contact loops from the energizing coils—a major consideration in previous designs—is rendered relatively unimportant, for it will differ in frequency from the signal by a factor of 2.



Experimental vibrator

The zero error introduced by the vibrator is well below 1 μV, the ultimate limitation being thermal and voltaic potentials in the contact circuits.

Patent application No. 28646/55 (Full patent pending). D. J. R. Martin and British Scientific Instrument Research Association.

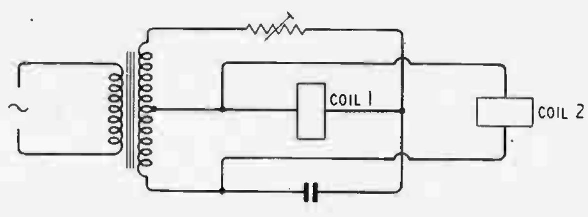


Fig. 1. *Circuit for energizing the vibrator coils. Coil 1 is supplied with current from an RC phase-shift network*

Combined Limiter and Discriminator

IMPROVING F.M. RECEIVER PERFORMANCE

By J. W. Head, M.A. and C. G. Mayo, M.A., B.Sc., M.I.E.E.*

SUMMARY. *The output voltage from a diode limiter is constant in amplitude and is associated with low impedance. The effect of applying such a voltage to a phase discriminator is discussed. A modified discriminator circuit† is described for which performance (which is easily computed) is not adversely affected by the limiter. Distortion less than 0·1% of any harmonic is easily obtained. Stray capacitances are less than for the normal phase discriminator, so that high sensitivity and high output can be obtained by efficient matching of elements. The equations giving the output are similar in form to those of the phase discriminator discussed earlier*[1]. *The circuit may be modified to allow the use of two consecutive limiters, but this should hardly be necessary when an improvement in the limiter circuit gives a.m. suppression of 50 dB for a single diode.*

In a receiver for frequency-modulated signals, we require not only a discriminator but also a limiter; that is, a means of suppressing unwanted amplitude modulation. Now it is much easier to make the voltage amplitude of a signal constant than to make the current amplitude constant. A voltage limiter is an essentially non-linear device; it is characteristic of such devices that they require operating voltages of the order of several volts. Such voltages would exist, in a typical f.m. receiver, on the windings of a Foster-Seeley discriminator, but not at any earlier stage unless extra amplification were specially provided. The effect of applying such a limiter directly to the primary or secondary of a Foster-Seeley discriminator is considered in Section 1 and found unsatisfactory. Alternative arrangements are therefore considered in Section 2. The formulae derived in Section 3 for the voltages associated with the most suitable of these arrangements have the same general form as those for the Foster-Seeley discriminator

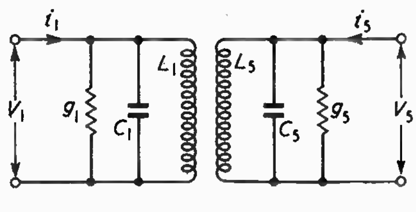


Fig. 1. *Circuit which can be tuned to behave as an impedance inverter*

discussed in an earlier paper[1]. The arrangement discussed in Section 2, however, has the advantages that amplitude modulation can be virtually eliminated, and that the centre-tapped coil has a low voltage and is well separated from the primary, so that stray capacitances are low. In this article, the approach is mainly theoretical. Practical work on these principles is now in progress and it is hoped that in due course an f.m. set embodying them will be described in detail, with performance-test results. We hesitate to give details in advance of actual construction. A model has been made, and preliminary results are in accordance with theory, but no attempt to obtain the best possible values of circuit parameters has so far been made.

1. Effect of Applying a Limiter to the Primary of a Foster-Seeley Discriminator

As in the main text of an earlier article[1], we shall regard the discriminator as adequately represented by the simplified circuit of Fig. 1. When the discriminator is in operation, i_5 is zero, and equations (13) of Reference 1 give, in the notation there used,

$$\left.\begin{aligned} V_1/i_1 &= (\beta + jx)/2s\Delta \\ V_5/i_1 &= j\lambda/2s\Delta \end{aligned}\right\} \quad \ldots \quad (1)$$

where

$$\Delta = (\alpha + jx)(\beta + jx) + \lambda^2 \quad \ldots \quad (2)$$

the impedances $j\omega_0 L_1$ and $j\omega_0 L_5$ of L_1 and L_5 at the mean frequency $\omega_0/2\pi$ are both assumed numerically equal to unity (in normalized units), sx is the fractional frequency deviation, $(\omega - \omega_0)/\omega_0$ at frequency $\omega/2\pi$, x being $+1$ at 100% modulation so that s is the ratio of the maximum frequency deviation to the mean frequency, 0·007 if the maximum frequency deviation is 75 kc/s and the mean frequency 10·7 Mc/s, and the conductances g_1, g_5 and the coupling coefficient k between L_1 and L_5 are respectively given by

$$g_1 = 2s\alpha, \quad g_5 = 2s\beta, \quad k = 2s\lambda \quad \ldots \quad (3)$$

It is assumed that the total capacitances C_1 and C_5 across L_1 and L_5 are chosen so that the resultant primary and secondary admittances are purely resistive at the mean

* B.B.C. Research Department, Kingswood Warren, Tadworth, Surrey.
† Provisional Patent Application No. 15323/57.

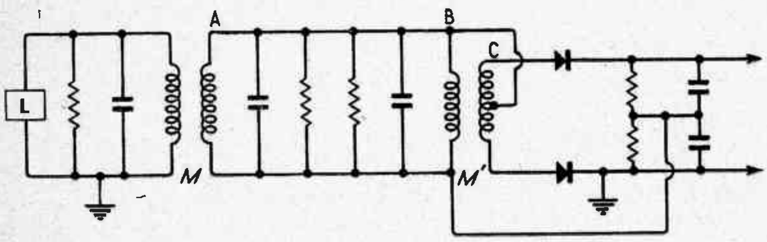


Fig. 2. *A discriminator circuit with limiter, suppied by the circuit of Fig.* 1

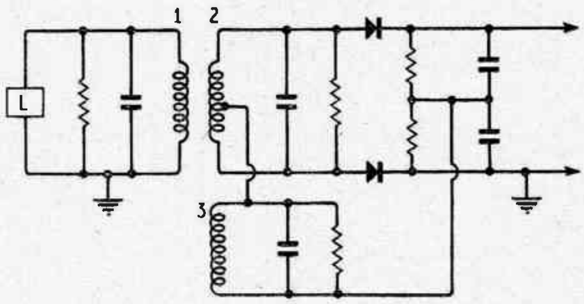


Fig. 3. *A preferred alternative arrangement of the circuit of Fig.* 2

frequency. The action of the diodes (not shown in Fig. 1) makes the output of the discriminator

$$V_0 = |V_1 + V_5| - |V_1 - V_5|^* \qquad \ldots \quad \ldots \quad (4)$$

If we now suppose that $|V_1|$ is made constant by means of a limiter, (1) becomes

$$V_5 = \frac{j\lambda}{\beta + jx} V_1 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

so that V_0 in (4) does not depend linearly on x unless β is so large that sensitivity is seriously reduced. With secondary limitation, so that $|V_5|$ is made constant, we have

$$V_1 = \frac{\beta + jx}{j\lambda} V_5 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

but β is now a variable, adjusted so as to make $|V_5|$ constant. If α, β, λ are constant, equal and sufficiently large, it is easily shown that $|\Delta|$ in (2) varies little with x, but if β varies with x, so will $|\Delta|$. Thus secondary limitation is even more unsatisfactory than primary, and the limitation of any combination of V_1 and V_5 is similarly unsatisfactory.

2. Discussion of Alternative Arrangements

When the circuit of Fig. 1 is tuned as here assumed, a low impedance placed across coil 1 gives a high impedance out from coil 5; this is discussed more fully in Reference 1. A diode limiter has the required low impedance; it causes the voltage of coil 1 to become nearly constant and the current from coil 5 to become nearly constant. The circuit of Fig. 1, with limiter added in this way, thus becomes suitable for supplying a separate discriminator as in Fig. 2. Now in Fig. 2, the coils at A and B are in parallel, so they could be replaced by a single coil. There is also no reason why the centre-tapped coil at C in Fig. 2 should not be exchanged with B, giving the structure of Fig. 3. This has the advantage that the voltage associated with the centre-tapped coil is less than in the Foster-Seeley arrangement under the conditions favourable to low distortion. As the coupling between the primary and the centre-tapped coil is designedly low (to take advantage of the impedance inverter effect), stray capacitances between the primary and the centre-tapped coil should be much lower than with the original Foster-Seeley arrangement. Further developments are also possible. The voltage limiter works best in a high-impedance circuit such as a pentode anode circuit; this can be secured with the arrangement of Fig. 3. The ordinary diode limiter alone does not in general provide sufficient a.m. suppression. One possible way of improving a.m. suppression is to precede the limiter in Fig. 3 by an additional limiter and an impedance inverter, the output of the impedance inverter being fed to the limiter shown in Fig. 3. The whole assembly would then be arranged as in Fig. 4 (a) with circuit diagram as in Fig. 4 (b). The coil marked 1 would be part of the impedance inverter associated with the limiter preceding Fig. 3; coil 2 is associated with the limiter **L** in Fig. 3, and coil 3 is the centre-tapped coil. The coupling between two consecutive coils in Fig. 4 (a) is about 5%. This arrangement is not discussed further because one limiter (of improved design)* can give about 50 dB of a.m. suppression. In this, a rejector circuit is added in series with the limiter diode as shown in Fig. 4 (c). This rejector circuit consists of an inductance and a capacitance in parallel, tuned to one of the harmonics (say, the second) of the mean discriminator frequency $\omega_0/2\pi$. With resistive circuits (e.g., at low frequency), the effect of a diode would be to distort the voltage waveform by clipping it. With tuned circuits, however, the effect is quite different. The tuned circuit maintains the waveform so that the diode acts as a resistive load varying with the input. The voltage waveform may have only a few per cent of harmonic distortion, and the use of two diodes would make little difference. In the construction here described, however, the diode is separated from the tank circuit (tuned to the mean discriminator frequency) by the parallel-tuned harmonic rejector, so that the voltage across the diode circuit is no longer constrained by the low impedance of the tank circuit to harmonics. This voltage may therefore have a clipped waveform although the voltage across the tank circuit remains undistorted. The waveform of the voltage across the diode is roughly indicated in Fig. 4 (d) (i) to (vi), if the rejector circuit is tuned to the second harmonic. In (i) the voltage across the diode is a pure sine-wave, since its maximum value M is less than the diode back-off voltage E, so that the diode never conducts even in the absence of the rejector circuit. In (ii) the waveform is still sinusoidal; the diode would conduct instantaneously in the absence of the rejector circuit; (iii) is the waveform appropriate when $E < M < E\sqrt{2}$; in this case, the diode would conduct in the absence of the rejector circuit, which prevents it from doing so; (iv) is the waveform appropriate when

* Equation (2) of Reference 1 differs from (4) here in having $\rho_1 V_1$ for V_1 and $\rho_5 V_5$ for V_5 to allow for inequalities in the inductances associated with the actual discriminator circuit, here simplified to that of Fig. 1.

* Provisional Patent Application No. 9568/57.

$M = E\sqrt{2}$; the diode is now just conducting for a quarter of a cycle surrounding the point of maximum input voltage; (v) and (vi) are waveforms appropriate for $M > E\sqrt{2}$, when there is appreciable diode current. For a third-harmonic rejector, the limiting voltage at which diode current begins becomes $2E/\sqrt{3}$ and so on. A combined second- and third-harmonic rejector would give a still greater increase, but the efficiency of a single third-harmonic rejector is adequate. Fig. 5 shows an experimentally-derived curve, in which fundamental voltage across the diode is plotted against fundamental current through the diode for such a limiter.

The construction of the rejector circuit is not critical. The effective impedance of the rejector should be high compared with the tuned impedance presented to the anode of the driving valve, but the Q of the circuit need not be high, say 10 or 20. The inductance should be high enough to tune very nearly with stray capacitances.

3. Formulae for the Preferred Arrangement

The circuit finally recommended is then as in Fig. 3, where the limiter L is a special one (having a third-harmonic rejector in series with the diode). L is either placed across the primary, or on a winding bifilar to the primary. The coupling of this primary (coil 1 in Fig. 3) to either half of the centre-tapped coil of the discriminator proper (coil 2 in Fig. 3) we shall take as $2s\lambda$ [where s is a mere abbreviation for 0·007, as in Equ. (1)]. The coupling of either half of the centre-tapped coil to the secondary (coil 3 in Fig. 3) we shall take as $2s\mu$, while the coupling between the primary and secondary is taken as $4s^2\lambda\mu$. This coupling, being of the second order in s, is neglected. Then λ and μ are of order 3; their actual values depend on the relative importance of output and fidelity. It has been shown in Reference 1 that any asymmetries likely to occur are of no practical significance, so that it is legitimate to regard the two halves of the centre-tapped coils as equally coupled to the other coils, and perfectly coupled to each other.

The performance of this arrangement can be calculated as in Reference 1. If we adopt the same normalization as in equations (1) to (3) above, we take the inductances of the coils 1 and 3 and of each half of the centre-tapped coil 2 to be the same, the impedances

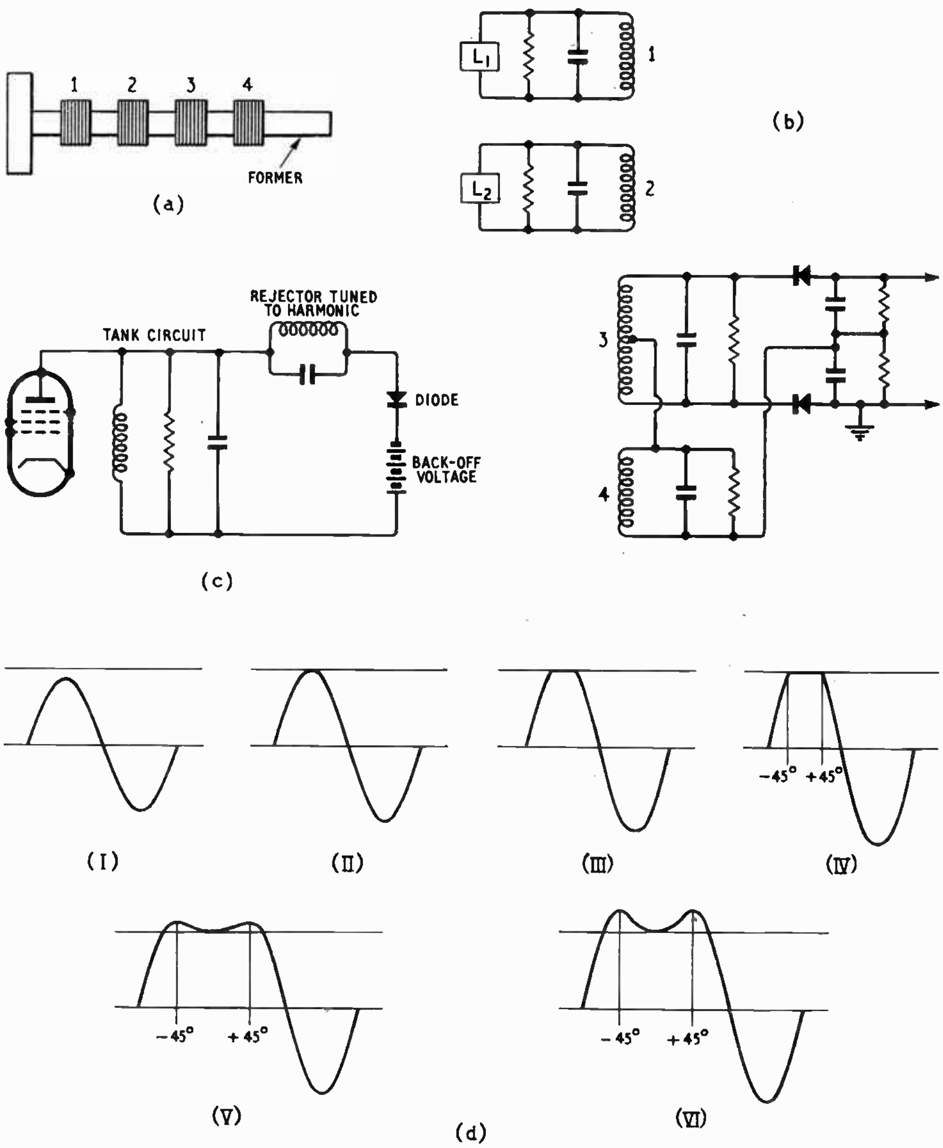


Fig. 4. *General arrangement when more than one diode limiter is needed*

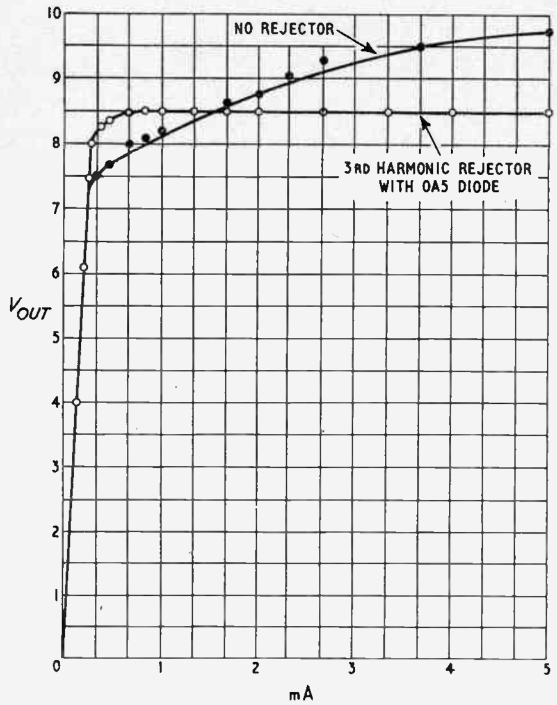


Fig. 5. *Fundamental current plotted against fundamental voltage for diode limiter with added third-harmonic rejector circuit*

of these coils at the mean frequency $\omega_0/2\pi$ are taken as unity*, and the total effective conductances across coil 1, each half of coil 2 and coil 3 are taken as $2s\alpha$, $2s\beta$ and $2s\gamma$ respectively. If V_1 is the voltage across coil 1, V_2 the voltage across either half of the centre-tapped coil 2 and V_3 the voltage across coil 3 while i_1, i_2 and i_3 are the corresponding currents, then we find, as explained in more detail in the Appendix,

$$\left.\begin{aligned} V_1 &= \Delta_1 i_1/\Delta \\ V_2 &= (\gamma + jx)j\lambda i_1/\Delta \\ V_3 &= -\lambda\mu i_1/\Delta \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

where sx is, as before, the fractional frequency deviation, and

$$\left.\begin{aligned} \Delta_1 &= (\beta + jx)\ (\gamma + jx) + \mu^2 \\ \Delta &= (\alpha + jx)\ (\beta + jx)\ (\gamma + jx) \\ &\quad + \lambda^2(\gamma + jx) + \mu^2(\alpha + jx) \end{aligned}\right\} \quad \ldots \quad \ldots \quad (8)$$

The presence of the limiter in Fig. 3 means that we must regard α as a variable, adjustable so that $|V_1|$ remains constant. Now α appears explicitly in Δ [equation (8)] but nowhere else. Equation (7) can therefore be written

$$\left.\begin{aligned} V_2 &= (\gamma + jx)j\lambda V_1/\Delta_1 \\ V_3 &= -\lambda\mu V_1/\Delta_1 \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

These equations are of the same form as (1) except in so far as the input current i_1 in (1) is replaced by $j\gamma V_1$, $|V_1|$ being nearly constant. Thus, the same precision can be obtained with the circuit of Fig. 3 as with the Foster-Seeley discriminator circuit discussed in Reference 1. If the circuit of Fig. 3 is preceded by a second limiter and phase inverter, equations (7) are practically unaltered.

When discussing the Foster-Seeley discriminator in Reference 1, it was found desirable that a quantity σ which corresponds here to $|V_3/V_2|$ should be large. In Reference 1, this implied that the coil which was centre-tapped should also have a relatively large output. The centre-tapping and high coupling between the halves of this coil involves relatively large stray capacitances, so that there is an upper limit to the possible impedance, and the impedances of the other coils must be correspondingly reduced if σ is to be sufficiently large. In the case of the Foster-Seeley discriminator, this difficulty cannot be overcome by centre-tapping the primary coil, because the primary coil must have relatively high impedance in order to match the drive from the valve efficiently. The difficulties of tapping down with close coupling when the tap would also require to be centre-tapped with close coupling are so formidable that this is seldom attempted. With the construction here described, however, the inter-winding capacitances are small. The most suitable coil can therefore be centre-tapped. Since the input to this coil is inductive, optimum values of the turns and impedance can be chosen, and optimum primary matching conditions can also be chosen independently.

We have thus achieved a construction which combines the discriminator with a limiter, and gives performance in both respects as good as may be required. Owing to the advantageous distribution of stray capacitances, very high efficiency is obtainable.

APPENDIX

With the normalization and notation of the main text for the circuit of Fig. 3, consider first the coils alone. The relation between currents and voltages is, to the first order in s,

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \end{bmatrix} = p \begin{bmatrix} 1 & 2s\lambda & 0 \\ 2s\lambda & 1 & 2s\mu \\ 0 & 2s\mu & 1 \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (A1)$$

where $p = j(1 + sx)$.

Inverting this matrix, we have [cf. Reference 1, Equation (C8)]

$$\begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} = \frac{1}{p} \begin{bmatrix} 1 & -2s\lambda & 0 \\ -2s\lambda & 1 & -2s\mu \\ 0 & -2s\mu & 1 \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \end{bmatrix} \quad \ldots \quad \ldots \quad \ldots \quad (A2)$$

If each winding is tuned with all other windings short-circuited, the current-voltage relation (A2) only has to have the principal-diagonal terms $1/p$ replaced by $q = p + 1/p \approx 2jsx$, in order to allow for the capacitances: to allow for losses and loads, the principal-diagonal terms must be further increased by $2s\alpha$, $2s\beta$ and $2s\gamma$ respectively, so that the complete current-voltage relation when all elements are taken into account is

$$\begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} = 8s^3 \begin{bmatrix} \alpha + jx & -\lambda/p & 0 \\ -\lambda/p & \beta + jx & -\mu/p \\ 0 & -\mu/p & \gamma + jx \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \end{bmatrix} \quad \ldots \quad \ldots \quad (A3)$$

We now have to obtain V_1, V_2, V_3 when $i_2 = i_3 = 0$ since only coil 1 in Fig. 3 has any external supply. The determinant Δ of the matrix in (A3) is

$$\Delta = (\alpha + jx)\ (\beta + jx)\ (\gamma + jx) + \lambda^2(\gamma + jx) + \mu^2(\alpha + jx) \quad (A4)$$

if we put -1 instead of p^2, so that only a second-order error is made in (A3), we then find that

$$\left.\begin{aligned} V_1 &= \Delta_1 i_1/\Delta \\ V_2 &= i_1(\gamma + jx)j\lambda/\Delta \\ V_3 &= i_1 \cdot j\lambda \cdot j\mu/\Delta \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (A5)$$

as in (7) of the main text.

REFERENCE

[1] C. G. Mayo and J. W. Head, "The Foster-Seeley Discriminator", *Electronic and Radio Engineer*, February 1958, p. 44.

* See footnote to equation (4); a similar adjustment to equation (7) may be required.

THE LAMB SHIFT—2: Lamb and Retherford's Experiment

Last month's article dealt with two rather difficult concepts, the self-energy of the electron in its interaction with the radiation field, and the polarization of the positron-filled vacuum of Dirac's theory. Both are embarrassing to the mathematicians, for they lead to infinite terms which have to be spirited away by what Heitler has called 'wishful mathematics'. It was stated that these concepts lead to two effects that can be measured experimentally—the separation of energy levels that were previously calculated to be coincident, and a change in the magnetic moment of the electron.

A good deal of the time was spent on the fine-structure of the H_α line, and in explaining what was being sought for in the 'Lamb shift' experiment, performed in 1947 by W. E. Lamb and R. C. Retherford. This concentrated on the difference between the $2^2S_{\frac{1}{2}}$ and $2^2P_{\frac{1}{2}}$ levels; it should be mentioned that the effect is not confined to just these two, and occurs for every higher n value. Also, there are a number of minor contributions, which will not be discussed, to the 'total' Lamb shift. The present article will be devoted to explaining the experiment of Lamb and Retherford, as it is described by W. E. Lamb (now Wykeham Professor at Oxford, and Nobel Laureate) in the Physical Society's 'Report on Progress in Physics' for 1951.

Lamb and Retherford's Experiment

Although the experiment was not actually done quite in this way, the general principle was to produce a collimated atomic beam consisting of H atoms excited to the $2^2S_{\frac{1}{2}}$ state. If this is above the $2^2P_{\frac{1}{2}}$ state, with an energy difference $\Delta E = h\nu_0$, then one possible transition is to the $2^2P_{\frac{1}{2}}$ state; if the two states are identical, this possibility is absent. The beam is passed through a radio-frequency field, and impinges on a detector which responds only to excited atoms. If the frequency ν of the field is close to ν_0, then the transition is stimulated, $2^2S_{\frac{1}{2}}$ atoms are lost to the detector, and the signal decreases; resonance, with minimum detector signal, occurs when $\nu = \nu_0$.

The first problem was to produce a beam of H atoms, to raise them to the $2^2S_{\frac{1}{2}}$ state (which is 10·2 eV above the normal ground state) and to keep them in this state during their passage through the field. Thermal dissociation of molecular hydrogen in a tungsten furnace provided the atoms, and they were excited by bombardment with electrons of energy slightly greater than 10·2 eV, directed at right angles to the beam. A good deal of theoretical work had been done on the lifetime of the $2^2S_{\frac{1}{2}}$ state, and it was believed to be metastable, with quite a long lifetime in the absence of an external electric field. This would have been rather a difficult condition to fulfil in any case, and adequate longevity was ensured by using a transverse magnetic field, as explained later on.

The detector was a tungsten target; as the work function for tungsten is about 4·5 volts, the 10·2-eV carried by the excited atoms enabled them to eject

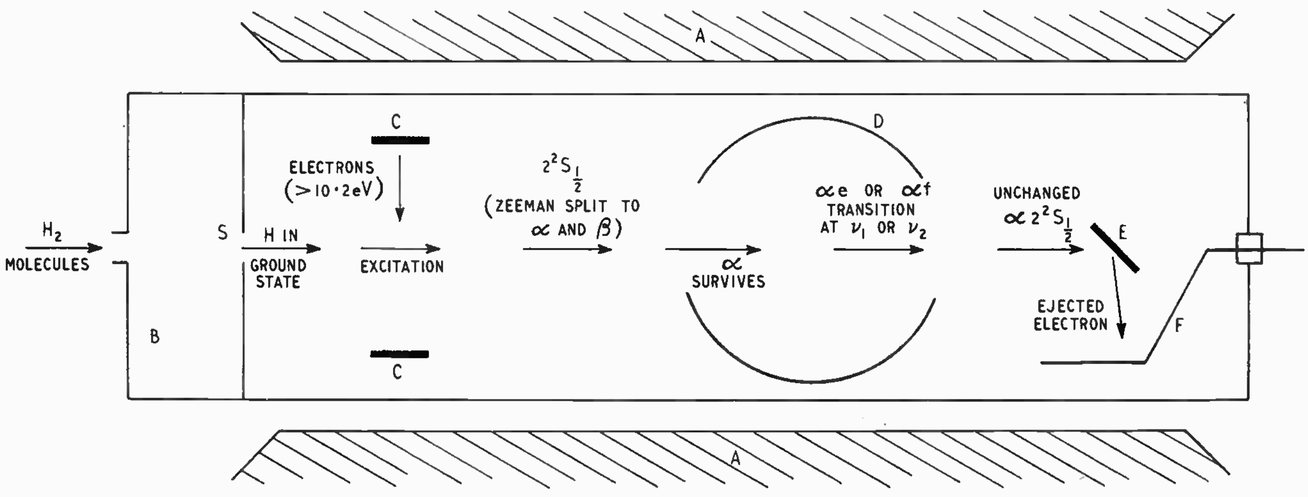


Fig. 1. *General scheme of Lamb and Retherford's experiment. A, magnet pole pieces; B, dissociation oven; C, electron bombarder; D, radio-frequency region; E, tungsten target; F, electron collector. All to the right of the fine slit S, by which atoms of hydrogen (or deuterium) enter the system, is evacuated*

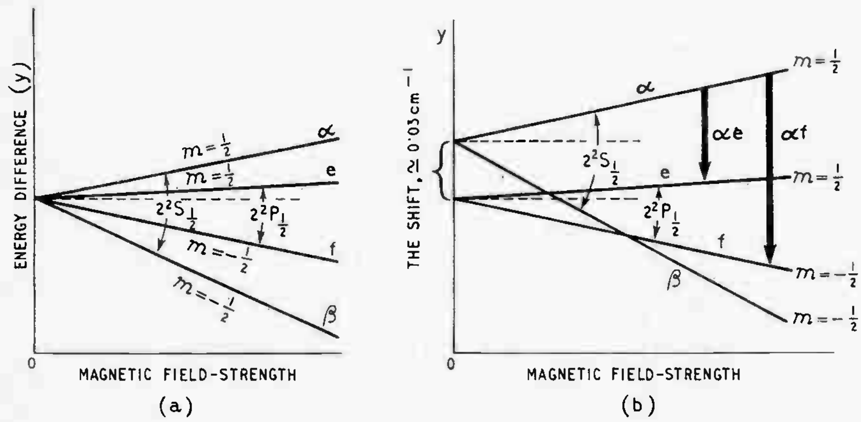


Fig. 2. Zeeman splitting of the $2^2S_{1/2}$ and $2^2P_{1/2}$ levels of H in a magnetic field; (a) according to the Dirac theory, (b) according to the newer theory. These diagrams have been simplified from those given in Prof. Lamb's paper so as to include only the information needed for the present article. The energy difference y is in wave-numbers and, in the original diagram, was measured with respect to the 'centre of gravity' of the level; the scale is not here shown, but the full spread $\alpha\beta$ is about 0·2 cm^{-1} and the greatest field-strength is about 5,000 c.g.s. units. The state β is so short-lived as to be quite unimportant, and the only two transitions actually measurable are those indicated, namely αe and αf

electrons from the metal on returning to their ground state, and these electrons were collected by an electrode close to the target. The general scheme is shown in Fig. 1.

The intricate part of the experiment was the use of the magnetic field. In the first place, this causes a Zeeman splitting of the $m = +\frac{1}{2}$ and $m = -\frac{1}{2}$ levels (Fig. 2), so that the 'natural' $2^2S_{\frac{1}{2}}$ and $2^2P_{\frac{1}{2}}$ states are not really there at all. Each has now become two components. The two $2^2S_{\frac{1}{2}}$ components are labelled α $(+\frac{1}{2})$ and β $(-\frac{1}{2})$, and the two $2^2P_{\frac{1}{2}}$ components e $(+\frac{1}{2})$ and f $(-\frac{1}{2})$ respectively. However, the difference between the Dirac calculation and the result being investigated persists all along the lines. The resonance frequency is no longer ν_0; indeed, there are two—ν_1 for the transition αe, and ν_2 for the transition αf. The values of ν_1 and ν_2 depend on the field strength. Instead of varying the radio-frequency ν, this was kept fixed, and the field strength varied until resonances for $\nu_1 = \nu$ and for $\nu_2 = \nu$ were obtained. From these, working back along the graphs of Fig. 3, as it were, ν_0 was calculated to be 1062 $\pm$ 5 Mc/s, or 0·0354 cm^{-1}.

It will be noted that although the state β is shown in the figures, only ν_1 and ν_2 have been mentioned; this is not just to simplify the story, for it happened that while state α had a lifetime of 7×10^{-5} sec, which was ample for survival over the 6-cm path to the detector, state β lasted only for about one five-hundredth of this under the conditions obtaining. The reason for this is pretty complicated, for the overall result depends on the strengths of the magnetic field applied and of the electric field present, and also on the value of m; and it turned out that with the actual field values $m = +\frac{1}{2}$ was favoured and $m = -\frac{1}{2}$ was not.

Hyperfine Structure

There is a further complication which now has to be faced. The $2^2S_{\frac{1}{2}}$ and $2^2P_{\frac{1}{2}}$ levels were Zeeman-split into α,β and e,f. But these four levels, as drawn in Fig. 2, are in their turn not really there at all—each is a *pair* of closely separated states. This hyperfine structure is due to coupling between the spin of the nucleus and the electron spin; in very weak magnetic fields, each m has two possibilities, either parallel or antiparallel to the nuclear spin; in strong magnetic fields the coupling itself breaks down, and each spin precesses independently, but there is still an interaction. A better way of putting this is to say that the magnetic field of the nucleus operates on the magnetic moment of the electron, and can do this in two ways. The separation depends on the resultant spin of the nucleus, and is also proportional to $1/n^3$, where n is the principal quantum number.

Hyperfine level separation is just measurable optically, using a high-resolution interferometer. In general, it lies nicely within the range of frequencies covered by microwave spectroscopy and paramagnetic resonance techniques. It has also been determined, at least for the ground states of unexcited atoms, by the atomic ray method of Rabi and his collaborators. For the $n = 1$ levels of hydrogen it is 0·048 cm^{-1}, which is greater than the Lamb shift itself; for the $n = 2$ levels selected for the experiment it is 0·006 cm^{-1} which, though still appreciable, is less than the width of the resonance curve. But with deuterium, the combination of proton and neutron in the nucleus has a much smaller magnetic moment than the single proton of hydrogen, and gives about a quarter of the hyperfine level separation. The resonances obtained by Lamb and Retherford with deuterium were therefore very much sharper; they appeared to be the same as for hydrogen, though later work revealed a small difference in the two cases.

Really accurate measurement of hyperfine level separation has shown it to be a few parts in a thousand greater than the older theory suggested, and also confirmed (by comparing results for H and D) that the discrepancy does not come from the nuclear side. It is consistent with an increase in the magnetic moment of the electron, above its Bohr-magneton value $he/4\pi mc$, due to interaction with the radiation field. Nothing better than an approximate calculation of the anomalous magnetic moment seems to have been managed theoretically, but the best approximation, 1·0016 Bohr magnetons, does fit the observations fairly well. This figure is quoted from Heitler; it may well be superseded now, for people have been working for the last ten years or so to try to get a closer approximation. The importance of this apparently small difference is that, like the Lamb shift itself, it is a direct confirmation of the self-energy and vacuum-polarization concepts of the newer theory. I hope, by the way, that the point of this section is clear. In Lamb and Retherford's experiment, hyperfine structure was more or less a nuisance that was circumvented when deuterium was used. But the effect itself, investigated by other workers, revealed the anomalous magnetic moment of the electron and so has reinforced the underlying theory with a quite independent check.

Agreement with Calculation

In the last article, the level shift was referred to as 0·033 cm^{-1}, which was in fact the suspected value before the measurements were made. In Lamb's paper, the result (1062 $\pm$ 5) Mc/s (0·0354 cm^{-1}) is compared with a calculated value 1051 Mc/s (0·0350 cm^{-1}). The value first calculated by Bethe in 1947 was 1040 Mc/s (0·0348 cm^{-1}).

The most recent figures I have found are those given by E. E. Salpeter (*Phys. Rev.*, 89, 92, 1953). He gives the experimental results obtained in 1952 by Triebwasser, Dayhoff, and Lamb as:

for H: (1057·77 $\pm$ 0·10) Mc/s, and
for D: (1059·00 $\pm$ 0·10) Mc/s.

Taking a number of correction terms into account, including an effect due to an anomalous magnetic moment of the nucleus, Salpeter gives theoretical values which agree to within half a megacycle with these results.

Conclusion

The picture presented in the previous article may have appeared a little fanciful; but I assure you that I checked carefully with W. Heitler's 'Quantum Theory of Radiation' (3rd Ed., 1954) to see that it was on the whole a fair representation. Perhaps it is best to quote Heitler's comments on the situation. He says, 'All the infinities that occur in the theory originate from a few divergent quantities that are unobservable, namely, a contribution to the mass and to the charge of a particle due to its interaction with the quantum field. If we pretend not to notice that the contributions in question turn out to be infinite and therefore cannot be handled in a way free from mathematical objection, there is no obstacle in the way of giving an unambiguous answer to every legitimate physical question. The outstanding success of this latest phase of the theory is the quantitative explanation of the radiative displacement of atomic energy levels and the magnetic moment of the electron additional to the usual magneton. These two effects belong to the most accurate confirmations of the theory extant. It appears therefore that the theory as it stands must be strangely near the final solution, yet it cannot possibly be correct, as the mathematical procedure used to extract these results is plainly unacceptable'.

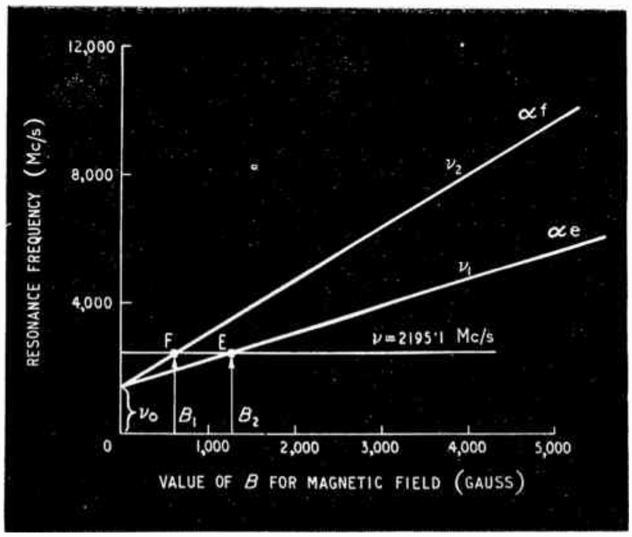


Fig. 3. *Variation of resonance frequency with magnetic field for each of the two transitions αe and αf. Though, like Fig. 2, this has been very much simplified from the original, it serves to illustrate the principle used to find ν_0. With ν fixed at 2195·1 Mc/s, the values B_1 and B_2 for the resonances ν_1 and ν_2 are found. Thus, points E and F on the two graphs are identified, whence the intercept ν_0 is obtained; at least, that is the general idea*

You notice that he says '*the* theory'. This is something much more than a mere explanation of the antics of a bound electron in a radiation field, and their effect on electronic energy levels. It is what they call *quantum electrodynamics*, and is an attempt to construct a complete theory of the short-range forces between particles, why the proton and the neutron hold together in the deuteron, the part played by the meson in nuclear architecture, and other fundamental mysteries which seem almost on the verge of being partially understood. Samuel Glasstone's 'Sourcebook on Atomic Energy' (1952) says of this that 'there are some charming speculations which may contain some grains of truth'. Several authorities combined together to give a valuable discussion, 'A Survey of Field Theory', in the 1953 volume of the Physical Society's 'Reports on Progress in Physics'. In this, R. E. Peierls comments that the only reason they have been able to get to the present stage in their sums is the small value of the 'fine-structure constant', the 1/137 which is always turning up, and powers of which are able to be discarded for approximate calculations.

These two 'Fringe' articles on the Lamb Shift cannot have done more than merely to indicate the difficulties that were facing the theory a little while ago, and to show the support that it received from a most delicate piece of electronic and radio engineering. I should have been able to find out a good deal more for you if I had waited a little, for I see that Prof. Lamb is to give the Physical Society's Guthrie Lecture in May, on 'Experimental Tests of Quantum Electrodynamics'. But my only real apology is for treating under the heading of 'Fringe' something which is so important that it is really at the very centre of the field of modern physics.

INTERNATIONAL CONVENTION ON MICROWAVE VALVES

The Institution of Electrical Engineers is organizing an International Convention on Microwave Valves to be held in London from 19th to 23rd May 1958. The Convention will survey present knowledge and discuss future trends, with particular attention to work in progress or recently completed.

EXHIBITIONS

May 12–21. The Production Exhibition, Grand Hall, Olympia, London, W.14.

May 19–24. First European Television Exhibition, Park Lane House, 45 Park Lane, London, W.1.

July 10–12 and 14–16. Institution of Electronics, College of Science and Technology, Manchester.

These exhibitions are in addition to those in the list on page 77 of our February issue.

Amplifier Low-Frequency Compensation

MAXIMALLY-FLAT INDICIAL, FREQUENCY, AND PHASE RESPONSES

By J. E. Flood, Ph.D., A.M.I.E.E.* and J. E. Halder, M.Sc., A.R.C.S.*

SUMMARY. *General expressions are deduced for the indicial response, the gain-frequency response and the phase-frequency response at low frequencies. The expressions are used to obtain the conditions for maximal flatness of the indicial response, gain-frequency response or phase-frequency response of particular circuits. A single resistance-capacitance coupled stage can have up to second-order compensation of its indicial response, up to fourth-order compensation of its phase-frequency response or up to fifth-order compensation of its gain-frequency response. The design of multistage amplifiers is also considered.*

When negative feedback is applied with a feedback fraction β which is independent of frequency, the overall indicial response or phase-frequency response is maximally flat when the corresponding response of the μ-path is maximally flat. A maximally-flat gain-frequency response, however, is not obtained when the μ-path has a maximally-flat gain-frequency response. Some simple relationships are found between the conditions for maximal flatness of the indicial response when anode decoupling circuits are added either inside or outside the feedback loop or a time-constant is added to the β-path. No such relationships hold for the gain-frequency response or the phase-frequency responses.

In a.c.-coupled amplifier circuits, the gain at low frequencies departs from its value at high frequencies and a phase shift is present. The indicial response (i.e., the response to a unit step input) is consequently affected, causing it to depart from the ideal flat top. The bandwidth of an amplifier is usually large enough for the effects of distortion at low frequencies and high frequencies to be considered independently. Analyses of low-frequency distortion can therefore be made from equivalent circuits which omit stray capacitances and components which are effective only at high frequencies. The gain-frequency response of the equivalent circuit may be made as flat as possible down to low frequencies by making as many as possible of its differential coefficients with respect to frequency vanish at infinite frequency; the gain-frequency response is then said to be maximally flat. The phase-frequency response can similarly be made maximally flat by causing as many as possible of its differential coefficients to vanish at infinite frequency. The indicial response can be made maximally flat by causing as many as possible of its differential coefficients with respect to time to vanish at $t = 0$.

In general, the conditions for maximally flat gain-frequency response, phase-frequency response and indicial response are different; the response which should be made optimum depends on the application for which the amplifier is intended. Thus, a video-frequency or pulse amplifier should have a maximally-flat indicial response, a voltage-measuring instrument needs a maximally-flat gain-frequency response and other applications require a maximally-flat phase-frequency response. The effects of low-frequency distortion on indicial response have been analysed for some typical amplifier circuits in an earlier paper[1]. This paper extends the analysis to include the conditions for maximally-flat gain-frequency and phase-frequency responses.

General Theory

Indicial Response

The Heaviside operational form of the indicial response is

$$h(p)\,\mathbf{1} = \frac{p^n + a_1 p^{n-1} + \ldots + a_n}{p^n + b_1 p^{n-1} + \ldots + b_n}\,\mathbf{1} \qquad \ldots \quad (1)$$

This can be expanded in descending powers of p, giving the following result[1].

$$h(p) = 1 + \frac{e_1}{p} + \frac{e_2}{p^2} + \ldots + \frac{e_r}{p^r} + \ldots \qquad \ldots \quad (2)$$

$$\text{where } e_r = \sum_{s=0}^{r-1} g_s\,(a_{r-s} - b_{r-s}) \qquad \ldots \quad (3)$$

$$\text{and} \quad g_s = \Sigma \frac{(-1)^m\, m!}{\alpha!\,\beta!\ldots\epsilon!}\, b_\lambda^{\alpha}\, b_\mu^{\beta} \ldots b_\rho^{\epsilon} \qquad \ldots \quad (4)$$

the second summation extending over all possible positive integral values of $\alpha, \beta, \ldots \epsilon$; $\lambda, \mu, \ldots \rho$ such that

$$\alpha + \beta + \ldots + \epsilon = m \;;\; \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$$
$$1 < \lambda < \mu < \ldots < \rho \leqslant s \;;\; a_0 = b_0 = g_0 = 1$$

Since the Heaviside transform of $t^r/r!$ is $1/p^r$, it follows from equation (2) that

$$h(t) = 1 + e_1 t + e_2 \frac{t^2}{2!} + \ldots + e_r \frac{t^n}{r!} + \ldots$$

* *Siemens Edison Swan Research Laboratory.*

If the rth differential coefficient of $h(t)$ at $t = 0$ is $h_r(0)$, it follows from Maclaurin's theorem that

$$h(0) = 1,\ h_1(0) = e_1, \ldots h_r(0) = e_r, \ldots \text{etc.}$$

The indicial response may be made to approach more and more nearly to the ideal by making more and more of the differential coefficients vanish at $t = 0$. Equation (3) shows that the first r differential coefficients will vanish if, and only if, $a_1 = b_1$, $a_2 = b_2$, ... $a_r = b_r$. The amplifier will then be said to have rth-order compensation of its indicial response. Since the condition $a_n = b_n$ cannot usually be attained, the indicial response will be called maximally flat if it has $(n - 1)$th-order compensation.

Gain-Frequency Response

It can be shown[1] that the gain-frequency response $G(\omega)$ corresponding to the indicial response of equation (1), is given by

$$\left\{G(\omega)\right\}^2 = \frac{\omega^{2n} + c_2\ \omega^{2n-2} + \ldots + c_{2n}}{\omega^{2n} + d_2\ \omega^{2n-2} + \ldots + d_{2n}} \quad \ldots \quad (5)$$

$$\left.\begin{aligned} \text{where } c_{2r} &= a_r^2 + 2 \sum_{s=1}^{r} (-1)^s\, a_{r+s}\, a_{r-s} \\ d_{2r} &= b_r^2 + 2 \sum_{s=1}^{r} (-1)^s\, b_{r+s}\, b_{r-s} \end{aligned}\right\} \quad \ldots \quad (6)$$

If $\{G(\omega)\}^2$ is expanded in descending powers of ω, then

$$\left\{G(\omega)\right\}^2 = 1 + \frac{e_2}{\omega^2} + \frac{e_4}{\omega^4} + \ldots + \frac{e_{2r}}{\omega^{2r}} + \ldots \quad (7)$$

where

$$e_{2r} = \sum_{s=0}^{r-1} g_s \left\{c_{2(r-s)} - d_{2(r-s)}\right\} \quad \ldots \quad \ldots \quad (8)$$

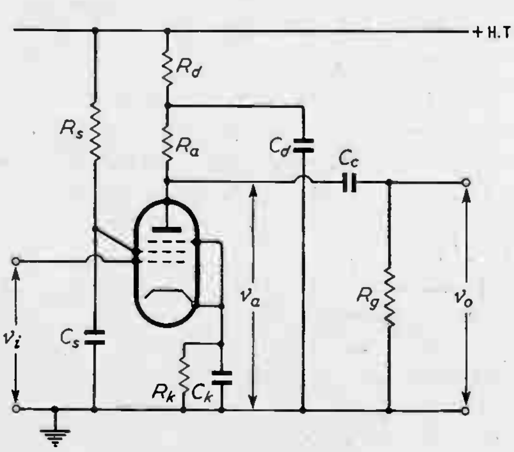
and g_s is given by equation (4). By Maclaurin's theorem, the rth derivative of $\{G(\omega)\}^2$ with respect to $1/\omega$ at infinite frequency is given, from equation (7), by

$$f_r(\infty) = r!\, e_r$$

The gain-frequency response has rth order compensation if the first r differential coefficients are zero at infinite frequency. Since $e_r = 0$ for odd values of r, the gain-frequency response has $(2r + 1)$th-order compensation if

$$e_2 = e_4 = \ldots = e_{2r} = 0$$

Fig. 1. *Resistance-capacitance coupled stage using pentode valve*



Equation (8) shows that this is so only if

$$c_2 = d_2,\ c_4 = d_4, \ldots c_{2r} = d_{2r}.$$

A sufficient, but not necessary condition for this is that $a_1 = b_1$, $a_2 = b_2$, ... $a_{2r} = b_{2r}$. Thus, if the indicial response has $2r$th-order compensation, the gain-frequency response has $(2r + 1)$th-order compensation. Since the condition $c_{2n} = d_{2n}$ cannot usually be attained, the gain-frequency response will be called maximally flat if its order of compensation is $(2n - 1)$.

Phase-Frequency Response

It can be shown[1] that the phase-frequency response corresponding to the indicial response of equation (1) is

$$\tan\phi = \frac{c_1\omega^{2n-1} + c_3\omega^{2n-3} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2\ \omega^{2n-2} + \ldots + d_{2n}} \quad \ldots \quad (9)$$

where

$$\left.\begin{aligned} c_{2r-1} &= (-1)^r \sum_{s=1}^{r} \left(a_{2s-1}b_{2(r-s)} - a_{2(r-s)}b_{2s-1}\right) \\ \text{and } d_{2r} &= (-1)^r \left\{ b_{2r} + \sum_{s=1}^{r} \left(a_{2s}b_{2(r-s)} - a_{2s-1}b_{2(r-s)+1}\right)\right\} \end{aligned}\right\} (10)$$

$\tan\phi$ can be expanded in descending powers of ω:

$$\tan\phi = \frac{e_1}{\omega} + \frac{e_3}{\omega^3} + \ldots \frac{+\ e_{2r+1}}{\omega^{2r+1}} + \ldots \quad \ldots \quad (11)$$

$$\text{where } e_1 = c_1,\ e_{2r} = 0$$

$$\text{and } e_{2r+1} = \sum_{s=0}^{r-1} g_s \left(c_{2r+1-s} - c_1\, d_{2r-s}\right) \quad \ldots \quad \ldots \quad (12)$$

where g_s is given by equation (4). The phase-frequency response has rth-order compensation if the first r derivatives of $\tan\phi$ with respect to $1/\omega$ are zero at infinite frequency. By Maclaurin's therorem, the rth derivative of $\tan\phi$ at infinite frequency, $f_r(\infty)$, is given, from equation (11) by

$$f_r(\infty) = r!\, e_r$$

Since $e_r = 0$ for even r, the phase-frequency response has $2r$th-order compensation if $e_1 = e_3 = \ldots = e_{2r-1} = 0$. Equation (11) shows that this is so only if $c_1 = c_3 = \ldots = c_{2r-1} = 0$. A sufficient, but not necessary condition for this is that

$$a_q = b_q \text{ for } q = 1, 2, \ldots 2r - 1.$$

Thus if the indicial response has $(2r-1)$ th-order compensation, then the phase-frequency response has $2r$ th-order compensation. The converse is not necessarily true. Since we usually cannot make $c_{2n-1} = 0$, the phase-frequency response will be called maximally flat if it has $(2n - 2)$th-order compensation.

The RC-Coupled Stage

Fig. 1 shows a typical resistance-capacitance coupled amplifier using a pentode valve. The usual assumptions will be made, namely :—

(1) The anode and screen currents are independent of anode voltage, and the ratio between them is independent of grid, screen and anode voltages.

(2) The impedance of the d.c. power supply is negligible.

(3) The grid resistance R_g is large compared with the anode load resistance, R_a, and the decoupling resistance, R_d. Then the gain between grid and anode can be

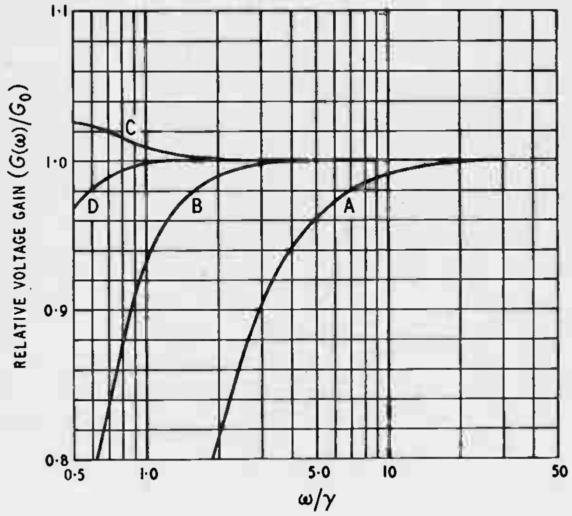


Fig. 2. *Gain-frequency response of amplifier stage; Curve A, coupling circuit alone* ($\alpha = \sqrt{2}\gamma$); *B, coupling circuit and anode decoupling; C, circuit of Fig.* 1, 3*rd-order compensation; D, circuit of Fig.* 1, 5*th-order compensation*

expressed[2] as

$$\frac{V_a}{V_i} = \frac{g_m\,(R_a + Z_d)}{1 + g_k\, Z_k + \sigma_s\, Z_s}$$

where g_m = grid-anode mutual conductance

g_k = sum of screen conductance and grid-anode, grid-screen and screen-anode mutual conductances

σ_s = screen conductance

Z_d, Z_k, Z_s, are the impedances of the anode, cathode and screen decoupling circuits, respectively.

The notation used will be:

$$\gamma = 1/R_aC_d,\ \lambda = 1/R_dC_d,\ \eta = 1/R_kC_k,\ \theta = 1/R_sC_s$$
$$\alpha = 1/R_gC_c,\ G_0 = g_mR_a$$

If v_i is a unit step, then

$$v_a = G_0 \frac{1+\gamma/(p+\lambda)}{1+g_kR_k\ \eta/(p+\eta)+\sigma_sR_s\ \theta/(p+\theta)}\,\mathbf{1} \qquad (13)$$

The operational expression for v_0/v_a is

$$\frac{v_0}{v_a} = \frac{p}{p+\alpha}\,\mathbf{1} \qquad (14)$$

Then, from equations (13) and (14), the overall indicial response is

$$v_0\,\mathbf{1} = G_0\,\frac{p^4 + a_1\,p^3 + a_2\,p^2 + a_3\,p}{p^4 + b_1\,p^3 + b_2\,p^2 + b_3\,p + b_4}\,\mathbf{1} \qquad (15)$$

where

$$a_1 = \gamma + \lambda + \eta + \theta$$
$$a_2 = \eta\,\theta + (\gamma + \lambda)\,(\eta + \theta)$$
$$a_3 = \eta\,\theta\,(\gamma + \lambda)$$
$$b_1 = \alpha + \lambda + \eta\,(1 + g_kR_k) + \theta\,(1 + \sigma_sR_s)$$
$$b_2 = \alpha\,\lambda + (\alpha + \lambda)\,\{\eta\,(1 + g_kR_k) + \theta\,(1 + \sigma_sR_s)\} + \eta\,\theta\,(1 + g_kR_k + \sigma_sR_s)$$
$$b_3 = \alpha\,\lambda\,\{\eta(1 + g_kR_k) + \theta\,(1 + \sigma_sR_s)\} + \eta\,\theta\,(\alpha + \lambda)\,(1 + g_kR_k + \sigma_sR_s)$$
$$b_4 = \alpha\,\lambda\,\eta\,\theta\,(1 + g_kR_k + \sigma_sR_s)$$

Compensation for Anode Decoupling by means of Cathode and Screen Decoupling

At low frequencies the increase in the anode load, due to the finite reactance of C_d, causes an increase in gain, but there is a fall in gain due to negative feedback introduced by the finite reactance of C_k and C_s. Edwards and Cherry[2] have shown that these effects can be made to cancel, so that the gain between grid and anode is independent of frequency. From equation (13) it is evident that the indicial response of the anode voltage is distortionless if

$$\left.\begin{array}{l}\gamma = \eta\ g_kR_k + \theta\ \sigma_sR_s \\ \text{and}\ \lambda = \eta = \theta\end{array}\right\} \qquad (16)$$

If these conditions are satisfied there will be no distortion of the anode indicial response and so no gain or phase distortion at low frequencies. Other circuits in which an increase in anode-load impedance offsets screen and cathode feedback at low frequencies have been described by Zeidler and Noe[3]. Unless there is d.c. coupling to the next stage, however, distortion will be introduced by the coupling capacitance.

Compensation for the Coupling Circuit by means of the Anode Decoupling Circuit

The gain is reduced at low frequencies by the increasing reactance of C_c. This may be offset to a certain extent by the increase in gain caused by the anode decoupling capacitance, although perfect compensation cannot be obtained because the increase in gain due to the decoupling circuit is always finite whereas, at zero frequency, the coupling circuit reduces the gain to zero. This compensating effect has been analysed by several authors[4, 5, 6]. The effect of the screen and cathode decoupling can be avoided, either by removing the circuits, in which case equation (13) becomes

$$v_a = G_0\left\{1 + \frac{\gamma}{p+\lambda}\right\}\mathbf{1} \qquad (17)$$

or by making their time-constants very large compared with those of the coupling and anode decoupling circuits, in which case equation (13) becomes

$$v_a = G_0\,\frac{1 + \gamma/(p + \lambda)}{1 + g_kR_k + \sigma_sR_s}\,\mathbf{1} \qquad (18)$$

Since the expressions for v_a in equations (17) and (18) differ only by a constant factor, the two alternatives are essentially equivalent.

From equations (14) and (17) the overall indicial response is

$$v_0 = G_0\,\frac{p^2 + p\,(\gamma + \lambda)}{p^2 + p\,(\alpha + \lambda) + \alpha\lambda}\,\mathbf{1} \qquad (19)$$

First-order compensation is obtained when $\alpha = \gamma$. This is the maximally-flat indicial response. The gain-frequency response corresponding to equation (19) has third-order compensation, and is maximally flat, if

$$\alpha = \sqrt{(\gamma^2 + 2\gamma\lambda)} \qquad (20)$$

The phase-frequency response corresponding to equation (19) has second-order compensation and is maximally flat if $\alpha = \gamma$. Thus, the condition for maximally flat phase-frequency response corresponds to a maximally flat indicial response, but that for maximally-flat gain-frequency response does not.

Higher-Order Compensation

Second-order compensation of the overall indicial response for the circuit of Fig. 1 can be obtained by taking into account the screen and cathode decoupling circuits. The response is given by equation (15).

From this, first-order compensation requires

$$\gamma = \alpha + \eta\, g_k R_k + \theta \sigma_s R_s \quad .. \quad .. \quad .. \quad (21)$$

for second-order compensation the additional requirement is[1]

$$\alpha = \frac{\eta\, g_k R_k\,(\eta - \lambda) + \theta\sigma_s R_s\,(\theta - \lambda)}{\lambda + \eta\, g_k R_k + \theta\sigma_s R_s} \quad .. \quad (22)$$

The maximally-flat response requires third-order compensation, but the condition for this cannot be satisfied[1]. Thus, second-order compensation is the highest obtainable and this occurs when equations (21) and (22) are satisfied.

The gain-frequency response corresponding to the indicial response given in equation (15) is

$$\{G(\omega)\}^2 = G_0^2\, \frac{\omega^8 + c_2\,\omega^6 + c_4\,\omega^4 + c_6\,\omega^2 + c_8}{\omega^8 + d_2\,\omega^6 + d_4\,\omega^4 + d_6\,\omega^2 + d_8} \quad (23)$$

where the coefficients are related to those in equation (15) by equations (6). From these, the condition for third-order compensation of the gain-frequency response is

$$a_1^2 - 2a_2 = b_1^2 - 2b_2$$

or

$$(\gamma+\lambda)^2 = \alpha^2+\lambda^2+2\eta\theta g_k R_k \sigma_s R_s + g_k R_k \eta(2+g_k R_k) + \sigma_s R_s \theta(2+\sigma_s R_s) \quad .. \quad .. \quad .. \quad (24)$$

Equation (24) is satisfied in the particular case when the indicial response has second-order compensation. Fifth-order compensation requires in addition that

$$a_2^2 - 2a_1 a_3 = b_2^2 - 2b_1 b_3 + 2b_4$$

or

$$\begin{aligned}(\eta^2 + \theta^2)\,(\gamma + \lambda)^2 &= \alpha^2\lambda^2 + \eta^2\theta^2\,(g_k R_k + \sigma_s R_s)(2 + g_k R_k + \sigma_s R_s) \\ &\quad + (\alpha^2+\lambda^2)\{\eta^2(1+g_k R_k)^2+\theta^2(1+\sigma_s R_s)^2 \\ &\quad + 2\eta\theta g_k R_k \sigma_s R_s\} \quad .. \quad .. \quad .. \quad .. \quad (25)\end{aligned}$$

The form or equation (23) suggest that the maximally-flat gain-frequency response requires seventh-order compensation. For this, in addition to equations (24) and (25), the parameters must satisfy

$$a_3^2 = b_3^2 - 2b_2 b_4$$

It appears, however, that this condition cannot be satisfied*. The highest order of compensation which can therefore be obtained for the gain-frequency response of the circuit shown in Fig. 1 is the fifth order. This is obtained when equations (24) and (25) are satisfied.

The phase-frequency response corresponding to the indicial response given by equation (15) is

$$\tan\phi = \frac{c_1\omega^7 + c_3\omega^5 + c_5\omega^3 + c_7\omega}{\omega^8 + d_2\omega^6 + d_4\omega^4 + d_6\omega^2 + d_8} \quad .. \quad (26)$$

where the coefficients are related to those in equation (15) by equations (10).

The condition for second-order compensation is

$$a_1 = b_1$$

or $\gamma = \alpha + \eta g_k R_k + \theta\sigma_s R_s \quad .. \quad .. \quad .. \quad (27)$

which is exactly the condition for first-order compensation of the indicial response. Thus, in this case, first-order indicial compensation and second-order phase-frequency compensation are entirely equivalent. To get fourth-order compensation of the phase-frequency response the additional requirement is

$$a_3 + a_1 b_2 = b_3 + b_1 a_2$$

or

$$\begin{aligned}&\eta^3 g_k R_k(1+g_k R_k)+\theta^3\sigma_s R_s(1+\sigma_s R_s)+\eta\theta(\eta+\theta)g_k R_k \sigma_s R_s \\ &= \alpha\lambda(\alpha+\lambda)+(\eta g_k R_k+\theta\sigma_s R_s)^2(\alpha+\lambda) \\ &\quad + (\eta g_k R_k+\theta\sigma_s R_s)(\alpha+\lambda)^2 \quad .. \quad (28)\end{aligned}$$

Equation (28) would be satisfied if the indicial response had third-order compensation, but this cannot be achieved. However, fourth-order phase-frequency compensation can be obtained, as will be indicated in the numerical example given below.

The form of equation (26) suggests that the maximally flat phase-frequency response corresponds to sixth-order compensation. This requires

$$a_2 b_3 = a_1 b_4 + a_3 b_2$$

It appears, however, that this condition cannot be satisfied.*

The highest order of compensation which can therefore be obtained for the phase-frequency response of the circuit shown in Fig. 1 is the fourth order. This is obtained when equations (27) and (28) are satisfied.

Numerical Examples

The example taken is an amplifier stage using an EF91 valve. Typical parameters for this valve are:

Grid-anode mutual conductance	=	7·6 mA/V
Grid-screen mutual conductance	=	2·0 mA/V
Screen-anode mutual conductance	=	0·1 mA/V
Screen conductance	=	0·029 mA/V

(under the conditions: $V_a = V_s = 250$V, $V_g = -2$V, $I_a = 10$ mA, $I_s = 2\cdot5$ mA).

From these figures $g_k = 9\cdot7$ mA/V and if $R_k = 160\,\Omega$ and $R_s = 18\text{k}\Omega$, then $g_k R_k = 1\cdot5$, $\sigma_s R_s = 0\cdot5$.

Compensation of the anode decoupling for indicial response by means of the screen and cathode decoupling circuits requires, from equations (16), that the time-constants of the anode, cathode and screen decoupling circuits should be equal, and $R_d = 2R_a$. The resulting zero-order indicial response is an exponential curve with the time-constant of the coupling circuit. If the effects of screen and cathode circuits can be made negligible, then the anode decoupling circuit can be used to compensate for the coupling circuit. If we keep $R_d = 2R_a$, the indicial response has first-order compensation and is maximally flat when $\alpha = \gamma$.

From equation (19) the indicial response is then

$$v_0(t) = G_0\,\{1 - 0\cdot25\,(\gamma t)^2 + 0\cdot125\,(\gamma t)^3 - \ldots\ldots\}$$

If all the time-constants are made finite and we keep $\gamma = 2\lambda$ and put $\eta = 2\theta$ then equations (21) and (22) show that second-order compensation can be obtained when

$$\alpha = 0\cdot016\gamma\ ,\ \theta = 0\cdot28\gamma$$

From equation (15) the indicial response is then

$$v_0(t) = G_0\,[1 - 0\cdot00037\,(\gamma t)^3 + 0\cdot0020\,(\gamma t)^4 - 0\cdot00067\,(\gamma t)^5 + \ldots\ldots]$$

* A rigorous proof of this has unfortunately not yet been obtained, but the authors have been unable to find any particular case which satisfies the required condition.

If γ remains the same in each case, it is evident that second-order compensation gives a much better response than that resulting from the compensation for the coupling circuit by means of the anode decoupling circuit.

When the anode decoupling circuit is used to compensate for the coupling circuit and $\gamma = 2\lambda$, the condition for the maximally-flat gain-frequency response, from equation (20) is

$$\alpha = \sqrt{2\gamma} = 2\sqrt{2\lambda}$$

and then

$$\{G(\omega)\}^2 = G_0^2\left\{1 - 0\cdot5\left(\frac{\gamma}{\omega}\right)^4 + \ldots.\right\} \quad .. \quad (29)$$

If all the time-constants are made finite, the gain-frequency response can have third or fifth-order compensation. Third-order compensation is obtained when equation (24) is satisfied and the indicial response has second-order compensation. The gain-frequency response is then:

$$\{G(\omega)\}^2 = G_0^2\left\{1 + 0\cdot096\left(\frac{\gamma}{\omega}\right)^4 + \ldots.\right\} \quad .. \quad (30)$$

Fifth-order compensation is obtained when equations (24) and (25) are satisfied. This can be done with

$$\eta = \theta = 0\cdot435\gamma$$
$$\alpha = \lambda = 0\cdot303\gamma$$

and then

$$\{G(\omega)\}^2 = G_0^2\left\{1 - 0\cdot0144\left(\frac{\gamma}{\omega}\right)^6 + \ldots.\right\} .. \quad (31)$$

The curves corresponding to equations (29), (30), and (31) are plotted in Fig. 2. This figure also shows as curve A the response due to the coupling circuit alone (when $\alpha = \sqrt{2\gamma}$). As would be expected, the higher the order of compensation the smaller the departure of the gain from its value at high frequencies. In the case of third-order compensation, however, the gain increases with decreasing frequency until a peak is reached which is only 3% above the gain at high frequencies. If this small peak in the gain-frequency is permissible, the third-order response is preferable to the fifth-order one, as is shown in Fig. 3.

When the anode decoupling circuit is used to compensate for the coupling circuit and $\gamma = 2\lambda$, the condition for the maximally-flat phase-frequency response is $\alpha = \gamma$ and then

$$\tan\phi = 0\cdot75\left(\frac{\gamma}{\omega}\right)^3 + 2\cdot06\left(\frac{\gamma}{\omega}\right)^5 + \ldots. \quad .. \quad (32)$$

If all the time-constants are made finite the phase-frequency response can have second or fourth-order compensation. Second-order compensation corresponds to first-order compensation of the indicial response. The phase-frequency response is then given by

$$\tan\phi = 0\cdot444\left(\frac{\gamma}{\omega}\right)^3 + 0\cdot641\left(\frac{\gamma}{\omega}\right)^5 + \ldots. \quad (33)$$

Fourth-order compensation is obtained when equations (27) and (28) are satisfied, which gives

$$\tan\phi = 0\cdot212\left(\frac{\gamma}{\omega}\right)^5 - 0\cdot0329\left(\frac{\gamma}{\omega}\right)^7 + \ldots. \quad (34)$$

The curves corresponding to equations (32), (33) and (34) are plotted in Fig. 4. This figure also shows as curve A the response due to the coupling circuit alone (when $\alpha = \gamma$). As is expected, the higher the order of compensation, the smaller is the phase-shift at low frequencies.

Stages in Tandem

Indicial Response

If two stages which are connected in tandem have rth-order and qth-order compensation of indicial response respectively, then the order of compensation of the complete amplifier is that of the stage with the lower order of compensation[1]. It is, however, sometimes possible to adjust the values of the circuit elements so that the overall indicial response has $(q + r)$th-order compensation[1].

Gain-Frequency Response

Let the gains of two stages be given by

$$\{G_1(\omega)\}^2 = \frac{\omega^{2n} + c_2\,\omega^{2n-2} + \ldots. + c_{2n}}{\omega^{2n} + d_2\,\omega^{2n-2} + \ldots. + d_{2n}}$$

and

$$\{G_2(\omega)\}^2 = \frac{\omega^{2m} + c'_2\,\omega^{2m-2} + \ldots. + c'_{2m}}{\omega^{2m} + d'_2\,\omega^{2m-2} + \ldots. + d'_{2m}}$$

The overall gain is then given by

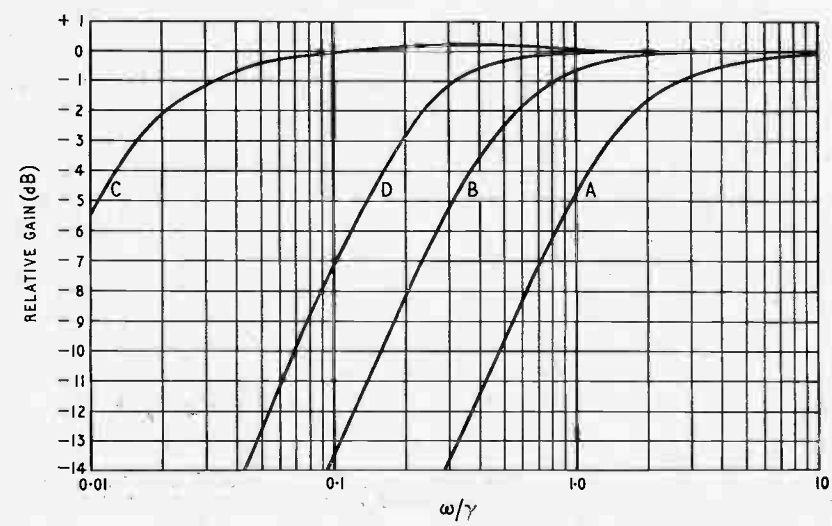
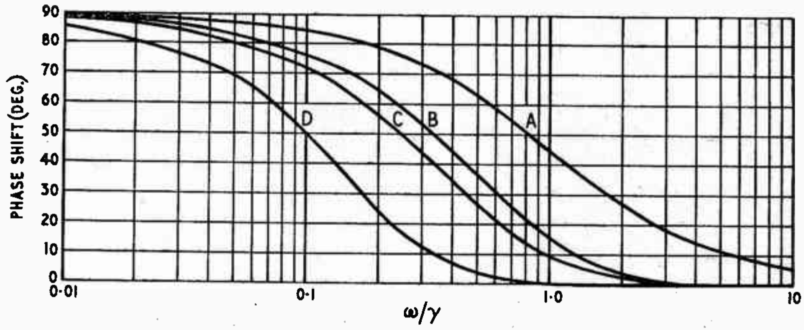


Fig. 3. *Gain-frequency response of amplifier stage: Curve A, coupling circuit alone* ($\alpha = \sqrt{2\gamma}$)*; B, coupling circuit and anode decoupling; C, circuit of Fig. 1, 3rd-order compensation; D, circuit of Fig. 1, 5th-order compensation*

Fig. 4. *Phase-frequency response of amplifier stage: Curve A, coupling circuit alone ($\alpha = \gamma$); B, coupling circuit and anode decoupling; C, circuit of Fig.* 1, *2nd-order compensation; D, circuit of Fig.* 1, *4th-order compensation*



$$\{G_1(\omega)\, G_2(\omega)\}^2 = \frac{\omega^{2(m+n)} + C_2\, \omega^{2(m+n-1)} + \ldots + C_{2(m+n)}}{\omega^{2(m+n)} + D_2\, \omega^{2(m+n-1)} + \ldots + D_{2(m+n)}}$$

where

$$C_{2r} = \sum_{s=0}^{r} c_{2s}\, c'_{2(r-s)}\ ,\ D_{2r} = \sum_{s=0}^{r} d_{2s}\, d'_{2(r-s)}$$

and $c_0 = c'_0 = d_0 = d'_0 = 1$

$$c_{2N} = d_{2N} = c'_{2M} = d'_{2M} = 0\ ,\ N > n\ ,\ M > m$$

If one stage has $(2r+1)$th-order compensation and the other has $(2q+1)$th-order, then

$$c_2 = d_2\ ,\ c_4 = d_4\ ,\ \ldots\ c_{2r} = d_{2r}\ ,\ c_{2r+2} \neq d_{2r+2}$$

and

$$c'_2 = d'_2\ ,\ c'_4 = d'_4\ ,\ \ldots\ c'_{2q} = d_{2q}\ ,\ c'_{2q+2} \neq d'_{2q+2}$$

so that if $r < q$

$$C_2 = D_2\ ,\ C_4 = D_4\ ,\ \ldots\ C_{2r} = D_{2r}$$

As for the indicial response, the order of compensation will be at least as great as that of the stage with the lower order of compensation. If, however, the first stage can have $(2r + 1)$th-order compensation, it is because the values of the first r of the coefficients c_{2r} and d_{2r} can be chosen. If the second stage can have $(2q + 1)$th-order compensation, then the values of the first q of the coefficients c'_{2q} and d'_{2q} can be chosen. It follows that the values of the first $(q + r)$ of the coefficients C_{2r} and D_{2r} can be chosen. If the Cs and Ds can be given any required values it is thus possible to obtain $C_0 = D_0$, $C_2 = D_2$, $\ldots$, $C_{2(q+r)+1} = D_{2(q+r)+1}$. The order of compensation of the overall response is then nearly the sum of the orders possible for the two separate stages. This condition cannot always be obtained in practice, however, due to restrictions on the range of values possible for the cs and ds.

Phase-Frequency Response

Let the phase-frequency response of two stages be

$$\tan \phi_1 = \frac{c_1 \omega^{2n-1} + c_3 \omega^{2n-3} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2 \omega^{2n-2} + \ldots + d_{2n}}$$

and

$$\tan \phi_2 = \frac{c'_1 \omega^{2m-1} + c'_3 \omega^{2m-3} + \ldots + c_{2m-1}\omega}{\omega^{2m} + d'_2 \omega^{2m-2} + \ldots + d'_{2m}}$$

The overall phase-frequency response is then

$$\tan(\phi_1 + \phi_2) = \frac{C_1 \omega^{2(m+n)-1} + C_3 \omega^{2(m+n)-3} + \ldots + C_{2(m+n)-1}\omega}{\omega^{2(m+n)} + D_2 \omega^{2(m+n-1)} + \ldots + D_{2(m+n)}}$$

where

$$C_{2r-1} = \sum_{s=1}^{r} \{c_{2s-1}\, d'_{2(r-s)} + c'_{2s-1}\, d_{2(r-s)}\}$$

and $D_{2r} = \sum_{s=1}^{r} \{d_{2s}\, d'_{2(r-s)} - c_{2s-1}\, c'_{2(r-s)+1}\}$

for $r = 1, 2, \ldots (m + n - 1)$,

$$D_0 = 1,\ D_{2(m+n)} = d_{2n}\, d'_{2m}$$

This will have $2r$th-order compensation if

$$C_{2s-1} = 0\ ,\ s = 1, 2, \ldots r$$

If one stage has $2r$th-order compensation and the other has $2q$th-order compensation then

$$c_1 = c_3 = \ldots = c_{2r-1} = 0,\ c_{2r+1} \neq 0$$

and

$$c'_1 = c'_3 = \ldots = c'_{2q-1} = 0,\ c_{2q+1} \neq 0$$

so that, if $r < q$

$$C_1 = C_3 = \ldots = C_{2r-1} = 0$$

The order of compensation of the two stages in tandem is therefore at least as great as that of the stage with the lower order of compensation. Since, however, the values of the first r coefficients c of one stage and the first q coefficients c' of the other stage can be chosen, it may be possible to adjust the values of the first $(q + r)$ of the coefficients C_{2r-1} so as to produce $2\,(q + r)$th-order compensation of the stages in tandem.

Negative Feedback Amplifiers

Amplifier with β Independent of Frequency

If the indicial response of the μ-path of the amplifier is

$$\mu(p)\mathbf{1} = \mu_0 \frac{p^n + a_2 p^{n-1} + \ldots + a_n}{p^n + b_1 p^{n-1} + \ldots + b_n}\,\mathbf{1} \quad \ldots \quad (35)$$

and the feedback fraction β_0, is independent of frequency, then the overall indicial response of the amplifier with feedback[1] is:

$$h(p)\mathbf{1} = \frac{\mu_0}{1 + \mu_0\beta_0} \cdot \frac{p^n + a_1 p^{n-1} + \ldots + a_n}{p^n + B_1 p^{n-1} + \ldots + B_n}\,\mathbf{1} \quad (36)$$

Where $B_r = (b_r + \mu_0\beta_0 a_r)/(1 + \mu_0\beta_0)$ .. (37)

The sign convention is chosen so that μ and β are both positive when the feedback is negative. If the μ-path has rth-order compensation of its indicial response then from equation (37) the overall response also has rth-order compensation. In particular, a maximally-flat overall response requires that the μ-path shall have a maximally-flat indicial response.

The gain-frequency response of the amplifier with feedback, from equations (5), (6), and (36) is given by

$$\{G(\omega)\}^2 = \left(\frac{\mu_0}{1 + \mu_0\beta_0}\right)^2 \left\{\frac{\omega^{2n} + c_2 \omega^{2n-2} + \ldots + c_{2n}}{\omega^{2n} + D_2 \omega^{2n-2} + \ldots + D_{2n}}\right\} \quad \ldots \quad (38)$$

where

$$\left.\begin{array}{l} c_{2r} = a_r^2 + 2\sum_{s=1}^{r} (-1)^s a_{r+s} a_{r-s} \\ \text{and } D_{2r} = B_r^2 + 2\sum_{s=1}^{r} (-1)^s B_{r+s} B_{r-s} \end{array}\right\} \quad .. \quad (39)$$

$(2r + 1)$th-order compensation requires

$$c_{2q} = D_{2q}\ ,\ q = 1, 2, \ldots . r.$$

This is obviously satisfied for

$$a_q = b_q\ ,\ q = 1, 2, \ldots . 2r.$$

Thus $2r$th-order compensation of the indicial response of the μ-path corresponds to $(2r + 1)$th-order compensation of the gain-frequency response of the amplifier. However, in general, $(2r + 1)$th-order compensation of the gain-frequency response of the μ-circuit does not correspond to a similar compensation for the overall response. Consequently a maximally-flat response of the μ-path does not result in a maximally-flat overall gain-frequency response.

The phase-frequency response of the amplifier with feedback, from equations (9), (10), and (36) is given by

$$\tan \phi = \frac{C_1\,\omega^{2n-1} + C_3\,\omega^{2n-3} + \ldots + C_{2n-1}\,\omega}{\omega^{2n} + D_2\,\omega^{2n-2} + \ldots . + D_{2n}} \quad .. \quad (40)$$

where

$$\left.\begin{array}{l} C_{2r-1} = (-1)^r \sum_{s=1}^{r} \left\{a_{2s-1} B_{2(r-s)} - a_{2(r-s)} B_{2s-1}\right\} \\ \qquad = c_{2r-1}/(1+\mu_0\beta_0) \\ \text{and } D_{2r} = (-1)^r \left\{B_{2r} + \sum_{s=1}^{r} [a_{2s} B_{2(r-s)} + a_{2s-1} B_{2(r-s)+1}]\right\} \end{array}\right\} \quad .. \quad (41)$$

$2r$th-order compensation requires

$$C_{2q-1} = 0\ ,\ q = 1, 2, \ldots . r$$

This is satisfied if $a_q = b_q\ , q = 1, 2, \ldots 2r-1$

i.e., $(2r-1)$th-order compensation of the indicial response of the μ-path corresponds to $2r$th-order compensation of the overall phase-frequency response.

More generally, since $C_r = 0$ when $c_r = 0$, $2r$th-order compensation of the phase-frequency response of the μ-path corresponds to a similar compensation of the overall phase-frequency response. In particular, a maximally-flat overall response requires that the μ-path shall have a maximally-flat phase-frequency response.

Effect of Networks Outside the Feedback Loop

If the amplifier contains a coupling circuit comprising a capacitance and resistance with time-constant $1/\alpha$, the indicial response of the coupling is $p/(p+\alpha)\mathbf{1}$. If the indicial response of the amplifier is given by equation (36) and the coupling circuit is outside the feedback loop, the overall indicial response is

$$\begin{aligned} h(p)\mathbf{1} &= \frac{p}{p+\alpha} \cdot \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^n + a_1 p^{n-1} + \ldots + a_n}{p^n + B_1 p^{n-1} + \ldots + B_n}\mathbf{1} \\ &= \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^{n+1} + a_1 p^n + \ldots . + a_n p}{p^{n+1} + B'_1 p^n + \ldots . + B'_{n+1}}\mathbf{1} \end{aligned} \quad .. \quad (42)$$

$$\begin{aligned} \text{where } B'_r &= B_r + \alpha B_{r-1} \\ &= [b_r + \alpha b_{r-1} + \mu_0\beta_0 (a_r + \alpha a_{r-1})]/[1+\mu_0\beta_0] \end{aligned} \quad .. \quad (43)$$

The overall gain-frequency response, from equations (5), (6), (42) and (43) is then

$$\{G(\omega)\}^2 = \left(\frac{\mu_0}{1+\mu_0\beta_0}\right)^2 \frac{\omega^{2n+2} + c_2\,\omega^{2n} + \ldots . + c_{2n}\omega^2}{\omega^{2n+2} + D_2\,\omega^{2n} + \ldots . + D_{2n+2}} \quad .. \quad (44)$$

$$\left.\begin{array}{l} \text{where } c_{2r} = a_r^2 + 2\sum_{s=1}^{r} (-1)^s a_{r+s} a_{r-s} \\ \text{and } D_{2r} = B'^2_r + 2\sum_{s=1}^{r} (-1)^s B'_{r+s} B'_{r-s} \end{array}\right\} \quad (45)$$

The overall phase-frequency response, from equations (9), (10), (42) and (43) is

$$\tan \phi(\omega) = \frac{C_1\,\omega^{2n+1} + C_3\,\omega^{2n+1} + \ldots . + C_{2n+1}\,\omega}{\omega^{2n+2} + D_2\,\omega^{2n} + \ldots . + D_{2n+2}} \quad .. \quad (46)$$

$$\left.\begin{array}{l} \text{where } C_{2r-1} = (-1)^r \sum_{s=1}^{r} \left\{a_{2s-1} B'_{(2r-s)} - a_{2(r-s)} B'_{2s-1}\right\} \\ \text{and } D_{2r} = (-1)^r \\ \qquad \left\{B'_{2r} + \sum_{s=1}^{r} (a_{2s} B'_{2(r-s)} - a_{2s-1} B'_{2(r-s)+1})\right\} \end{array}\right\} \quad (47)$$

In order to compare the effect of adding a coupling circuit outside the feedback loop with a coupling circuit inside the loop, consider the effect of adding a coupling circuit with time-constant $1/\alpha'$ to the amplifier whose indicial response is given by equation (35). The indicial response of the μ-path, including the coupling circuit, is

$$\mu(p)\,\mathbf{1} = \mu_0 \frac{p^{n+1} + a_1 p^n + \ldots . + a_n p}{p^{n+1} + b'_1 p^n + \ldots . + b'_{n+1}} \quad .. \quad (48)$$

$$\text{where } b'_r = b_r + \alpha' b_{r-1} \quad .. \quad (49)$$

The overall indicial response is therefore given by equation (42) where, from equations (37) and (49)

$$B'_r = [b_r + \alpha' b_{r1} + \mu_0\beta_0 a_r]/[1+\mu_0\beta_0] \quad .. \quad (50)$$

The overall gain-frequency response is given by equations (44) and (45) where the value of B'_r is that given by equation (50).

The difference in form between equations (43) and (50) shows that, in general, the addition of a coupling circuit has a different effect on each of the overall responses depending on whether it is inside or outside the feedback loop.

If the amplifier whose overall indicial response is given by equation (36) has an additional stage with anode decoupling outside the feedback loop, the indicial response of the additional stage, from equation (17) is

$$\frac{p + (\gamma+\lambda)}{p+\lambda}\mathbf{1}$$

The overall indicial response, including the additional stage, from equations (36) and (17) is

$$h(p)\,\mathbf{1} = \frac{\mu_0}{1+\mu_0\beta_0} \cdot \frac{p^{n+1} + A_1 p^n + \ldots . + A_{n+1}}{p^{n+1} + B'_1 p^n + \ldots . + B'_{n+1}}\,\mathbf{1} \quad .. \quad (51)$$

$$\begin{aligned} \text{where } A_r &= a_r + (\gamma+\lambda)\,a_{r-1} \\ \text{and } B'_r &= B_r + \lambda B_{r-1} \\ &= [b_r + \mu_0\beta_0 a_r + \lambda(b_{r-1} + \mu_0\beta_0 a_{r-1})]/[1+\mu_0\beta_0] \end{aligned} \quad (52)$$

The overall gain-frequency response from equations (5), (6) and (51) is given by

$$\left\{G(\omega)\right\}^2 = \left(\frac{\mu_0}{1+\mu_0\beta_0}\right)^2 \frac{\omega^{2n+2} + C_2\omega^{2n} + \ldots . + C_{2n+2}}{\omega^{2n+2} + D_2\omega^{2n} + \ldots . + D_{2n+2}} \quad .. \quad (53)$$

$$\left.\begin{aligned} \text{where } C_{2r} &= A_r^2 + 2\sum_{s=1}^{r}(-1)^s A_{r+s}A_{r-s} \\ \text{and } D_{2r} &= B'^2_r + 2\sum_{s=1}^{r}(-1)^s B'_{r+s}B'_{r-s} \end{aligned}\right\} \quad \cdots \quad (54)$$

The overall phase-frequency response, from equations (9), (10) and (51) is

$$\tan\phi(\omega) = \frac{C_1\,\omega^{2n+1} + C_3\,\omega^{2n-1} + \ldots + C_{2n+1}\,\omega}{\omega^{2n+2} + D_2\,\omega^{2n} + \ldots + D_{2n+2}} \quad (46)$$

$$\left.\begin{aligned} \text{where } C_{2r-1} &= (-1)^r \sum_{s=1}^{r}\left(A_{2s-1}B'_{2(r-s)} - A_{2(r-s)}B'_{2s-1}\right) \\ \text{and } D_{2r} &= (-1)^r \\ &\left\{B'_{2r} + \sum_{s=1}^{r}\left(A_{2s}B'_{2(r-s)} - A_{2s-1}B'_{2(r-s)+1}\right)\right\} \end{aligned}\right| \quad \cdots \quad (55)$$

In order to compare the effect of adding a decoupling circuit outside the feedback loop with adding a decoupling circuit inside the loop, consider the effect of adding a decoupling circuit with time-constants $1/\gamma'$ and $1/\lambda'$ to the amplifier whose indicial response is given by equation (35). The indicial response of the μ-path including the decoupling circuit, from equations (17) and (35) is

$$\mu(p)\,\mathbf{1} = \mu_0 \frac{p^{n+1} + A_1 p^n + \ldots + A_{n+1}}{p^{n+1} + B_1 p^n + \ldots + B_{n+1}}\mathbf{1} \quad (56)$$

$$\left.\begin{aligned} \text{where } A_r &= a_r + (\lambda' + \gamma')\,a_{r-1} \\ B_r &= b_r + \lambda'\,b_{r-1} \end{aligned}\right\} \quad \cdots \quad \cdots \quad (57)$$

The overall indicial response from equations (56) and (36) is therefore given by equation (51)

$$\begin{aligned} \text{where } B'_r &= [B_r + \mu_0\beta_0 A_r]/[1 + \mu_0\beta_0] \\ &= [b_r + \mu_0\beta_0\,a_r + \lambda'(b_{r-1} + \mu_0\beta_0\,a_{r-1}) \\ &\quad + \mu_0\beta_0\,\gamma' a_{r-1}]/[1 + \mu_0\beta_0] \quad \cdots \quad (58) \end{aligned}$$

The gain-frequency response from equations (5), (6) and (58) is given by equations (53) and (54). Where A_r and B'_r are given by equations (57) and (58) The phase-frequency response, from equations (9), (10) and (58), is given by equations (46) and (55) where A_r and B'_r are again given by equations (57) and (58). By comparing equations (52) and (58) it is seen that the condition for an amplifier with a decoupling circuit outside the feedback loop to have rth-order compensation of indicial response is the same as that for an amplifier with an additional decoupling circuit inside the loop having $\lambda' = \lambda$ and $\gamma' = \gamma\,(1 + \mu_0\beta_0)$. However, when the equations for the gain-frequency responses are compared, it is seen that they are not equivalent for any values of $\gamma, \lambda, \gamma', \lambda'$ except the trivial case $\lambda = \lambda'$; $\gamma = \gamma' = 0$. When the equations for the phase-frequency responses are compared, it is seen that they are not equivalent for any values of $\gamma, \lambda, \gamma', \lambda'$. It is therefore possible to regard the effect of a decoupling circuit outside the feedback loop as equivalent to one inside the loop only when considering the indicial response.

Amplifier with a Time-Constant in the Feedback Path

If a capacitor is connected in series with the resistance potential divider which comprises the β-path to give a time-constant $1/\sigma$, the indicial response of the feedback path becomes

$$\beta(p)\,\mathbf{1} = \beta_0 \frac{p}{p + \sigma}\mathbf{1}$$

The overall indicial response is then

$$\begin{aligned} h(p)\,\mathbf{1} &= \frac{\mu(p)}{1 + \beta(p)\mu(p)}\mathbf{1} \\ &= \frac{\mu_0}{1 + \mu_0\beta_0} \cdot \frac{p^{n+1} + a_1' p^n + \ldots + a'_{n+1}}{p^{n+1} + B_1' p^n + \ldots + B'_{n+1}}\mathbf{1} \quad (59) \end{aligned}$$

where

$$\left.\begin{aligned} a'_r &= a_r + a_{r-1}\sigma \\ B'_r &= [b_r + \sigma\,b_{r-1} + \mu_0\beta_0 a_r]/[1 + \mu_0\beta_0] \end{aligned}\right\} \quad (60)$$

The gain-frequency response is then given by equations (53) and (54) where $A_r = a'_r$ and a'_r and B'_r are given by equations (60). The phase-frequency response is given by equations (46) and (55) where $A_r = a'_r$ and a'_r and B'_r are given by equations (60). By comparing equations (58) with equations (60) it is seen that the conditions for rth-order compensation of the indicial response when a time-constant is added to the β-path are equivalent to the conditions when a stage of anode decoupling is added to the μ-path with values of R_aC_d and R_aC_d given by $1/\sigma$ and $1/\mu_0\beta_0\sigma$ respectively. Alternatively, by comparing equations (60) with equations (52) it follows that the effect on the order of the indicial response obtained by adding the time-constant to the β-path is equivalent to adding a decoupling circuit outside the feedback loop with values of R_dC_d and R_aC_d given by $1/\sigma$ and $(1 + \mu_0\beta_0)/\mu_0\beta_0\sigma$ respectively. However, when the equations for the gain-frequency responses are compared, it is seen that they are not equivalent for any component values of decoupling circuit either inside the feedback loop or outside. When the equations for the phase-frequency responses are compared it is also seen that they are not equivalent for any component values of a decoupling circuit, either inside or outside the feedback loop.

Conclusions

General expressions for the gain-frequency and phase-frequency response of amplifiers at low frequencies have been deduced from the corresponding indicial response. The general expressions can be used to obtain the conditions for maximal flatness of the indicial response or of the gain-frequency response or of the phase-frequency response of particular amplifier circuits; these conditions are generally different. A single resistance-capacitance coupled stage can have second-order compensation of its indicial response, up to fifth-order compensation of its gain-frequency response or up to fourth-order compensation of its phase-frequency response.

As the number of stages in an amplifier is increased, the order of compensation which can be obtained increases, but so does the complexity of the calculations required to obtain the conditions for compensation. However, if two stages are connected in tandem having qth- and rth-order compensation of indicial response or gain-frequency response or phase-frequency response, the overall response will be of order q or r, whichever is the lower. A complicated circuit can therefore be dealt with by considering it as a number of simpler stages connected in tandem, each of which can be made to have the required order of compensation of its indicial or gain-frequency or phase-frequency response.

When an amplifier has negative feedback applied with a feedback fraction β which is independent of frequency, the overall indicial response or phase-frequency response is maximally flat when the corresponding response of the μ-path is maximally flat. A maximally-flat gain-frequency response, however, is not obtained when the μ-path has a maximally-flat gain-frequency response.

Some simple relationships have been found between the conditions for maximal flatness of the indicial response when anode decoupling circuits are added to feedback amplifiers either inside or outside the feedback loop or a time-constant is added to the β-path. No such relationships hold for the gain-frequency response or the phase-frequency response.

Acknowledgement

Acknowledgement is made to Siemens Edison Swan Limited for permission to publish this paper.

REFERENCES

1 J. E. Flood, "L.F. Compensation for Video Amplifiers", *Wireless Engineer*, Vol. 31, pp. 175–186 and 214–221.
2 G. W. Edwards and E. C. Cherry, "Amplifier Characteristics with Particular Reference to a New Method of Frequency Compensation of Single Stages", *J. Instn elect. Engrs*, 1940, Vol. 87, p. 178.
3 H. M. Zeidler and J. D. Noe, "Pentriode Amplifiers", *Proc. Inst. Radio Engrs*, 1948, Vol. 36, p. 1332.
4 W. E. Thomson, "Transient Response of V.F. Couplings", *Wireless Engineer*, 1947, Vol. 24, p. 20.
5 H. L. Donley and D. W. Epstein, "Low-Frequency Characteristics of the Coupling Circuits of Single- and Multi-stage Video Amplifiers", *R.C.A. Rev.*, 1942, Vol. 3, p. 416.
6 W. A. Lynch, "Video Amplifier L-F Correction", *Communications*, 1943, Vol. 23, p. 16.

MATHEMATICAL TOOLS

By Computer

Matrices 2: The Laws of Matrix Algebra, and the 'A'-Matrices of Standard Networks

In last month's article, we considered the general purpose and significance of a matrix (with two rows and two columns) in relation to the performance of a four-terminal network, with due regard to certain important special cases, such as that of an ideal transformer. We now have to consider how the matrices of complicated circuits can be derived from those of their component parts ; this is equivalent to saying that we require an algebra for the handling of matrices.

Given any four-terminal network, there are several different matrices which can be regarded as equally valid for describing its essential performance. We shall discuss these various matrices, and the relations between

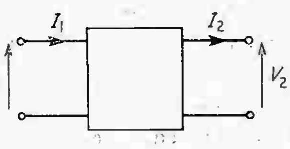


Fig. 1. *A linear four-terminal network*

them, in our next article. For the time being, we shall confine our attention to one particular kind of matrix, usually called the A-matrix, which arises when the input voltage and current are expressed in terms of the output voltage and current, and we shall derive the A-matrices of certain fundamental networks from which we can build up all others; this building-up process will be made clear when we derive the A-matrices of a general T-network and of a general Π-network. The most useful property of the A-matrix is that if two networks are connected in cascade, the A-matrix of the combined network is the product of the A-matrices of the separate networks. This useful property, however, is useless until we can give a sensible meaning to the word 'product' as applied to matrices.

Suppose then that we have a four-terminal network N_1 where the input current I_1 and voltage V_1 are given in terms of the output current I_2 and voltage V_2 by the linear relations

$$V_1 = a_{11} V_2 + a_{12} I_2$$
$$I_1 = a_{21} V_2 + a_{22} I_2$$
$$\text{or} \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad \ldots \quad \ldots \quad (1)$$

with the sign convention of Fig. 1, that is to say, V_1 is positive if the upper input terminal is at a higher potential than the lower, V_2 is positive if the upper output terminal is at a higher potential than the lower, and the currents I_1 and I_2 are positive if in the direction of the arrows. We can, for simplicity, regard the V and I terms as sinusoidal quantities so that the familiar j-notation can be used. The coefficients a_{11} and a_{22} are then numbers which may be complex, while a_{12} has the essential nature of an impedance and a_{21} that of an admittance. Equation (1) can be extended to the case of arbitrary currents and voltages by means of operational calculus; it is then an equation in the "p-world" which has been frequently mentioned in earlier articles. The terms a_{11}, a_{12}, a_{21} and a_{22} are then functions of p, and a_{12} is still an 'impedance' in the generalized p-world sense, while a_{21} is similarly a generalized admittance. Now suppose that, for a second four-terminal network, the input current I_3 and voltage V_3 are related to the output current I_4 and voltage V_4 by the equations

$$\left.\begin{aligned} V_3 &= b_{11} V_4 + b_{12} I_4 \\ I_3 &= b_{21} V_4 + b_{22} I_4 \end{aligned}\right\}$$

$$\text{or} \quad \begin{pmatrix} V_3 \\ I_3 \end{pmatrix} = \begin{pmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{pmatrix} \begin{pmatrix} V_4 \\ I_4 \end{pmatrix} \quad \ldots \quad \ldots \quad (2)$$

Then if these two networks are connected in cascade, so that

$$V_2 = V_3; \quad I_2 = I_3 \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

we can obtain V_1 and I_1 in terms of V_4 and I_4 by direct

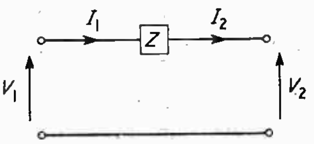


Fig. 2. *Single series element*

substitution from (2) into (1) in view of (3). The result reduces to

$$\begin{aligned} V_1 &= c_{11} V_4 + c_{12} I_4 \\ I_1 &= c_{21} V_4 + c_{22} I_4 \end{aligned}$$

$$\text{or} \quad \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} c_{11} & c_{12} \\ c_{21} & c_{22} \end{pmatrix} \begin{pmatrix} V_4 \\ I_4 \end{pmatrix} \quad \ldots \quad \ldots \quad (4)$$

where

$$\left.\begin{aligned} c_{11} &= a_{11} b_{11} + a_{12} b_{21}\,; \quad c_{12} = a_{11} b_{12} + a_{12} b_{22} \\ c_{21} &= a_{21} b_{11} + a_{22} b_{21}\,; \quad c_{22} = a_{21} b_{12} + a_{22} b_{22} \end{aligned}\right\} \quad (5)$$

This suggests that it would be useful to define the 'product' of two matrices

$$A = \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} \quad \text{and} \quad B = \begin{pmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{pmatrix} \quad \ldots \quad (6)$$

as

$$AB = C = \begin{pmatrix} c_{11} & c_{12} \\ c_{21} & c_{22} \end{pmatrix} \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

where the c-elements are given in terms of the a-elements and b-elements by (5). Now it so happens that mathematicians have somewhat arbitrarily chosen a definition of a 'matrix product' which reduces to (5) for a square matrix having two rows and two columns, because they have found that with this definition, a satisfactory matrix algebra can be built up.

A further examination of (1) shows that if we are given (1) in the matrix form and wish to derive the equation form, we can do so by carrying out matrix multiplication in a manner analogous to the rule (5) already adopted for square matrices.

The remaining definitions are reasonably straightforward, namely

(a) There are three kinds of matrices having not more than two rows and columns, namely row matrices (r_1, r_2) having one row and two columns, column matrices, like (V_1/I_1), (V_2/I_2), etc., having two rows and one column, and square matrices like A and B in (6) having two rows and two columns.

(b) Two matrices are equal only if they are both row matrices, both column matrices or both square matrices, and corresponding elements are equal.

(c) Two matrices can be added (or subtracted) only if they are of the same kind (i.e., both row, both column or both square) and each element of the sum (or difference) matrix is obtained by adding (or subtracting) corresponding elements of the two original matrices.

(d) A matrix all of whose elements are zero is called a null matrix, and the square matrix A given by (6) where $a_{11} = a_{22} = 1$ and $a_{12} = a_{21} = 0$ is called the unit matrix and often denoted by I.

(e) The reciprocal of a square matrix A is another square matrix A^{-1} such that

$$AA^{-1} = A^{-1} A = I \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

Note that when a matrix product is formed by means of (5), it makes a difference whether A precedes B, as in (5), or B precedes A but, in the exceptional case of the reciprocal matrix, the order of the matrices A and A^{-1} is immaterial.

(f) To find the reciprocal of A, we need to know its determinant $|A|$ defined by

$$|A| = a_{11} a_{22} - a_{12} a_{21} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

It will be noticed that the determinants of all the A-matrices of specific networks discussed below are unity. In fact, the determinant of the A-matrix of any passive network which obeys the law of reciprocity is unity. If two matrices A, B are such that the product AB is a third matrix C, it can be shown that the determinant $|C|$ is the product of the determinants $|A|$ and $|B|$. In terms of the determinant (9), if A is given by (6), then we have

$$A^{-1} = \begin{pmatrix} a_{22}/|A| & -a_{12}/|A| \\ -a_{21}/|A| & a_{11}/|A| \end{pmatrix} \quad \ldots \quad \ldots \quad (10)^*$$

(g) Division by a square matrix A is equivalent to multiplication by its reciprocal A^{-1}.

(h) For ladder networks and similar structures, we may need to multiply not two but any number of A-matrices. A neat way of doing this was recently given by W. Proctor Wilson† but this is outside the scope of these articles.

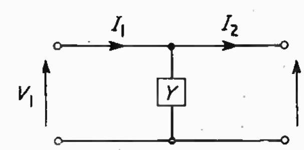


Fig. 3. *Single shunt element*

(i) With the above definitions, matrices can be regarded as obeying all the laws of algebra, except that in a multiplication, the order of the factors must not be changed.

Now consider the three fundamental networks from which we can build up all others, namely an ideal transformer of ratio $n:1$, a network having a single series element as in Fig. 2 and a network having a

* For an A-matrix, $|A|$ is usually 1, but this is not true for other matrices discussed later. (10), however, is true for any square matrix having two rows and columns provided that the determinant $|A|$ of the matrix is different from zero.

† *Electronic and Radio Engineer*, June 1957, pp. 229–231.

single shunt element as in Fig. 3. For the ideal transformer of ratio $n:1$ we have

$$\begin{aligned} V_1 &= nV_2 \\ I_1 &= (1/n)I_2 \end{aligned} \quad \text{or} \quad \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} n & 0 \\ 0 & 1/n \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \qquad (11)$$

and the determinant of the matrix (11) is clearly unity, from (9). For the single series element of Fig. 2, with impedance Z, we have

$$\begin{aligned} V_1 &= V_2 + ZI_2 \\ I_1 &= I_2 \end{aligned} \quad \text{or} \quad \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \qquad (12)$$

so that the determinant of the matrix (12) is also unity. Again, for the single shunt element of Fig. 3, with admittance Y, we have

$$\begin{aligned} V_1 &= V_2 \\ I_1 &= YV_1 + I_2 \end{aligned} \quad \text{or} \quad \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \qquad (13)$$

and once more the determinant of the matrix (13) is unity.

Now consider the network of Fig. 4(a), which can

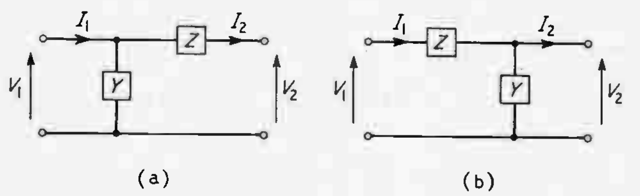


Fig. 4. *L-networks in which a single shunt element (a) precedes and (b) follows a single series element*

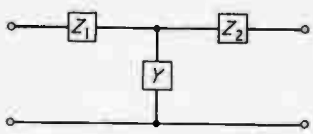


Fig. 5. *A general T-network*

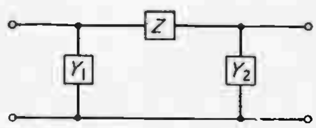


Fig. 6. *A general π-network*

be regarded as that of Fig. 3 followed in cascade by that of Fig. 2. The A-matrix of the combination is therefore obtained directly by multiplying the matrices of (13) and (12) in that order; it is thus

$$\begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} = \begin{pmatrix} 1 & Z \\ Y & 1+YZ \end{pmatrix} \quad \cdots \quad (14)$$

and the correctness of this result can easily be verified by direct analysis of Fig. 4(a); once the product-rule (5) for square matrices is mastered, however, direct derivation of (14) by matrix multiplication is easier. As the determinants of the two A-matrices on the left-hand side of (14) are both 1, we should expect the determinant of the A-matrix on the right-hand side to be 1×1, that is, 1 also; this is easily verified from (9).

Similarly, for the network of Fig. 4(b), the A-matrix is

$$\begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} = \begin{pmatrix} 1+YZ & Z \\ Y & 1 \end{pmatrix} \quad \cdots \quad (15)$$

for the network of Fig. 4(b) can be regarded as that of Fig. 2 followed in cascade by that of Fig. 3.

We have now only one further stage to derive the A-matrix of the general T-network of Fig. 5; this can be regarded as the network of Fig. 4(b) [with Z_1 for Z] followed in cascade by that of Fig. 2 [with Z_2 for Z]. The A-matrix is therefore

$$\begin{pmatrix} 1+YZ_1 & Z_1 \\ Y & 1 \end{pmatrix} \begin{pmatrix} 1 & Z_2 \\ 0 & 1 \end{pmatrix}$$

$$= \begin{pmatrix} 1+YZ_1 & Z_1+Z_2+YZ_1Z_2 \\ Y & 1+YZ_2 \end{pmatrix} \quad \cdots \quad \cdots \quad (16)$$

and again it has determinant unity. The corresponding result for the A-matrix of the general Π-network of Fig. 6 is

$$\begin{pmatrix} 1 & Z \\ Y_1 & 1+Y_1Z \end{pmatrix} \begin{pmatrix} 1 & 0 \\ Y_2 & 1 \end{pmatrix}$$

$$= \begin{pmatrix} 1+Y_2Z & Z \\ Y_1+Y_2+ZY_1Y_2 & 1+Y_1Z \end{pmatrix} \quad \cdots \quad \cdots \quad (17)$$

since this network can be regarded as that of Fig. 4(a) with Y_1 for Y followed by that of Fig. 3 with Y_2 for Y. The same result is obtained by regarding it as the network of Fig. 3 with Y_1 for Y followed by that of Fig. 4(b) with Y_2 for Y.

It is perhaps worth passing note that if we obtain the A-matrix of any network, we can immediately derive the elements of the equivalent-T from (16) or those of the equivalent-Π from (17). For if the A-matrix is given by A in (6), and the network is passive and reciprocal so that $|A| = 1$ [this last condition must be satisfied if it has an equivalent-T or equivalent-Π circuit] then the equivalent-T is as in Fig. 5 where

$$Y = a_{21};\ Z_1 = (a_{11}-1)/a_{21};\ Z_2 = (a_{22}-1)/a_{21} \qquad (18)$$

and the equivalent-Π is as in Fig. 6 where

$$Z = a_{12};\ Y_1 = (a_{22}-1)/a_{12};\ Y_2 = (a_{11}-1)/a_{12} \qquad (19)$$

and these equivalent networks will be symmetrical if and only if $a_{11} = a_{22}$.

It is also worth passing note that the star-delta transformation can in effect be carried out in either direction by means of (18) and (19). Given Y, Z_1 and Z_2, if we require Z, Y_1 and Y_2, we can first obtain a_{21} from the first of (18), then a_{11} and a_{22} from the remainder of (18).

a_{12} is then derived from the fact that the network is passive and reciprocal, so that the determinant (9) of the A-matrix is unity. We can now determine Z, Y_1 and Y_2 direct from (19). In the reverse case where Z, Y_1 and Y_2 are given and we require Y, Z_1 and Z_2, (19) gives a_{12}, a_{11} and a_{22}, (9) with $|A| = 1$ gives a_{21} and then (18) gives Y, Z_1 and Z_2.

Thus we have been able to derive the relations between input and output current and voltage for the general T-network of Fig. 5 and the general Π-network of Fig. 6 by algebraic manipulation of similar relations applied to simpler networks contained within them. This process could be greatly extended. So far, however, we have not used the simpler law for adding matrices. In our next article we shall consider other matrices such as the Z-matrix by means of which the input and output voltages are expressed in terms of the input and output currents. If two networks are in series, we shall show that the Z-matrix of the combination is the sum of the Z-matrices of the separate networks. Thus, further ways become available for expressing the behaviour of a complicated network in terms of the behaviour of its component parts.

Transmission-Line Low-Pass Filters

DESIGN METHODS FOR THE V.H.F. AND U.H.F. BANDS

By F. Charman, B.E.M.*

SUMMARY. *This article describes the design of low-pass filters in the v.h.f. range and the appendices give the mathematical analysis of the design work.*

In the frequency range of, say, 100–1000 Mc/s, it is not possible to build networks using the lumped reactance elements which are normally used at lower frequencies. The simple 'lumped' elements cease to be lumped, and become distributed structures; their reactance–frequency curves change and they eventually exhibit resonances. Inductances must, in general, be abandoned first, though with care it is possible to build capacitance elements which are reasonably 'true to type' even at a frequency of several hundred Mc/s.

It is therefore necessary to find a new approach to network design in this range; networks must be designed in terms of distributed structures, and new mathematical methods must be introduced. The most general form of distributed structure is the transmission line, and this forms a convenient unit element both in theory and in practice. It is possible, up to a point, to introduce capacitances into the structures and this can be a great advantage although at frequencies above, say, 300 Mc/s these may have tangible self-inductance. At higher frequencies, line junctions themselves cause fresh difficulties, to which no practicable solution has yet been found.

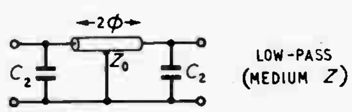
The first approach to transmission-line filters was made by Mason and Sykes in 1937, since when progress has been made,[2,3] but there is need for suitable classification and workable design formulæ, and there is much to be learned about practical construction.

This article deals with one of the simplest forms of transmission-line filter, and describes some experimental attempts to align theory and practice. The filter section comprises transmission-line series arms with capacitance shunt arms. It has been found that useful low-pass filters can be made in this way. Design procedure has been developed and several filters have been designed and made, showing progressive improvements in technique. These filters are described in detail and the complete theory is given in the Appendices.

* *E.M.I. Electronics, Ltd., Hayes, Middlesex.*

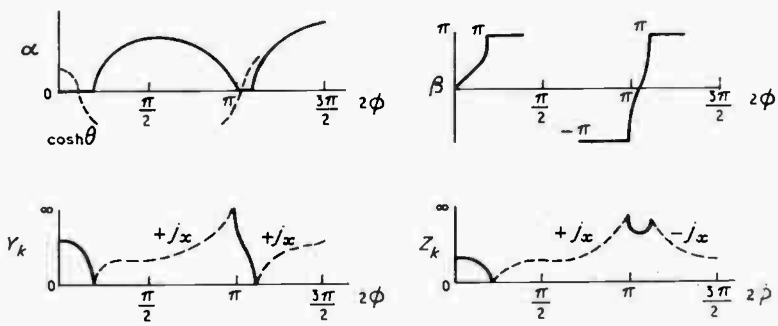
Transmission-Line Filters

The literature gives many examples of attempts to take the well-known low-frequency networks and to copy them in transmission-line structures; these attempts are, however, mostly failures. The reason is that the distributed structures are higher order networks with more variables; generalization should, if at all, proceed in the other direction. Certainly we can still use conventional network theory to cascade sections, but the terminal properties of the sections themselves require new methods for their solution.



$$\cosh\theta = \cos 2\phi - a\sin 2\phi$$
$$(Y_k/Z_0)^2 = 2a\cot 2\phi + 1 - a^2$$
$$(Z_k/Z_0)^2 = (1 - a\tan\phi)/(1 + a\cot\phi)$$

Notes. Y_k = mid-shunt $\quad Z_k$ = mid-series $\quad a = \omega C_2 Z_0$



Design Data

Let a $(= \omega C_2 Z_0) = 1$ at ω_0. Then $a = \omega/\omega_0 = \phi/\phi_0 = b\,.\,2\phi$; $(b = 1/2\phi_0$ and ω_0, ϕ_0 are purely auxiliary parameters.)
Then $\cosh\theta = 0$ when $\cot 2\phi/2\phi = b = \pm 1$ when $\cot\phi/\phi = 2b$ or $\tan\phi/\phi = -2b$ = a maximum when $\tan 2\phi/2\phi = -b/(1+b)$ and is of value $\cos 2\phi[1 + a^2/(1+b)]$.
$Y_k = Y_0$ (mid-shunt only) when $\cot 2\phi/2\phi = b/2$. Never in mid-series. (<1 in bands 1, 3, etc., $\gg 1$ in bands 2, 4, etc.)
Parameter $b = 1/2\phi_0$ can be varied to control either Y_k in pass-band or values of maximum attenuation:
See curves:
$Y_k Z_0$ as $f(b)$ and cut-off (Fig. 3) $\quad Y_k Z_0 \,.\, Z_k/Z_0 \,.\, \alpha_{max}$ as f (cut-off angle) (Fig. 2)

Fig. 1. *Design data for low-pass filters*

In lumped element circuits, the reactances or susceptances are linear functions of frequency ; e.g., $Z = j\omega L$ or $Y = j\omega C$. In distributed circuits they take forms like $Z = jZ_0 \tan\phi$ and the solutions come in terms of circular functions or transcendental functions such as $\tan\phi/\phi$ but, in addition, time is consumed in passing through a section of transmission line in a way which is not very evident in, for example, a low-pass lumped filter, and this is a pertinent reason why the new methods of analysis are needed.

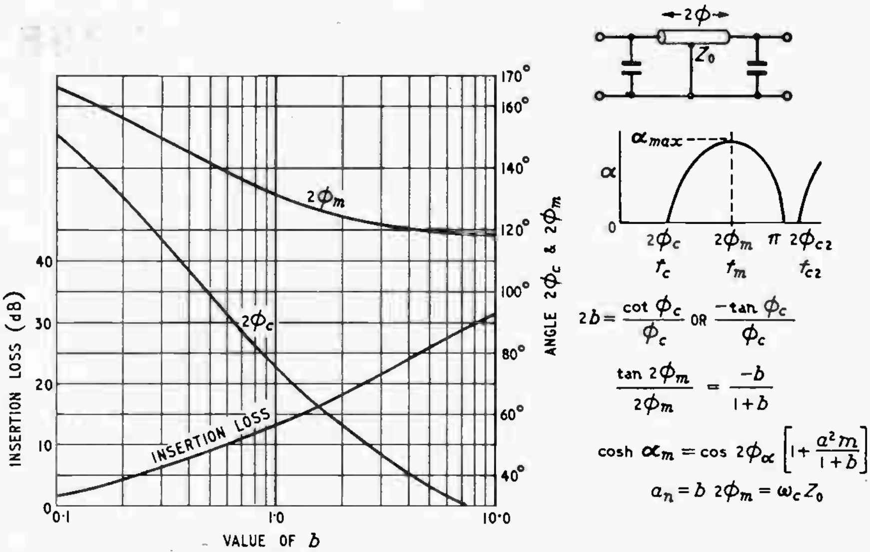


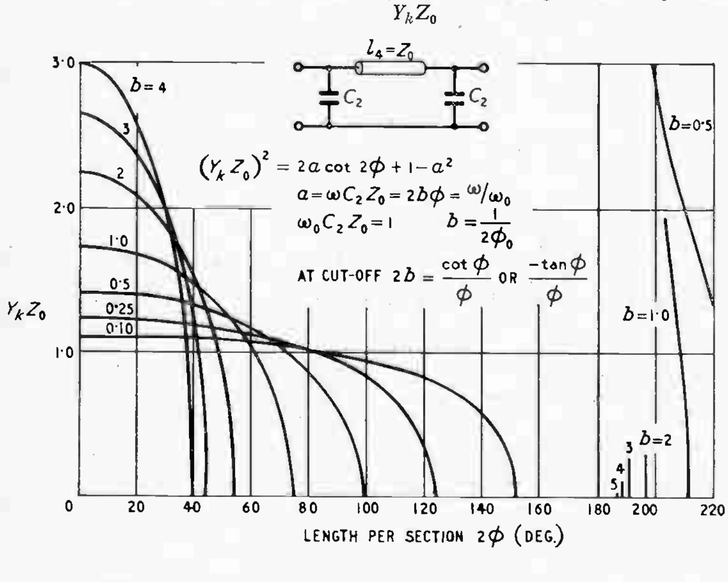
Fig. 2. *Low-pass filter characteristics*

Since the properties of the networks depend upon periodic instead of linear functions, the solutions are always multi-valued, [$\tan x = \tan (x \pm n\pi)$ and so forth]. Therefore, the simplest filters exhibit a series of pass-bands and, in order to use the filters, we may have to find a way to cover up all but one of these bands. In many cases, these bands are all of the same frequency width but, if some lumped reactance can be introduced, the higher order bands can be made to close very rapidly and become unimportant. The use of fixed capacitances as far up the frequency scale as possible should therefore be tried. The filters described here comprise lengths of line with capacitance shunts at regular intervals. They represent one of the simplest forms of T-L Filter and, since it appears that the higher order pass-bands can be made evanescent, they have been referred to as Low-Pass Filters.

Design Procedure

The complete analysis of this filter section is given in Appendix 1, and from it the data sheet (Fig. 1) and the design curves Figs. 2, 3 and 4 have been prepared. It is found that the properties of the network can be defined completely by two variables, one of which is the characteristic impedance Z_0 of the line and the other its electrical length at the first cut-off frequency.

Fig. 3. *Influence of electrical length of line on image admittance function* $Y_k Z_0$



The latter variable is introduced in terms of a basic frequency ω_0 or corresponding electrical line length 2ϕ at which the capacitive reactance is equal to Z_0. This gives a coefficient $b = 1/2\phi_0$ as the main controlling parameter; the line-characteristic impedance Z_0 is now only an image impedance scale factor.

The effect of the capacitive susceptance is introduced into the normalized equations by means of a factor $a = \omega C Z_0$, and this factor is obviously unity at the frequency ω_0, thus $a = \omega/\omega_0$ and is a normalized frequency variable. When b is increased, the line becomes shorter and the capacitance greater, whereas values of b much less than unity tend to degenerate the filter into a multiple band-stop filter with low attenuation in the stop bands.

In useful cases, therefore, the parameter b will be greater than unity and 2ϕ will always be less than one radian at cut off. Study of the design curves (Figs. 1–4) shows the way in which the value of b affects the filter properties. It controls the fractional ratio between cut-off frequency and the start of the second pass-band and, the greater this ratio, the greater the maximum attenuation in the stop bands. It should be noted that this attenuation is always finite.

In some cases it may be an advantage to adjust b so as to place the maximum attenuation where required, or to place the second pass-band in a suitable region. On the other hand, if b is made large, high attenuation is reached and the secondary pass-bands become very narrow. The limit in this direction is set by the impracticably large value of capacitance, or by extreme values of the normalized image functions which are not compatible with a practicable value of Z_0; i.e., the normalized image impedance becomes too low.

If the filter is to be used as a low-pass filter, attention must be paid to the higher pass-bands. As a result, the parameter b must be greater than unity, and the length of line always less than one radian per section. This makes the secondary bands very narrow, and their image impedances very far removed from those of the lower band. Their effect can be removed by the use of two sections with different values of b, or constructional differences can be employed to put them out of register in a multi-section filter.

A valuable feature is that for practicable values of b the shape of the image function curve is almost the same as that of a lumped filter and is independent of b. It is, therefore, possible to match two sections with different values of b, and also to use the existing charts for lumped filters to assess the performance in the pass-band. The phase characteristic is slightly inferior to that of the lumped filter.

The solutions for the various critical frequencies which are in terms of the transcendental function cot x/x are given in the attached charts for various values of b, and can be found for any value of x from the tables quoted in Ref. 7.

When a suitable value for b has been selected, the properties of the section can be examined on the charts and, if satisfactory, the required values of line lengths, capacitance, and Z_0 can be extracted from the design formulæ of the data sheet, Fig. 1.

Experimental Procedure

The equipment available for experimental work extended only to 700 Mc/s in frequency. In such circumstances, the correct procedure might have been to build models with cut-off frequencies well below 100 Mc/s, so as to leave a very wide 'attenuation' band for study, making over-size models in order to reproduce the u.h.f. limitations at lower frequencies. But the work was not started simply as basic research; the filters were made in response to definite demands for work in the range 100–500 Mc/s. It was therefore necessary to learn as much as possible using existing resources. If in the future it is possible to extend the working range of the equipment to, say, 2000 Mc/s, it may be possible to examine the designs more critically.

Equipment included a 70-ohm slotted line as a basic reference, with the aid of which variable-frequency standard terminations were made, and also some 10-dB buffer pads. Generators and receivers were available to the limit of 700 Mc/s and also a good piston attenuator. In addition, reflection detector instruments were also available for impedance measurement. Transmission-circuit connections were based on Telcon plugs and sockets which are claimed to be satisfactory up to 3000 Mc/s, though at times it was necessary to use inferior connectors. For insertion measurements between generator and receiver, 10-dB pads were maintained in circuit. The behaviour of the shunt elements could be determined by insertion-loss measurements between these pads. The adjustment of the elements is still, to some extent, empirical, nevertheless it has been possible to reproduce the theoretical characteristics of the models with some success.

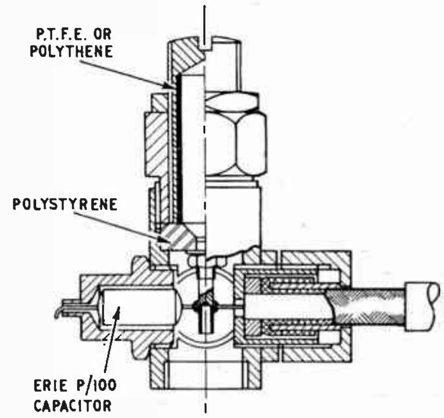
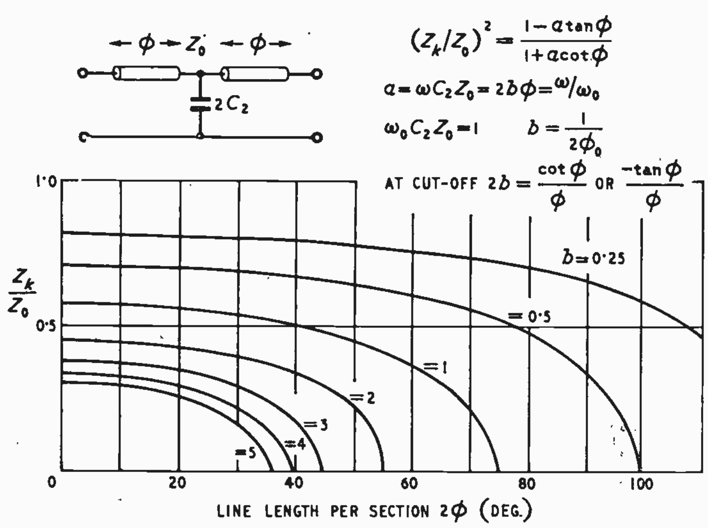


Fig. 5. *Section of trimmer block assembly showing cable terminal and fixed capacitor socket*

Fig. 4. *Electrical length of line plotted against $Z_k Z_0$*



Practical Models

Mains Filter

The first application of the capacitance-loaded line was made in 1950, as part of the filter for the mains and incoming wiring of a screened room. The complete filter was required to attenuate all frequencies from 100 kc/s to at least 3000 Mc/s.

The range 100 kc/s to 100 Mc/s was covered by a lumped filter, using carefully designed iron-core inductances but, because of component limitations, this would not necessarily attenuate higher frequencies. The v.h.f. region was therefore covered by a line-capacitance filter and the u.h.f. region was completed by a 'short line' filter of the type described in Ref. 3.

In power-supply filters, the pass-band is of no interest and, since it is only in this region that the image impedance could be important, the filter is usually designed around any convenient image impedance. This filter was therefore based on a convenient type of cable, U.R.1 (70 ohms) shunted by T.C.C. 'Large M' capacitors. Two different sections were used to minimize the effect of spurious pass-bands; one section was 31 in. long with 200-pF capacitors and the other 40 in. long with 300-pF capacitors. They were mounted either side of the u.h.f.

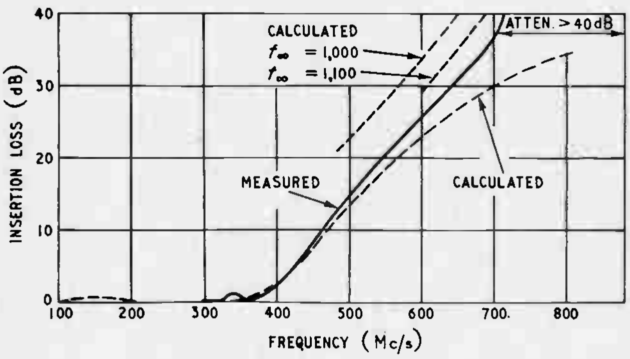


Fig. 6. 290-*Mc/s low-pass filter insertion loss*

section which was built into the walls of the screened room. The theoretical pass-bands were:

31 in.	0–28	126–134	253–256	398–400 Mc/s.
40 in.	0–24	96–103	192–196	288–290 Mc/s.

The attenuation of this filter rose rapidly from about 30 Mc/s and was over 70 dB at 600 Mc/s; no spurious pass-bands were located. The u.h.f. filter was known to start attenuation at about 600 Mc/s, but it may also have helped to eliminate spurious bands in the other filter.

Harmonic Filter

A 70-ohm filter was required to suppress generator harmonics in an unselective line-measuring set, and also to improve the accuracy of a wattmeter which depended on line-voltage measurement. Frequencies of interest centred around 290 Mc/s.

Three preliminary models were made, two single-section designs and a double section, with progressive improvement, until the final design was reached. All models used capacitor sections as shown on Fig. 12, but these were unsatisfactory, and the final model (Fig. 13) used specially designed compact concentric trimmer units.

In the first model, 70-ohm line sections were used, and an attempt was made to bring the second harmonic of 290 Mc/s near to maximum attenuation. This led to a value of the control parameter $b = 3/2$ bringing the electrical line length to $2\phi = 54°$ for a 70-ohm image impedance at 290 Mc/s; the other constants were

$\omega C Z_0 = 1$ at $2\phi = 2/3$ radian (or 38° or 204 Mc/s) giving $C = 11$ pF. Cut-off occurs at $2\phi = 62°$ (330 Mc/s) and max. attenuation, 16 dB per section, at 126° or 680 Mc/s.

The method of adjustment was to preset the capacitors before connection, and finally trim them for zero reflection at 290 Mc/s in a 70-ohm circuit. Insertion-loss measurements between 10 dB pads showed a cut-off at 400 Mc/s, and low attenuation beyond cut-off.

Two factors were considered to be wrong. First, the assumption that 70-ohm line would be suitable for a 70-ohm filter had led to working too near cut-off, so that the method of adjustment could lead to serious error if the filter elements were, in fact, not correct. This design had called for a rather large value of capacitance to go with its own residual inductance. Secondly, no allowance had been made for the fact that the capacitor box represented a large fraction of the line section with an impedance higher than 70 ohms.

The design was therefore modified to a somewhat higher cut-off frequency. A value of $b = 3{\cdot}07$ was chosen to give the following design figures:

$$2\phi = 31° \text{ at } 290 \text{ Mc/s}$$
$$2\phi = 18{\cdot}6° \text{ (175 Mc/s)}$$
$$f_c = 420 \text{ Mc/s } (2\phi = 45)$$
$$C = 6{\cdot}3 \text{ pF}$$
$$2\phi \text{ max} = 121° \text{ (1200 Mc/s)}$$
$$\alpha \text{ max} = 21 \text{ dB}$$

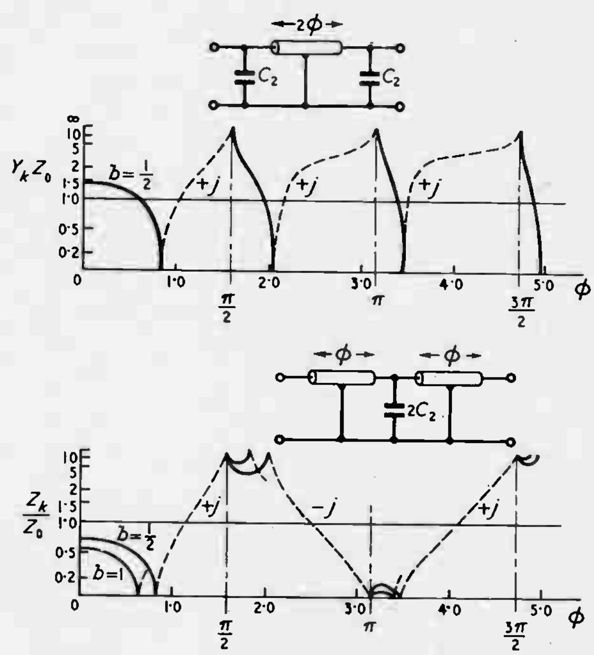
At 290 Mc/s, $Y_K Z_0 = 2$, giving $Z_0 = 140$ ohms. In setting up this filter, the trimmer was adjusted to include the capacitance of a section of 140-ohm line of the same length as the box, namely +1 pF; the final adjustment was carried out as before in terms of impedance match at 290 Mc/s.

A single section gave moderately close agreement with the theoretical insertion loss and was used for its intended purpose; i.e., to improve the accuracy of the impedance-measuring equipment. It made a noticeable improvement and enabled us to obtain a better performance from a two-section filter.

In the two-section model Fig. 13 the trimmers were set at 6·5 + 1 pF at the ends and 13 + 1 pF at the centre. (The allowance for line capacitance is $100/3Z_0$ pF per cm.) After a small adjustment at 290 Mc/s, open/short circuit measurements gave a value for Z_k of $68\underline{/1{\cdot}5°}$ ohms, and a reasonable insertion-loss curve.

In the final model, the trimmer capacitor unit was designed to be more compact and its use helped to bring theory and practice closer together. The trimmer was made in a number of capacitances ranges up to a maximum of 17 pF, together with a variety of accessory fittings, two of which are illustrated in Fig. 5. This trimmer and its descendants have since proved extremely useful in laboratory work for the rapid construction of

Fig. 7. *The march of the image functions Z_k/Z_0 and Y_kZ_0 for case $b = \frac{1}{2}$ ($\omega C_2 Z_0 = 1$ at $\phi_0 = \frac{1}{2}$) and partly for $b = 1$*



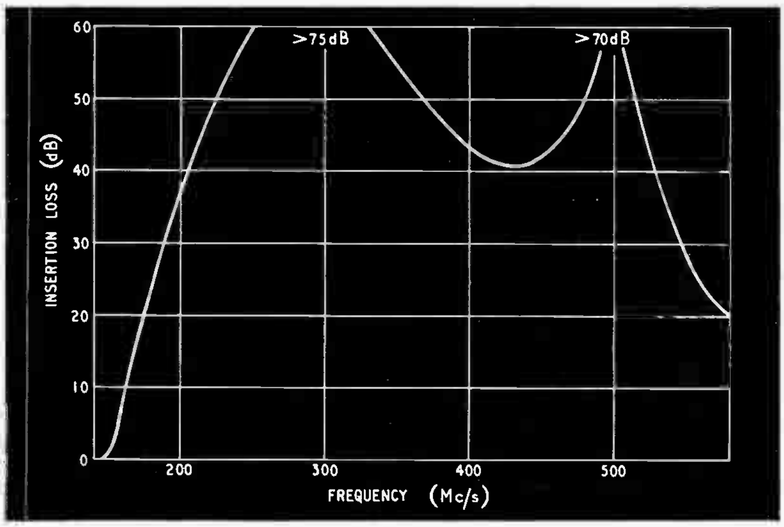


Fig. 8. *Insertion loss of* 190-*Mc/s filter*

networks such as π-transformers, hybrid rings, etc., up to a frequency of 1000 Mc/s.

Several copies of this design have been made, and in all of them it was found that, after pre-setting the capacitors in the way described above, very little correction was necessary to match the impedances at 290 Mc/s. A typical measured insertion-loss curve is given in Fig. 6 together with theoretical responses, calculated according to the method given in Appendix 2. It will be seen that the agreement is good up to 600 Mc/s, after which the practical insertion loss rises more rapidly than the theoretical loss.

Now this rising curve is characteristic of an m-derived filter, with a frequency of infinite attenuation. Such a state of affairs would exist if the shunt capacitors had a series resonance as a result of their self-inductance. It is easily possible to allow for this in the calculations by the use of a modified coefficient a' in place of $a = \omega C Z_0$. The expression then becomes

$$a' = a/(1 - x^2)$$

where $x = f/f_\infty$

Curves have been included in Fig. 6 for the cases where $f_\infty = 1000$ and 1100 Mc/s. A precise agreement cannot be expected, since two different values of capacitance exist in the filter, but it will be seen that the observed response could be adequately explained in this way if there were a resonance somewhere in the region of 1200 Mc/s. The self-inductance value (for the 13-pF capacitor) corresponding to such a resonance is about 2 mμH, which is of the same order as the value obtained by treating the connecting spigot of the capacitor as a short section of line. Filters similar to this design, but working at a frequency of 50% higher, have also been made successfully with the same components but with a shorter line section.

150-*Mc/s Low-Pass Filter*

This filter was required for aircraft installation in the aerial line of the v.h.f. R/T communications equipment, to prevent interference from experimental transmitters working at 290 and 465 Mc/s. The specification was:

(1) Attenuation not exceeding 1 dB in the frequency band 110–156 Mc/s.

(2) Insertion loss greater than 50 dB at 290 and 465 Mc/s.

(3) To work in 70-ohm line circuit.

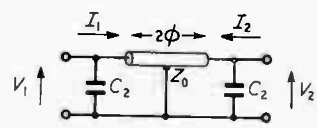
It appeared that a low-pass filter would be suitable and that three sections would be sufficient for the attenuation requirements. Some compromise in design was, as always, necessary: because the image impedance changes very rapidly near cut-off, a decision had to be made where to place the cut-off; if it were placed too near the upper working frequency, the insertion loss might be too high at this point. On the other hand the higher it is placed, the less will be the attenuation at 290 Mc/s.

It was decided to place this cut-off at 170 Mc/s and to adjust for equal attenuations at 290 and 465 Mc/s. From the design data of Figs. 1 and 2 this led to a choice of $b = 2$ as the control parameter, giving a full-section electrical length of 55° at the 170-Mc/s cut-off, and the following figures:

Frequency Mc/s	110	140	150	156	170
Elect. length 2ϕ°	35·5	45	48·4	50·5	55
y_k	1·7	1·3	1·11	0·85	0

With the line-section characteristic impedance chosen to match the network to 70 ohms at 140 Mc/s, the mismatch should be approximately the same at the two ends of the working band. The correct characteristic

Fig. 9. *Mid-shunt filter section*



impedance is thus, from the above, $1 \cdot 3 \times 70$ or 91 ohms, and could be obtained by the use of UR31 cable. Renormalizing the above values of y_k to 70 ohms; i.e., dividing by $1 \cdot 3$, then gave:

Frequency	110	140	150	156
y_k	1·31	1·0	0·85	0·66
Reflection coeff.	0·14	0	0·08	0·21

In the attenuating band, the frequencies 290 and 465 Mc/s correspond to $2\phi = 93 \cdot 5$ and 150° respectively with a nominal attenuation $\cosh^{-1}(\cos 2\phi - a \sin 2\phi)$ equal to $16 \cdot 3$ dB per section. This was barely enough to meet the specification and to improve it with only three sections would have required a lower cut-off with increased pass-band loss, or a higher value of the

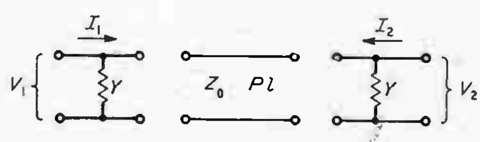


Fig. 10. *Filter broken into matrix sections*

parameter b, which would have made the capacitances excessively large. The addition of another section was not considered worth while, and it was agreed to proceed.

The control parameter $b = 2$ brings $\omega C Z_0 = 1$ at $\frac{1}{2}$ radian, or $28 \cdot 7°$, corresponding to 89 Mc/s. This gave a capacitance value of $19 \cdot 7$ pF per junction. The line is $\lambda/8$

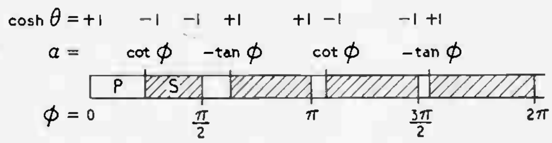


Fig. 11. *Diagram showing how the solutions of Equ.* (14) *fit together*

long at 140 Mc/s; i.e., 11,800/(8 × 140) or $10 \cdot 6$ in., in free space. To determine the physical solid-polythene cable length, $0 \cdot 375$ in. is subtracted for terminal junctions, and the remainder multiplied by the velocity factor of 2/3, giving $6 \cdot 75$ in. of polythene plus terminal leads. The terminal sleeves shown in Fig. 5 allow this length to be set accurately.

Now the capacitances of approximately 20 pF at the ends, or 40 pF at the inner junctions are both beyond the range of the trimmer unit of Fig. 5 and far too large to treat as simple capacitances at, say, 500 Mc/s. It was, therefore, necessary to add parallel capacitance, and the capacitor socket also shown in Fig. 5 was brought into use. This socket was designed to contain Erie Type P100 capacitors in a minimum inductance condition, and the arrangement made it possible to introduce a scheme which was not in the original plan. The loading capacitors, with their leads into the trimmers, could be made to resonate, and thus short-circuit the line at the two important attenuator frequencies.

The resonances were determined by inserting such a unit into a line between buffer pads, and observing the frequency response. The final values chosen were 12 pF for the terminal blocks, and 27 pF for the central ones, resonating at 465 and 290 Mc/s respectively with $\frac{1}{2}$–$\frac{3}{4}$ in. of folded leads.

The complete filter was made up as shown in Fig. 14 and adjusted in the manner previously described. At first, the cut-off frequency was somewhat low, and it was necessary to reduce the inner section capacitances slightly, to allow for the effect of their inductance which was noticeable at 150 Mc/s.

A typical response is shown in Fig. 8. Over most of the pass-band the insertion loss is below the prescribed value of 1 dB but, in the six models made, individual attention was necessary to keep below 2 dB at 156 Mc/s, since the cut-off position tended to vary. The two attenuation peaks caused by the series resonances were beyond the capabilities of our measuring equipment, but it will be noted that the upper peak has been moved to a higher frequency to help in the removal of the low attenuation at 580–600 Mc/s, the region of the second pass-band of the filter.

At this point it is pertinent to comment on a mistake in the design. Looking back, it will be seen that 290 Mc/s corresponds to a value of 2ϕ near 90°. Now the second pass-band in these filters always starts at $2\phi = 180°$; thus, if the attenuation becomes low at this point, the filter is liable to transmit the second harmonic of the 290-Mc/s transmitter. If the design had been pursued further, the next step would have been to introduce a different value for b in the two inner sections, and so move their second pass-bands. This is always a possible, though not guaranteed, cure because, for the same cut-off frequency, the shape of the image-admittance curve does not change appreciably with change of b.

Finally, it may be noted that the series resonant devices were extremely effective but that, in practice, each needed individual adjustment, and that considerable variations of attenuation response could be made. This is possibly because the tubular capacitors are not sufficiently alike at these high frequencies —something similar but more definable would be better in future.

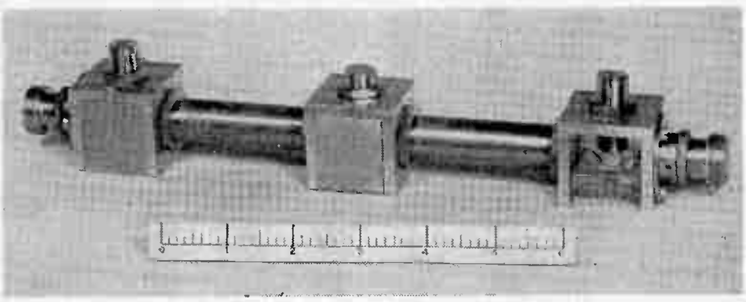
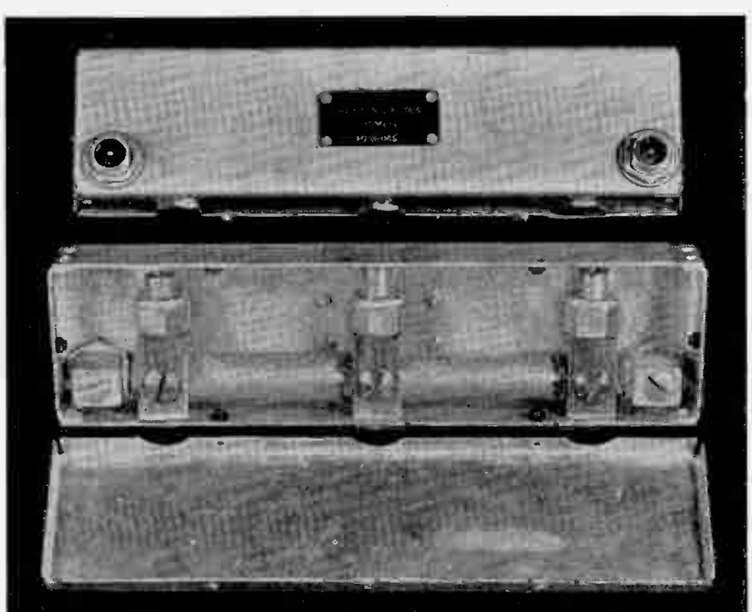


Fig. 12. *Early model of filter*

Fig. 13. *Harmonic filter*



Conclusion

A method of design has been evolved for filters using transmission lines in conjunction with lumped capacitance. The examples given show that this design can be carried into practice and the theoretical performance of the filters quite well reproduced up to at least 600 Mc/s. It is unfortunate that the equipment available would not enable the work to be extended to higher frequencies, to about 2000 Mc/s where the performance of the components and the effect of transmission-line discontinuities become of first-order importance. However, it may be noted that the analytical methods which have been developed, have also been used in the R.C.A. Laboratories[9] where the work has been carried up to 1000 Mc/s and the theory extended to make allowance for line discontinuities.

While the work described in this article was in progress considerable effort was expended in an attempt to find a method of design for filters corresponding to the Zobel *m*-derived filters, with improved pass-band response and an infinite attenuation frequency beyond cut-off. Working on a transmission-line basis with the analytical methods of this article, no solution was found; it was not possible to match the image functions at the junction between *m*- and prototype-sections. A type of *m*-derived filter has been made but, since its prototype design is rather different from that of the filters described here, it would more suitably be described separately.

APPENDIX 1

Analysis of the Network

A working knowledge of lumped element transmission circuits is assumed but, while these can be analysed by the application of Kirchoff rules to the meshes of the network, something more powerful is needed to deal quickly with circuits involving standing waves. The series transmission-line section is best treated in terms of its input and output currents and voltages. This element of the filter section is thus represented as a four-terminal network with prescribed properties. Four-pole matrix analysis has been developed for this purpose but, since it may not be well known, the calculations will be followed in some detail.

Matrix algebra is simply a generalized method of solving simultaneous equations; it reduces the work by reducing the processes to simple rules.[2, 5, 6, 8]

Fig. 9 shows a filter section in mid-shunt connection. The electrical length of the line component is given as 2ϕ, so that ϕ and C_2 are half-section element values. It should be noted that ϕ is proportional to frequency, being in fact $\omega l/v$ where l is the physical length of the half section, and v is the velocity of a wave in it.

The transmission constant θ will be calculated for a whole section, as is conventional, but it will be found more convenient to measure the elements in terms of their half-section values. This makes the equations simpler, and is probably the more rational way to treat any four-pole network. The factor 2 which causes so much confusion in dealing with full and half-section lumped networks, is however, still present.

The properties of any four-pole network can be described completely by a pair of simultaneous equations between the input and output currents and voltages. There are three arbitrary variables, since we can fix one current or voltage.

The equations can be written in three ways, starting with V_1 and V_2, I_1 and I_2 or V_1 and I_1. The last will be the most preferable since it can give us the form of an impedance V_1/I_1.

Thus

$$\left.\begin{aligned} V_1 &= AV_2 - B\,I_2 \\ I_1 &= CV_2 - D\,I_2 \end{aligned}\right\} \qquad (1)$$

where the coefficients $A\,.\,.\,D$ depend on the structure of the network, or in matrix form*

* The matrix $\begin{bmatrix} V_1 \\ I_1 \end{bmatrix}$ is known as the general network parameter matrix and is suitable for cascade connection; the matrix $\begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$ is known as the impedance matrix (*z*) and is suitable for networks in series, while $\begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$ is the admittance matrix used for networks in parallel. They can, of course, be transformed one to another—see Refs. 2 and 8.

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \times \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \qquad (2)$$

indicating $V_1 = A \times V_2 + B \times (-I_2)$, etc.

The negative sign has been transferred to I_2 in order to give a symmetrical layout, as in Fig. 9.

For any symmetrical four-pole network, the transmission constant θ is related by

$$\left.\begin{aligned} A &= D = \cosh\theta \\ B &= Z_k \sinh\theta \\ C &= \sinh/Z_k \end{aligned}\right\} \qquad (3)$$

and we can find Z_k and θ by solving these equations. The complete section of Fig. 1 consists of three four-poles in cascade (admittance–line–admittance) (Fig. 10) and the three matrices are shown below.

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \times \begin{bmatrix} \cosh Pl & Z_0 \sinh Pl \\ (1/Z_0)\sinh Pl & \cosh Pl \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \times \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \qquad (4)$$

If these three are multiplied out in the correct way,[5,6,8] we obtain the equivalent *ABCD* matrix for the whole section, and from this we can find θ and Z_k. The components will be assumed to be loss-free and, hence, we can write $Y = j\omega C_2$, $\cosh Pl = \cos 2\phi$ and $\sinh Pl = j \sin 2\phi$. The multiplication then gives:

$$\left.\begin{aligned} A &= D = \cos 2\phi - \omega C_2 Z_0 \sin 2\phi \\ B &= jZ_0 \sin 2\phi \\ C &= j/Z_0\,[2\omega C_2 Z_0 \cos 2\phi + \sin 2\phi(1 + \omega^2 C_2{}^2 Z_0{}^2)] \end{aligned}\right\} \qquad (5a)$$

For convenience, let $\omega C_2 Z_0 = a$ = normalized admittance of the shunt arm.

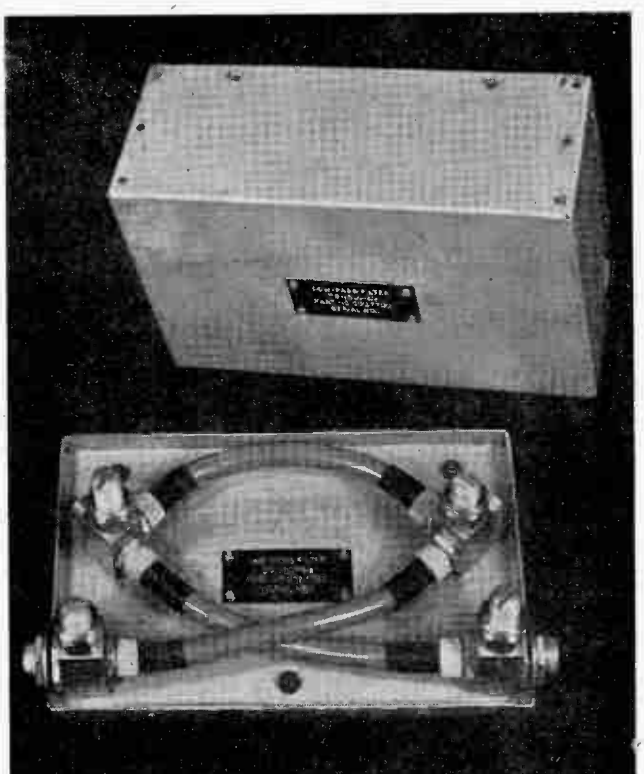
Then

$$\left.\begin{aligned} A &= D = \cos 2\phi - a \sin 2\phi \\ B &= jZ_0 \sin 2\phi \\ C &= j/Z_0\,[2a \cos 2\phi + (1 - a^2)\sin 2\phi] \end{aligned}\right\} \qquad (5b)$$

Transmission Constant

From Equ. 5, when compared with (3), which is true for any symmetrical network

Fig. 14. 150-*Mc/s low-pass filter*



$$\cosh\theta = \cos 2\phi - a\sin 2\phi \quad \ldots \quad (6)$$

gives the transmission constant θ of the section.

Now, with reference back to Equ. 1, if terminals 2 are open-circuited, then $I_2 = 0$ and the right-hand matrix column $= 0$, giving

$$Z_{oc} = \frac{V_{1o}}{I_{1o}} = \frac{A}{C}$$

whereas if terminals 2 are short-circuited

$$Z_{sc} = \frac{V_{1s}}{I_{1s}} = \frac{B}{D}$$

From these two results we have, for any network

$$Z_k^2 = Z_{oc}\,Z_{sc} = \frac{AB}{CD} \quad \ldots \quad (7)$$

and if the network is symmetrical, $A = D$, then

$$Z_k^2 = B/C \quad \ldots \quad (7a)$$

Terminology

It is usually more satisfactory to work in impedance coefficients when dealing with series circuits, and in admittance coefficients when using parallel circuits, rather than to use impedance for both. It seems rational, therefore, to confine our interests to impedances for mid-series connections, and to admittances for mid-shunt. This may be a bold move, but it does, in fact, make the equations of this filter neater and it also permits us to use the terms like Y_k and Z_k without the extra subscripts, T or π.

It is also helpful to normalize these image functions against, say, the Z_0 of the line, e.g.,

$$z_k = Z_k/Z_0 \qquad y_k = Y_k Z_0$$

Mid-Shunt Image Admittances

Inserting the values of B and C in Equ. (7a) and inverting, we find

$$y_k^2 = 2a\cot 2\phi + 1 - a^2 \quad \ldots \quad (8)$$

The form of this function is shown in Fig. 7 and it has been found by trial that it approximates, in the first pass-band, very closely to the form $1 - (\omega/\omega_c)^2$ of the corresponding lumped element low-pass filter, where ω_c is the cut-off frequency.

Mid-Series Image Impedance

To find the mid-series image function, we must find a new set of *ABCD* values by rearranging the circuit and the matrices. The algebra has been found simpler if a half-section is taken this time. The matrices then become

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \cos\phi & jZ_0\sin\phi \\ (j/Z_0)\sin\phi & \cos\phi \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \times \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}$$

from which

$$\left.\begin{aligned} A &= \cos\phi - a\sin\phi \\ B &= jZ_0\sin\phi \\ C &= j/Z_0\,(\sin\phi + a\cos\phi) \\ D &= \cos\phi \end{aligned}\right\} \quad \ldots \quad (9)$$

This is, of course, no longer symmetrical, and the angles are half-section angles. The transmission constant is for a half section, and is given by

$$AD = \cosh^2\theta/2 = \cos^2\phi - a\sin\phi\cos\phi$$

The value of $\cosh\theta$ for a whole section given in Equ. (6) can be obtained from this by operating

$$\cosh\theta = 2\cosh^2\theta/2 - 1$$

The mid-series image function is, from (9),

$$Z_k^2 = \frac{AB}{CD} = \frac{\sin\phi\cos\phi - a\sin^2\phi}{\sin\phi\cos\phi + a\cos^2\phi}$$

$$= \frac{1 - a\tan\phi}{1 + a\cot\phi} \quad \ldots \quad (10)$$

An example is also given in Fig. 7.

As an exercise, the matrix order may be reversed, when equations (6) and (8) for y_k should eventually be obtained.

Critical Frequencies

The frequencies of most interest are the edges of the pass-bands, some form of 'mid-frequency' and the first frequency where $y_k = 1$.

The cut-off frequencies can be found by solving for $\cosh\theta = \pm 1$. Cosh $\theta = 0$ can be taken as a mid-frequency, and this often turns out to be a useful point. The solution $y_k = 1$ must be taken from Equ. (7a), but there is no solution of $Z_k = 1$ for this filter. The cut-off frequencies may also be found by putting the image functions equal to zero and infinity. In all cases the solution will be multiple since we are dealing with circular functions.

The function $\cosh\theta$ ranges from unity, when $\theta = 0$, to ∞ when $\theta = \infty$ and, since $\cosh\theta = \cosh(-\theta)$ (in the same way as $\cos\theta$), the range must also run from -1 to $-\infty$. Outside the limits -1 to $+1$, therefore, θ has a real value and, from its definition, must represent an attenuation, $\theta = \alpha$. Inside the range $x = -1$ to $+1$, $\cosh^{-1}x$ has imaginary values $j\cos^{-1}x$ and so, in this region, there is a pass-band with a transmission phase-angle: $\theta = j\beta$.

To find the cut-off and mid-frequencies we solve

$$\cos 2\phi - a\sin 2\phi = -1,\ 0,\ +1$$

If the algebra becomes too complex, we can always try the alternative $z_k, y_k = 0, \infty$

In this case, however, we can proceed from $\cosh\theta$.

From Equ. (6) the 'mid-frequency' is ($\cosh\theta = 0$) obviously given by:

$$\cot 2\phi = a \quad \ldots \quad (11)$$

Also, since $\cos 2\phi = 0$ when $\sin 2\phi = \pm 1$ and vice versa, then clearly one set of solutions for f_c is given by

$$\cos 2\phi = \pm 1 \ldots 2\phi = n\pi \quad \ldots \quad (12)$$

Another set can be found by putting $\cosh^2\theta = 1$, then

$$\cos^2 2\phi + a^2\sin^2 2\phi - 2a\sin 2\phi\cos 2\phi = 1$$

$$\therefore\ a^2\sin^2 2\phi - 2a\sin 2\phi\cos 2\phi = 1 - \cos^2 2\phi = \sin 2\phi$$

$$\therefore\ \sin 2\phi\,(a^2 - 1) = 2a\cos 2\phi$$

$$\text{and } \cot 2\phi = \frac{a^2 - 1}{2a} \quad \ldots \quad (13)$$

This expression can be reduced with the aid of a page of trigonometry, but there is a more simple method. The standard expression for $\cot 2\phi$ is

$$\cot 2\phi = \frac{\cot^2\phi - 1}{2\cot\phi}$$

When this is compared with Equ. (13) it is seen at once that the reduction is

$$\cot\phi = a$$

However, if we invert Equ. (13) we also recognize the form of $-\tan 2\phi$, so another solution is

$$\tan\phi = -a$$

Thus the total of solutions is

$$\left.\begin{aligned} &(f_m).\ \cosh\theta = 0 \text{ when } \cot 2\phi = a \\ &(f_c).\ \cosh\theta = \pm 1 \text{ when } \phi = n\pi/2 \\ &\qquad \text{or } a = \cot\phi \\ &\qquad \text{or } a = -\tan\phi \end{aligned}\right\} \quad \ldots \quad (14)$$

This appears to be a complete set, since we now have a value of ϕ to suit each quadrant.

These solutions have multiple values. If the coefficient a were constant, they would recur at regular intervals of $\pi/2$. But a is proportional to frequency and to ϕ and will, therefore, be greater in each successive band. Since $\cot\phi$ is positive in odd quadrants and has its high value near 0, π, etc., whereas $\tan\phi$ is negative in even quadrants, with its high value near $\pi/2$, $3\pi/2$, etc., there will be a pass-band in the lower end of each quadrant of ϕ, and the bands will become progressively narrower. Fig. 11 shows how the solutions fit together.

Attenuation

Equation (6) for $\cosh\theta$ can never give an infinite value (except at infinity) since $\cos 2\phi$ and $\sin 2\phi$ never exceed ± 1. It will, therefore, rise to a maximum finite value somewhere in each stop-band. Since a is increasing with ϕ, this maximum will be progressively greater in each such band. However, because $\sin 2\phi$ and $\cos 2\phi$ alternate like and different signs, there will be an undulation superimposed on this set of maxima. The positions and amplitudes of these maxima will be found later when we have rationalized the equations for design purposes.

Design Data

The equations above are not in the best form for general design work. It is necessary to establish a link between ϕ and $\omega C_2 Z_0$ so that the frequency variable can be put into the normalized form ω/ω_c.

Suppose we take a datum frequency ω_0 such that $\omega_0 C_2 Z_0 = 1$. The coefficient a now becomes ω/ω_0 (and also ϕ/ϕ_0). To link ϕ to ω a new coefficient b is introduced so that

$$a = \omega/\omega_0 = \phi/\phi_0 = b\, 2\phi$$

Then $b = 1/2\phi_0$ where $2\phi_0$ is the electrical length of the full section line when $\omega_0 C Z_0 = 1$, and is a constant for any particular filter. The factor b is thus a single parameter which we can vary to control all the normalized properties of the filter.

Equations (14), which give the critical frequencies, then become

$$(f_m) \ldots \cosh\theta = 0 \quad \ldots \cot 2\phi = b.2\phi$$

$$\text{i.e., } \frac{\cot 2\phi}{2\phi} = b$$

$$(f_c) \ldots \cosh\theta = \pm 1 \ldots \quad \phi = n\pi/2$$

$$\frac{\cot\phi}{\phi} = b$$

$$\frac{-\tan\phi}{\phi} = b$$

These transcendental equations are not soluble but tables of them exist[7] and, therefore, if the value of 2ϕ for cut-off can be decided, b becomes fixed, and also the CZ_0 product. To split this, we must consider the relation between Z_k or Y_k and Z_0 through the pass-band. All these operations are best carried out with the aid of a set of generalized curves of y_k and z_k as a function of ϕ and b, such as the attached charts Figs. 3 and 4.

The decision on cut-off ϕ_c will be determined in terms of the attenuation response, or the width or relative position of the higher order pass-bands. It will be clear from the various curves that the smaller we can make $2\phi_c$ for the required cut-off frequency, then the greater will be the attenuation rate, and also the narrower will be the upper pass-bands. This is desirable in a low-pass filter, but the lower limit of $2\phi_c$ will be determined by practical considerations of how large we can make a true capacitance, or what value of Z_0 can be used. In one of the examples given, the cut-off angle was chosen critically in the hope of enclosing the harmonics of an oscillator inside attenuation bands.

Maximum Attenuation

Consideration of attenuation was left until the factor b had been introduced. The point of maximum attenuation may be found as usual by differentiating and equating to zero.

Thus

$$\cosh\theta = \cos 2\phi - b.2\phi \sin 2\phi \quad \ldots \quad (16)$$

$$\frac{d}{d(2\phi)}\cosh\theta = -\sin 2\phi - [b \sin 2\phi + b.2\phi.\cos 2\phi] = 0$$

$$\therefore (1 + b)\sin 2\phi = -b.2\phi.\cos 2\phi$$

$$\text{or } \frac{\tan 2\phi_m}{2\phi_m} = \frac{-b}{1+b}$$

where $2\phi_m$ is the angular line-length corresponding to maximum attenuation.

If, from the above, we take

$$\sin 2\phi = \frac{2b\phi}{1+b}.\cos 2\phi$$

and substitute Equ. (16) we find the value of the maximum attenuation from

$$\cosh\theta_m = \left(1 - \frac{(2b\phi)^2}{1+b}\right)\cos 2\phi_m \quad \ldots \quad (17)$$

Values of attenuation ($\alpha = \theta$) and the angles $2\phi_m$ and $2\phi_c$ are plotted in Fig. 2 as a function of b.

Image Functions

The image functions can be plotted from equations (8) and (10) and are both illustrated in Fig. 7. It will be seen that the mid-series impedance never crosses the value Z_0 in any of the pass-bands. The higher pass-bands show the characteristics of a band-pass filter, but alternate between mid-shunt (high) and mid-series (low) forms. A little reflection will show that this should be so, since we are looking at a mid-shunt type of junction (C_2) through alternate quarter and half-wavelengths of line. The impedance is always high or low in these secondary bands, a fact which we may use to prevent their passing much energy.

The mid-shunt admittance characteristic is of similar form in the low-pass band, but rides from infinity to zero in each other band. This type of characteristic corresponds with those marked 'not usually of interest' in Shea's filter charts, and is not likely to be useful. y_k is equal to unity at a point where $\cot 2\phi/2\phi = b/2$ (found by putting Equ. (8) equal to unity), but the normalized mid-series impedance never reaches unity.

To complete the charts of the image functions, it is necessary to show the curves in the attenuating bands where the functions are imaginary. To allot a sign to each branch we may use Foster's Reactance Theorem, which in its stated form applies only to reactances. However, if we define admittance in such a way that $Y = j\omega C$ corresponds with $X = j\omega L$, the theorem applies also to susceptances. Thus we can say that, in any loss-free combination of elements, the total reactance or susceptance always rises with increasing frequency. They can only pass from positive to negative values by passing through the point at infinity and in this way the charts can be completed.

APPENDIX 2

Evaluation of Insertion Loss of a Two-section Filter

The insertion loss of a four-pole network can be expressed in terms of the elements $ABCD$ of its general parameter matrix. When the network is symmetrical and purely reactive, the expression[8] reduces to

$$L = 10\log_{10}\left[1 + \tfrac{1}{4}\left(\frac{B}{jR} - \frac{CR}{j}\right)^2\right] \quad \ldots \quad (1)$$

where R is the value of the generator and load impedances.

We can calculate B and C for the network, or we can use the results derived from this expression for any symmetrical four-pole, using the relations [equations (3) and (7a) of Appendix 1].

$$\left.\begin{array}{l} A = D = \cosh\theta \\ B = Z_k \sinh\theta \; C = Y_k \sinh\theta \\ Z_k^2 = B/C \end{array}\right\} \quad \ldots \quad (2)$$

which give

$$L = 10\log_{10}\left[1 + \tfrac{1}{4}\sinh^2 n\alpha\left(\frac{Z_k}{R} - \frac{R}{Z_k}\right)^2\right] \quad \ldots \quad (3)$$

in the attenuating band, or:

$$L = 10\log_{10}\left[1 + \tfrac{1}{4}\sin^2 n\beta\left(\frac{Z_k}{R} - \frac{R}{Z_k}\right)^2\right] \quad \ldots \quad (3a)$$

in the pass-band, where α or $\beta = \cosh^{-1} A$ and n is the number of identical sections in cascade. In the former case, it is necessary to find the values of B and C for the complete cascade network; in the latter case it is necessary to evaluate Z_k but, once this is done, the loss curves can be calculated for any number of sections by just changing the value of n. The latter method has the disadvantage that Z_k becomes zero or infinite at cut-off frequencies, so that the result becomes indeterminate at these points. The former method was used for the curves of Fig. 6.

For the two-section filter, a consideration of the early part of Appendix 1 will show that to find the matrix elements $A'B'C'D'$ for the complete filter in terms of the single-section elements $ABCD$, it is necessary to square the matrix.

This gives

$$\left.\begin{array}{l} A' = A^2 + BC = D' \\ B' = 2AB \\ C' = 2AC \end{array}\right\} \quad \ldots \quad (4)$$

Inserting these in equations (1) above, and remembering that $Z_0 = 2R$ in our particular case, we have

$$L = 10\log\left[1 + A^2\left(\frac{2B}{jZ_0} - \frac{CZ_0}{2j}\right)^2\right] \quad \ldots \quad (5)$$

using for A, B, C the values given in equation (5a) of Appendix 1.

REFERENCES

1 Mason and Sykes, "The Use of Coaxial and Balanced Transmission Lines in Filters and Wide-band Transformers for High Radio Frequencies", *Bell Syst. tech. J.*, 1937, p. 275.
2 Ragan, "Microwave Transmission Circuits", M.I.T. Radiation Laboratory Series, No. 9, McGraw-Hill, 1947.
3 Oatley and Burrell, "Ultra-High-Frequency Filters", *J. Instn elect. Engrs*, Part IIIa, 1946, No. 8, p. 1338.
4 Karakash, "Transmission Lines and Filter Networks", Macmillan, 1950.
5 Kron, "Tensor Analysis of Networks", (1939).
6 Aitken, "Determinants and Matrices", Methuen Monograph.
7 Tables of Functions like tan x/x will be found in Jahnke and Emde, "Funktion Tafeln", and can also be obtained from the authors.
8 P. I. Richards, "Applications of Matrix Algebra to Filter Theory", *Proc. Inst. Radio Engrs*, March 1946, p. 145P. (This is excellent.)
9 H. B. Yin and T. L. Foley, "Filter using Coaxial Transmission Lines as Elements", *RCA Rev.*, March 1954, p. 62 (also as RCA Reprint No. 1034).

ZETA

RELEASING THERMONUCLEAR ENERGY

THE experimental Zero Energy Thermonuclear Apparatus at Harwell employs pulse circuitry on a grand scale. The object of the experiments is to raise heavy hydrogen to an extremely high temperature and maintain it in this state long enough for the gaseous nuclei to collide and fuse, releasing energy. The process is the same as that which takes place in a hydrogen bomb but, in the latter case, the required high temperatures are produced by letting off an atomic (i.e., fission) bomb, a procedure which does not yield industrially-usable power.

In ZETA and similar pieces of apparatus developed elsewhere, a small quantity of heavy hydrogen (deuterium) is introduced into an evacuated tube, ionized by a weak r.f. field, and then heated by passing a very high current pulse through it. In the Harwell machine, currents of about 200,000 A are used, the pulse duration being a few milliseconds, and temperatures of about 5×10^6 °C are attained. The power output, which is measured in terms of the number of neutrons produced per pulse, is quite small. Temperatures of about 10^8 °C are required to yield usable outputs, and further work will be directed to the attainment of such high temperatures.

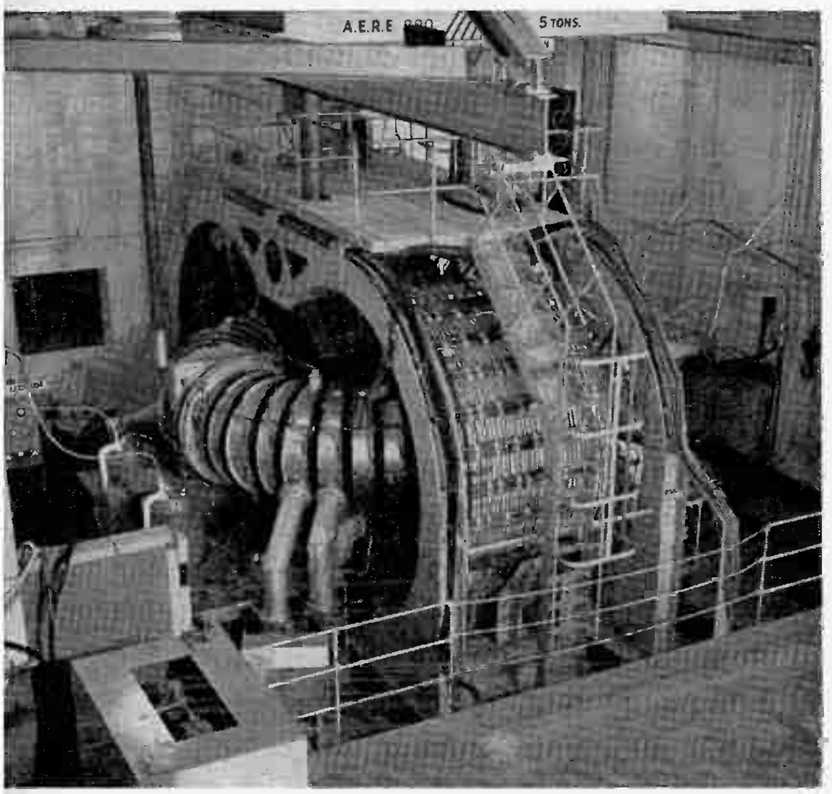
The required high-current pulses are obtained by discharging a capacitor through a pulse transformer. The secondary of the transformer has one turn, which is the gas-discharge path itself, bent into a ring. The capacitor is charged via S_1 with S_2 and S_3 open. It is discharged via S_2, initiating the output pulse. When most of the energy stored in the capacitor has been transferred to the inductance of the pulse transformer, S_2 is opened and S_3 closed, permitting current to continue to flow through the transformer primary and thereby lengthening the pulse. When we said that the circuitry is on the grand scale, we were referring to its physical size. The capacitor is actually a bank of 30-μF, 25-kV working capacitors totalling 1,600 μF, and is housed in a large room. The core of the pulse transformer weighs 150 tons. There are sundry additional chokes, not shown in Fig. 1, used for shaping the pulses. A 500-μH choke is housed in a can the size of three or four telephone kiosks.

The charging current is obtained by means of a high-voltage transformer and a selenium rectifier. It is limited to 4 A by means of a saturated thermionic valve (actually a large triode). An alternative discharge path is provided for use in case of a fault, and takes the form of a pair of ignitrons in series with a resistor to dissipate the power.

In order to prevent cooling of the hot gas, the latter must be kept away from the walls of the containing tube. The magnetic field produced by the high current flowing in the gas is sufficient to constrict the discharge but, unfortunately, such a constricted discharge is unstable. Any small kink in the discharge path results in a modified magnetic field which in turn accentuates the kink. To confine the discharge to the required path, an axial magnetic field, produced by a winding on the containing tube, is employed. The field required is quite small (200 gauss). The discharge path can still move about bodily inside the tube but, if it tends to do so, eddy currents induced in the tube walls, which are of metal, produce fields in the direction required to counteract such movement. The limitation on the temperature achieved in ZETA is therefore believed to be imposed by the power supply, not the instability of the discharge, and more capacitors are to be installed to enable higher temperatures to be obtained.

A considerable amount of electronic instrumentation is employed for making measurements on ZETA and other smaller equipments at Harwell. Neutrons are counted by means of scintillators. The position of the current path is investigated by inserting a pair of probes between which a p.d. is applied, the extent of conduction depending on the density of the ionized gas. The presence of ionized gas is also detected by millimetric waves. A transmitting path across the tube is blocked when the gas becomes highly conducting. Use is also made of the fact that the current induced by the ion stream in a coil round the tube varies according to the position of the stream. A system of pickup coils can then be arranged to yield information on a co-ordinate basis.

Eight sets of readings can be recorded simultaneously by photographing the traces on an eight-channel precision oscillograph supplied by A. E. Cawkell. In addition to the items mentioned,



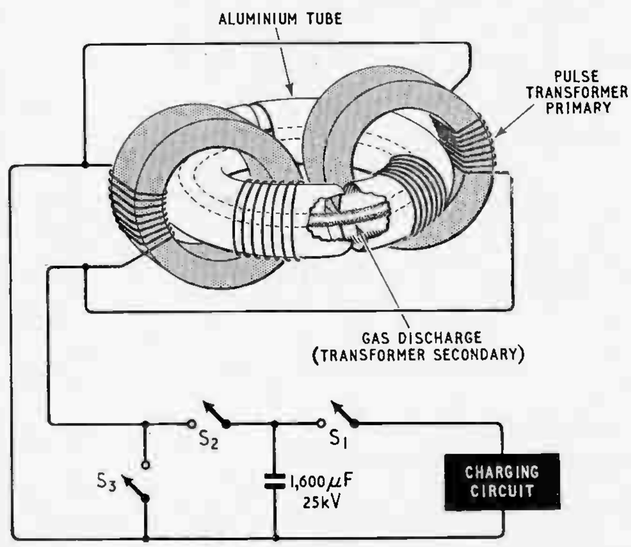


Fig. 1. *Basic circuit of ZETA. The field winding round the aluminium tube is for preventing kinks in the current path*

General view of ZETA

measurements may be made of circuit voltages and currents, intensity of light emitted by the discharge, X-ray output, etc.

A special feature of the eight-channel oscillograph is a circuit for keeping the trace brightness constant irrespective of the sweep duration. The linearity is also very good ; for example, the velocity error of the X-deflection system is less than 1% between beams and at opposite ends of one beam.

Another equipment by the same makers enables displays of transient phenomena to be retained for as long as desired. This is a two-channel display equipment which employs 'Memotron' storage tubes. The latter are two-gun devices. One gun is employed for writing and one for reading. The reading gun continuously 'floods' a special type of mesh target with low-speed electrons. The target consequently acquires a negative charge and electrons from the reading gun do not pass through the holes to strike the fluorescent screen. The writing gun employs a beam which, when it strikes the target, releases secondary electrons, the areas struck then assuming a positive potential. Electrons from the reading gun now pass through these areas to produce a picture on the screen which persists and so can be examined at leisure. Erasure is affected by lowering the potential of a collector electrode near the target.

Some 'windows' in the discharge tube enable measurements based on light output to be made. The hot deuterium itself does not emit light, but oxygen does, and a small quantity is admitted for measurement purposes. Movement of the conducting path is investigated by means of image-converter cameras, and spectroscopic measurements are made. An elegant application of the spectroscope enables the very high temperatures encountered to be measured. Light-emitting gases at low temperatures produce the familiar line spectra. As the temperature is increased, the spectral lines are found to broaden. The cause of this broadening is Doppler effect. The emitted light comes from particles which are rushing about in all directions. At any moment, some particles will be moving away from the observer and some towards him, producing Doppler frequency shifts about the mean spectral frequency. The resultant broadening is a function of the temperature and may, therefore, be used to measure it.

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.

Stacked Valve Circuits

SIR,—Referring to this article in your November issue, I would like to point out that the position of the cathode is incorrectly shown in Fig. 2(a) and (b). The resistor R_k $(1 + \mu)$ should be separated into two parts R_k and μR_k in series with μR_k uppermost and the dashed line indicating the cathode between them.

On p. 405, second column, second line, the inequality sign should be reversed so that R_0 is very large, not very small.

Kenilworth, Warwicks. A. W. KEEN
7th December 1957.

SIR,—I cannot agree with Mr. Keen's criticism of Fig. 2 in my article. These diagrams are equivalent circuits for the purpose of current calculation only and in no way attempt to explain potential changes at points in the actual circuit. The position of the cathode was chosen to show the symmetry of $(\mu + 1)R_k$ when viewed from 'above' the cathode and $(R_L + r_a)/(\mu + 1)$ when viewed from 'below' the cathode. (The grid and anode are considered as points above the cathode, and the cathode connection as being below the cathode.)

As you can see, to alter the position of the cathode would alter the whole basis of the article and, while it may be argued pedantically that its correct position is where Mr. Keen suggests, pragmatically it is satisfactory in the position as shown.

I agree, of course, about the reversal of the inequality sign.

Auckland University College, J. B. EARNSHAW
New Zealand.
23rd December 1957.

The Transactor

SIR,—I am interested to see that Mr. Keen[1,2] and I[3,7] have been working on similar lines in connection with ideal active elements and have independently coined the name 'transactor' for them. It must be stated, however, that only the voltage–current and current–voltage transactors are fundamental and that the current–current and voltage–voltage transactors may be derived from them, as I have previously pointed out to Mr. Keen[5].

Furthermore, I am not satisfied with Mr. Keen's treatment of certain aspects of the subject, for his method of splitting the matrix of the general active four-pole does not, in fact, separate the active from the passive parts. A general analysis is complex and I shall give it elsewhere. I shall restrict the present analysis to real admittance coefficients. If Mr. Keen's derivation can be shown not to hold in this restricted case, then it obviously does not hold in the general case.

The active four-pole of Fig. 1 can be specified by

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = Y \times \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \quad \ldots \quad (1)$$

with $$Y = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \quad \ldots \quad (2)$$

and all y terms are real.

The active four-pole defined by (1) and (2) must obey certain conditions of stability if it is to be classed as a transmission element ;

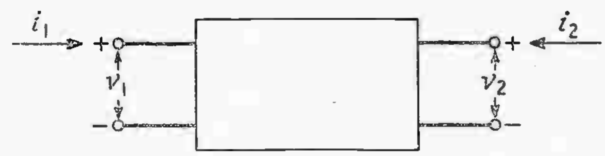


Fig. 1

otherwise it might oscillate when inserted between parts of a system. It is therefore necessary to have

$$\text{either } y_{11} > 0 \quad \text{or } y_{22} > 0 \quad \ldots \quad (3)$$

$$\text{and } y_{11}y_{22} - y_{12}y_{21} > 0$$

$$\text{Let } y_{12} = a + s, \quad y_{21} = a - s \quad \ldots \quad (4)$$

so that (2) becomes

$$Y = \begin{bmatrix} y_{11} & a \\ a & y_{22} \end{bmatrix} + \begin{bmatrix} 0 & s \\ -s & 0 \end{bmatrix} \quad \ldots \quad (5)$$

Substitute (4) into the last inequality of (3), when

$$y_{11}y_{22} - a^2 + s^2 > 0 \quad \ldots \quad (6)$$

The power entering the four-pole is

$$P = i_1v_1 + i_2v_2 = y_{11}v_1^2 + (y_{12} + y_{21})v_1v_2 + y_{22}v_2^2$$
$$= y_{11}v_1^2 + 2a\,v_1v_2 + y_{22}v_2^2 \quad \ldots \quad (7)$$

on substituting from (4).

If the four-pole is active P must be negative; i.e., it must supply power to the external circuit. P cannot be negative for all values of v_1, v_2 because the squared terms of (7) are positive by virtue of (3). Hence the four-pole will only be active for values of v_1, v_2 for which the quadratic form of (7) is negative ; that is, for which

$$y_{11}y_{22} - a^2 < 0 \quad \ldots \quad (8)$$

The s coefficient thus has no effect on the power, but is a stabilizing component as per (6). Since $y_{11}y_{22} > 0$ because of (3) activity is only associated with the coefficient a.

One can therefore split (5) into three terms.

$$Y = \begin{bmatrix} y_{11} & 0 \\ 0 & y_{22} \end{bmatrix} + \begin{bmatrix} 0 & a \\ a & 0 \end{bmatrix} + \begin{bmatrix} 0 & s \\ -s & 0 \end{bmatrix} \quad \ldots \quad (9)$$

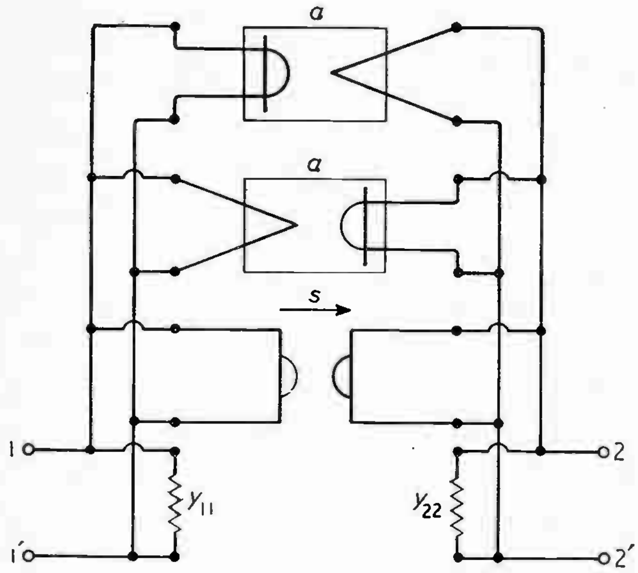


Fig. 2

in which the y part represents conductance connected across the 11′ and 22′ terminals (dissipation), the a part represents two voltage-current transactors[3] connected in opposition between these terminals (activity) and the s part represents a gyrator connected between them (lossless coupling) as shown in Fig. 2.

Mr. Keen starts off with equations (1) and (2) but writes $y_{12}' = y_{21} - y_{12}$ to obtain

$$Y = \begin{bmatrix} y_{11} & y_{12} \\ y_{12} & y_{22} \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ y_{12}' & 0 \end{bmatrix}$$

or in terms of a and s

$$Y = \begin{bmatrix} y_{11} & a+s \\ a+s & y_{22} \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ -2s & 0 \end{bmatrix}$$

and he ascribes the activity to the s part. This seems to me wrong, for I have shown above that s does not affect the activity. It is his left-hand matrix which is responsible for the activity but, as it also includes the other admittance terms, it is not in its simplest form.

The four-pole may however be 'unilateralized' by arranging for $a+s = 0$ or $a-s = 0$ when it will transmit from 11′ to 22′ or from 22′ to 11′. However, it is s that must be *varied* when attempting such a lossless transformation and *not a*, which is an *invariant* coefficient characterizing the activity of the four-pole.

There is one further point I should like to make with regard to Mr. Keen's transactor realization given in Fig. 5 of his paper. On the left-hand side he depicts valve circuits comprising only *three* terminals, on the right-hand side he depicts transactor circuits comprising *four* terminals. This does not seem right to me[8], and it appears that both the theoretical derivation and the physical realization of the transactor elements are more sophisticated than Mr. Keen indicates.

British Telecommunications Research Limited, GERALD E. SHARPE
Taplow, Berks.
30th January 1958.

REFERENCES

1. A. W. Keen, "Transactive Network Elements", *J. Instn elect. Engrs*, April 1957, Vol. 3, New Series, No. 28, pp. 213–214.
2. A. W. Keen, "The Transactor; An Idealised Network Element", *Electronic and Radio Engineer*, Dec. 1957, Vol. 34, New Series, No. 12, pp. 459–61.
3. G. E. Sharpe, "Ideal Active Elements", *J. Instn elect. Engrs*, Jan. 1957, Vol. 3, New Series, No. 25, pp. 33–34.
4. See also: D. Morris, "Ideal Active Elements", *J. Instn elect. Engrs*, May 1957, Vol. 3, New Series, No. 29, pp. 272–273.
5. G. E. Sharpe, "Ideal Active Elements", *J. Instn elect. Engrs*, July 1957, Vol. 3, New Series, No. 31, pp. 430–431.
6. G. E. Sharpe, "Transactors", *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, pp. 692–693.
7. G. E. Sharpe, "The Pentode Gyrator", *Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CT4, No. 4.
8. A physical realization of a voltage-current transactor will be found in ref. 5.

SIR,—In my original contribution on this subject (*J. Instn. elect. Engrs*, April 1957, p. 213) I defined transactive network elements as the class of "two-terminal-pair networks characterized by *G*, *H*, *Y* or *Z* matrices having a null principal diagonal, particularly those having one zero element also in their secondary diagonal". There is no basis in this definition for Mr. Sharpe's contention that certain transactors are more fundamental than others; moreover, the derivations given in my subsequent article are entirely consistent with this definition. Nothing I have written precludes the three-terminal form of transactor or, conversely, the association of an ideal one-to-one transformer with a three-terminal realization to convert it to four-terminal form.

Mr. Sharpe's analysis of the active network is similar to that of J. Shekel (*Proc. Inst. Radio Engrs*, August 1954) and he uses it to show that my matrix subdivision does not separate the active from the passive parts of the network. It was not my intention to do so. My expressed intention was to separate from the entire matrix a component which conforms to the above definition, and is therefore strictly transactive, leaving a residual symmetric matrix which will, in general, inevitably contain both active and passive parts. There is no need to represent this symmetric portion by transactors since it is always capable of representation in terms of the conventional idealized active and passive elements. In the very first example of my original contribution I show that the admittance gyrator is separable into a symmetric network consisting of two active and one passive elements, shunted by a unilateral admittance transactor. This example is in complete accord with J. Shekel's proof and provides an excellent example of the influence of a transactive element on a symmetrical active network.

Kenilworth, A. W. KEEN.
Warwickshire.
8th February 1958.

Time Signals

SIR,—To meet the need of people interested in accurate time comparisons, the long-wave stations (e.g., GBR, 16 kc/s; GBZ, 19·6 kc/s) in the U.K. have since October 1950 transmitted mean-time seconds dots with prolonged marks at minute intervals during the times 0955–1000 and 1755–1800 U.T. The much older rhythmic time signals have been continued within the periods 1001–1006 and 1801–1806 U.T., and the reason of this letter is to suggest that these older signals now serve no useful purpose and can be abandoned.

It appears to me that all interested parties are adequately served by the mean-time seconds on long waves, by the precision transmissions of MSF, WWV, etc., and by the 6-dot seconds on many B.B.C. carriers.

If your readers are in agreement with this view the way is clear to achieve at least one economy !

Marconi's Wireless Telegraph Co. Ltd., NORMAN LEA
Great Baddow, Essex.
30th January 1958.

'Earthed' or 'Common' ?

SIR,—I wish to call to your attention a term used in your journal which leads to confusion. In April, the article by H. Paul Williams and, in July, the article by W. Guggenbühl and M. J. O. Strutt, used the term 'earthed' in place of the better word 'common'. The

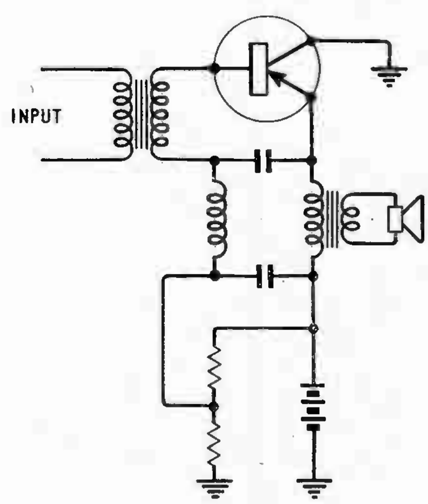


figure shows a circuit diagram of a transistor power-amplifier circuit in which the collector is 'earthed' or 'grounded' for both the supply and signal voltages, yet the transistor is operated common emitter. This circuit has been used in order that the d.c. resistance of the primary of the output transformer can serve for temperature stabilization of the emitter current, with good heat conduction from the transistor collector to the 'earth'.

Marlton, N.J., U.S.A. WINFIELD R. KOCH
7th January 1958.

Numerical Extraction of Roots

SIR,—The method given by 'Computer' in your January issue is familiar, but there is an alternative more powerful method which deserves to be better known. This was given by Mr. W. B. Jordan in a short note in *Mathematical Tables and Aids to Computation*, July 1951, Vol. 5, page 183.

If it is required to solve numerically an equation $F(y) = 0$, and a trial solution is y_1, then a much better approximation is

$$y_2 = y_1 - \frac{2\,F(y_1)\cdot F'(y_1)}{2\,F'^2(y_1) - F(y_1)\cdot F''(y_1)}.$$

To extract the square root of a number N, $F(y) = y^2 - N = 0$; after taking the derivatives, substituting and simplifying, we obtain the formula

$$y_2 = y_1 \times \frac{y_1^2 + 3N}{3\,y_1^2 + N}$$

This is reasonably convenient for computation using a desk calculator.

Taking the first example given in the article, if $y_1 = 1{\cdot}4$, as a first approximation to the square root of 2, then $y_2 = 1{\cdot}4142132$, which shows an error of 4 in the eighth significant figure compared with Computer's error of 1 in the fifth. This is a considerable improvement in accuracy without any corresponding increase in calculation. (It is worth remarking that if we take $y_1 = 1$, admittedly a very rough shot, Mr. Jordan's method yields $y_2 = 1{\cdot}4$ comparing very favourably with the value of $1{\cdot}5$ obtained by Computer's method.)

Mr. Jordan quotes as an example the extraction of the square root of 10, taking 3 as the first approximation. The second approximation has an error of 1 in the fifth significant figure, which is quite creditable, but the third approximation is correct in the fifteenth decimal place.

In practice, the first approximations are often obtainable from four- or five-figure tables, possibly by interpolation, in which case a single application of the formula usually gives adequate accuracy.

Firth Cleveland Instruments Ltd., D. T. BROADBENT
Pontypridd, Glamorgan.
20th January 1958.

New Books

Fundamentals of Electron Devices

By KARL R. SPANGENBERG. Pp. 505 + xii. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 75s.

The number of books on semiconductor devices that has been written since the discovery of the transistor ten years ago appears to be of the same order as the total number on thermionic valves in their fifty years' existence. In some of the former, hardly more than a passing reference is made to the vacuum devices, in spite of the similarity of their applications. It is therefore very welcome that Professor Spangenberg, who is well known for his earlier classical book on vacuum tubes, should have prepared a book for undergraduate and graduate study which treats both types of device on an equal footing, showing the common basis wherever possible.

Some criticism can be made of the title of the book, which should more correctly be 'Fundamentals of Lower Frequency Electron Devices', since no mention is made of microwave devices. The various facets of the theory of valves such as klystrons, magnetrons and travelling-wave tubes, whose modes of operation depend on transit time, and often on special internal circuits, are just as much fundamentals as the theory of space-charge control. In fact, in dealing only with diode and grid-controlled valves, emphasis is placed on those types which are most likely to be superseded by transistors in the future, except possibly where high voltage and high power are required.

The book deals admirably with its chosen field. The first third is devoted to the physics of the devices, covering the fundamental particles involved, their motion in electric and magnetic fields, the properties of atoms and the consequent properties of conductors, semiconductors, and insulators, junction effects, and electron emission. The remainder is concerned with the devices themselves, their characteristics, and applications, including one chapter on the various forms of photo-cell, emissive, conductive and voltaic.

It is in the handling of equivalent circuits, and in the treatment of small and large signal amplifiers and oscillators that the advantage to the student of the unified treatment appears. The slightly greater emphasis placed on the transistor is justified in view of its somewhat more complicated conditions of operation.

The text is written in a very readable style, the mathematical derivations of some of the basic relationships being carried out in a series of appendices. The rationalized system of m.k.s. units is used for the most part, but it would seem a pity from the student's point of view that occasionally the author should feel it necessary to slip into other units for convenience in visualization. At least the new generation should be taught to think directly in m.k.s. units. Equally, what seem to be thinly-veiled doubts about the use of rationalization are out of place in such a book.

These are, however, minor criticisms of a book that is well written and well produced. It can be thoroughly recommended to the student of electronics. W.H.A.

The Reproduction of Colour

By R. W. G. HUNT, B.Sc., Ph.D., D.I.C., A.R.C.S., F.R.P.S. Pp. 204; 11 colour plates. Fountain Press, 46-47 Chancery Lane, London, W.C.2. Price 63s.

Based on lectures given at the Royal Institution in 1953, Dr. Hunt's admirable book gives a clear account of the physical principles of colour mixing, and also discusses the effects of the result from the aesthetic point of view. The early chapters explain the basis of the trichromatic reproduction schemes, both additive (as used in colour television) and subtractive (with a very full account of the Kodachrome process).

The chapter on visual appreciation emphasises that the real aim is to produce a pleasing and satisfying picture. The viewer will accept anything that is supposed to be grass as being green, and the 'correct' green, and it might almost seem that a reproduction which he is happy about is good enough. But the job must, of course, be done as accurately as possible, and a great deal of labour has been expended in developing 'correction' processes which compensate for the imperfections of pigments.

There is a short and useful account of the transmission of colour television signals, and the impact of electronics on the printing process itself, which seems an even newer development (see the article on "Colour Printing", *Electronic and Radio Engineer*, Jan. 1958, p. 26), is given the prominence that its ingenuity deserves. The Hardy and Wurzburg method of producing colour-corrected dot images, and the Time-Springdale colour scanner for extracting separation negatives from a single transparency, are explained in some detail. When one considers the number of linear equations that have to be solved in the latter, not only simultaneously but also continuously, it is not surprising to learn that an electronic computer is included in the apparatus.

The colour plates, which merit inclusion for their beauty alone, illustrate the results obtained by different processes and corrections.

The author, who is on the research staff of Messrs. Kodak Ltd., has produced a book which can be recommended as a most valuable guide on basic principles as well as their most recent applications. G.R.N.

The Exploration of Space by Radio

By R. HANBURY BROWN, D.I.C., A.M.I.E.E., F.R.A.S., and A. C. B. LOVELL, O.B.E., F.R.S. Pp. 207 + xii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 35s.

Only two chapters in this book deal with the radio techniques of radio astronomy. The authors say that "the general reader may find some sections in these chapters somewhat difficult". To the radio engineer, however, they are the easiest and it is rather hard to imagine the man in the street grappling with the others. Although they may not require specialized knowledge, they do need a good general background of physics and some small understanding of mathematics. By general reader, the authors probably do not mean the man in the street so much as the non-radio physicist, however.

The book has eleven chapters of which the first provides a general astronomical background, while the second and third deal with the properties of radio waves and radio techniques. The last chapter is descriptive of the Jodrell Bank radio telescope. The remaining seven chapters, which form the bulk of the book, deal with the results of radio-astronomical observations. The whole field is covered from observations of extra-galactic noise to radar echoes from the moon.

To the expert radio astonomer, the treatment probably appears simplified. To the radio engineer it will appear quite detailed and almost more than adequate to satisfy his interest. W.T.C.

High Quality Sound Reproduction

By JAMES MOIR, M.I.E.E. Pp. 591 + xiv. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 70s.

The merit of this book is that it gives in one volume a complete general picture of the problems involved in sound reproduction, and it does put these problems in perspective so that the reader gets a good idea of their relative importance. It is intended for the professional engineer and for the knowledgeable amateur, but it is not a designer's book in the ordinary meaning of the term. In spite of its size, it is not nearly detailed enough for the designer.

The first three chapters cover the characteristics of sounds, the ear and the hearing system. An attempt is then made to draw up performance specifications for apparatus in terms of bandwidth, distortion levels, and so on. The next four chapters cover microphones and mixers, gramophone records and magnetic recording. Succeeding chapters are entitled: Voltage Amplifiers, Power Amplifiers, Output Transformers, Negative Feedback, Tone-Control Circuits, Rectifier Circuits and Dividing Networks. Chapters 16 and 17 cover loudspeakers and the acoustic problem, and the book concludes with a discussion of the special requirements of the cinema and of stereophony.

The treatment is by way of general discussion with very little mathematics. It gives the newcomer to 'sound' a very good general introduction to the subject and a wealth of related fact and opinion. Since the result is subjective and perfection is unattainable in a monaural system, there is plenty of room for opinion about what is best and it is valuable to have informed opinion so clearly presented. Although the book is one for the beginner in 'sound', it is not for the newcomer to electronics. A reasonable standard of prior knowledge of this is assumed. W.T.C.

Selection and Application of Metallic Rectifiers

By STUART P. JACKSON. Pp. 326. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 60s.

Chapters on the application of metal rectifiers, rectifier circuits, complex loads and filters, selecting the correct type of rectifier, selenium, copper oxide, magnesium copper sulphide, titanium dioxide, germanium and silicon rectifiers, junction rectifier theory and its application to germanium and silicon rectifiers, battery charging, electroplating supplies, industrial power supplies and their applications, magnetic amplifiers, and special rectifier circuits.

Analytical Design of Linear Feedback Controls

By G. C. NEWTON, L. A. GOULD and J. F. KAISER. Pp. 419. John Wiley & Sons Inc., available in the U.K. from Chapman & Hall, 37 Essex Street, London, W.C.2. Price 96s.

The authors are in the electrical engineering department of the Massachusetts Institute of Technology. The book is intended for post-graduate workers.

MEETINGS

I.E.E.

10th March. "The Industrial Designer—An Ally or Enemy of the Engineer ? ", discussion to be opened by Misha Black, O.B.E.

19th March. "Efficiency and Reciprocity in Pulse-Amplitude Modulation: Part 1, Principles", by K. W. Cattermole, B.Sc.; Part 2, "Testing and Applications", by J. C. Price, B.Sc.; "Transistor Pulse Generators for Time-Division Multiplex", by K. W. Cattermole, B.Sc.

20th March. "Storage and Manipulation of Information in the Brain", by R. L. Beurle, Ph.D., B.Sc.(Eng.).

25th March. "The Atomic Clock", by L. Essen, D.Sc., Ph.D.

31st March. "Future Radiocommunication Methods for Civil Aircraft", by W. E. Brunt.

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

Brit. I.R.E.

18th March. A discussion on the Technical Committee's Report: "Recommended Method of Expressing Electronic Measuring Instrument Characteristics: 1. A.M. or F.M. Signal Generators".

26th March. "Electronics in Medicine", by R. F. Farr, B.Sc.

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

Institute of Physics

18th March. "Microwave Ferrites", by L. A. Thomas, to be held at 5.30 at the Institute of Physics, 47 Belgrave Square, London, S.W.1.

R.S.G.B.

21st March. "The Junction-Type Transistor and its Application to Short-Wave Radio", by E. Wolfendale, and L. E. Jansson, to be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, commencing at 6.30.

Radar and Electronics Association

12th March. "The Jodrell Bank Radiotelescope in Action", by Professor A. C. B. Lovell, O.B.E., B.Sc., Ph.D., to be held at 7.30 in the Anatomy Theatre, University College, Gower Street, London, W.C.1.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for January* 1958

Date 1958 January	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	− 2	N.M.
2	− 2	N.M.
3	− 2	N.M.
4	− 2	N.M.
5	− 2	0
6	− 2	0
7	− 2	0
8	− 2	0
9	− 2	N.M.
10	− 2	0
11	− 2	0
12	− 2	+ 1
13	− 2	+ 1
14	− 2	+ 2
15	− 2	+ 2
16	− 2	+ 2
17	− 2	+ 1
18	− 2	+ 1
19	− 2	+ 1
20	− 1	+ 3
21	− 1	+ 3
22	− 1	+ 4
23	− 1	+ 4
24	− 1	N.M.
25	− 1	N.M.
26	− 1	N.M.
27	− 2	+ 1
28	− 2	+ 2
29	− 2	+ 1
30	− 2	+ 2
31	− 2	+ 3

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

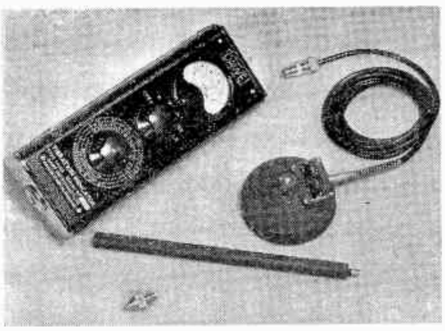
New Products

Vibration Indicator

An inexpensive, simple, yet accurate vibration indicator is available for routine testing and vibration surveys, for which elaborate equipment is not required.

Known as the Type 1414 vibration indicator, the unit consists of the indicating instrument itself and a transducer probe. The equipment is self-contained, measures $8\frac{3}{4}$ in. by $3\frac{3}{8}$ in. by $2\frac{7}{8}$ in., and weighs $2\frac{1}{2}$ lb. It is capable of giving a direct reading of peak acceleration, velocity or displacement of vibration. Provision is made for checking on the meter the condition of the h.t. and l.t. batteries.

The vibration pick-up is of the inertia-operated crystal type delivering a voltage proportional to acceleration up to a maximum value of some 10 g. The velocity and displacement readings are obtained by the



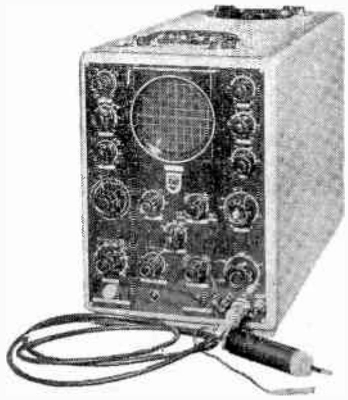
use of single and double-integrating networks. The pick-up housing is 3 in. diameter by $2\frac{1}{2}$ in. high, the weight with the 8-in. long extension probe shown in the illustration being 1 lb. The extension probe makes it possible to check vibrations in locations which are difficult to reach otherwise.

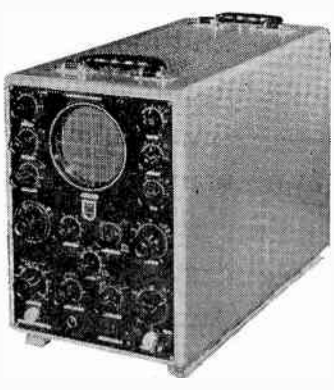
Dawe Instruments Ltd.,
99 *Uxbridge Road, London, W.*5.

Oscilloscopes

Philips Electrical Ltd. have introduced two new oscilloscopes.

Type GM5666, a direct-coupled oscilloscope designed for low-frequency investigations in the mechanical, chemical, biological





and medical fields, and type GM5662, which incorporates a wide-band Y amplifier having a slow-falling frequency characteristic and a versatile time-base. The latter is said to be particularly suitable for the observation of pulse phenomena.

Makers' Specifications:

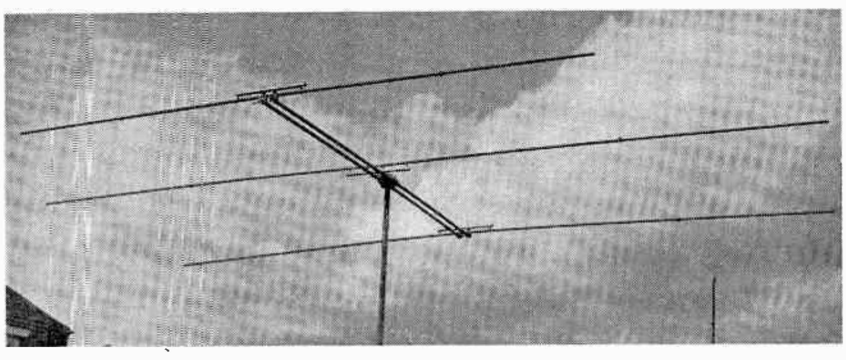
GM5666—C.R. tube: 10 cm with post-deflection acceleration. Y amplifier: D.C. —40 kc/s (−3 dB) 3 mV p-p/cm. X amplifier: D.C. —200 kc/s (−3 dB) 350 mV p-p/cm. Time-base: Frequency, 0·3 c/s–30,000 c/s; Velocity, 3 μsec/cm 0·3 sec/cm (with 10 × expansion). Trigger/sync.: Positive or negative, internal or external. Mains supply: 110–245 V, 40–100 c/s; consumption 130 W.

GM5662—C.R. tube: 10 cm with post-deflection acceleration. Y amplifier: 3 c/s–14 Mc/s (−3 dB) 50 mV p-p/cm. X amplifier: D.C. —800 kc/s (−3 dB) 600 mV p-p/cm. Time-base: Frequency, 10 c/s–800 kc/s; Velocity, 0·08 μsec/cm–10 μsec/cm (plus 5 × expansion). Trigger/sync.: Positive or negative, internal or external. Mains supply: 110–245 V, 50–100 c/s; consumption 230 W.

Philips Electrical Ltd.,
*Century House, Shaftesbury Avenue, London, W.C.*2.

Multi-Band Aerial

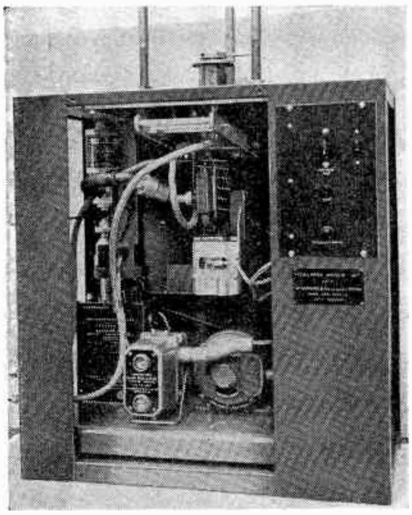
A beam aerial for the 10, 15 and 20-metre amateur radio bands is available as a kit of parts. It is claimed to be easily erected, to require no elaborate supporting tower (the weight is only 35 lb) and to require no adjustments when changing from one band to another.



Northern Aluminium Co. Ltd.,
Banbury, Oxfordshire.

C.R.T. Re-Pumping Unit

This unit has been specially developed for evacuating cathode-ray tubes which have been repaired for re-sale. The outfit consists of a 'Speedivac' 1S50 rotary vacuum pump displacing approximately 1·7 cubic feet per minute, backing a type 203 oil diffusion pump capable of attaining a final vacuum of 5×10^{-6} mm Hg. The diffusion pump is fitted with a thermosnap switch for protection if the cooling water supplies should fail and a fluid economizer is mounted in the backing line to reduce the amount of oil lost from the diffusion pump when the system is open to atmosphere. A magnetic



isolation valve is mounted above the rotary pump to isolate this from the rest of the system, thus enabling the system to be left under vacuum when the unit is switched off. Other safety switches and non-return valves are supplied for the complete protection of the plant in the event of power failure.

The top surface of the cabinet is made from heat-resistant material and the neck-gripper of the cathode-ray tube is water-cooled so that the tube may be baked in an

oven prior to sealing. The complete plant is entirely self-contained and occupies a floor space of approximately 16 in. × 28 in. the cabinet being 30 in. high.
Edwards High Vacuum Ltd.,
Manor Royal, Crawley, Sussex.

Relays with Heavy-Duty Contacts

Magnetic Devices Ltd. state that their compact Series 590/596 relays can now be supplied with heavy-duty contacts. These contacts are rated at 15 amperes with a single-pole normally open arrangement.

Since they are only $1\frac{1}{2}$ in. long by $1\frac{1}{8}$ in. wide by $1\frac{3}{4}$ in. deep approximately, these relays are suitable for use where space is limited.

Contact arrangement: single pole normally open (double break); contact

rating 15 A resistive load at 250 V a.c. or 30 V d.c.; maximum operating voltage of the Series 500 relay is 250 V 50 c/s with a coil rating of 4 VA, and that of the Series 596 relay is 150 V d.c., with a coil rating of 1 W.
Magnetic Devices Ltd.,
Exning Road, Newmarket, Suffolk.

L.F. Phase Meter

The Airmec phase meter Type 206 has been designed for measuring the gain, attenuation and phase-shift of any four-terminal network operating in the frequency range 20 c/s to 100 kc/s. Phase is indicated directly on a six-inch meter having four scales, and gain or loss values are indicated by the difference between two attenuator settings.

Two cathode-follower probes are provided

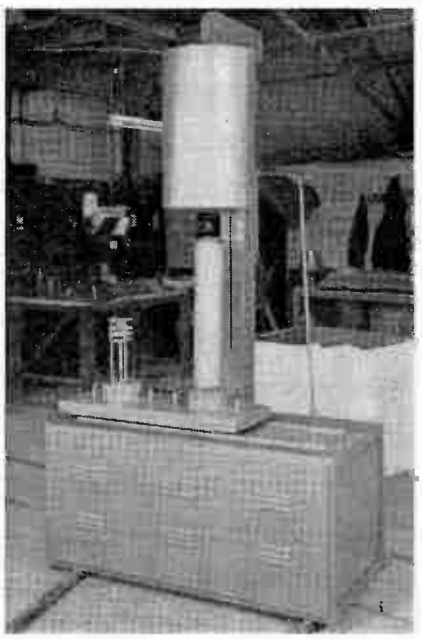
with input impedances of 12 MΩ. The range of input level on either channel is 1 mV to 1 V and insertion losses and gains of up to 60 dB can be measured.

The circuit consists of two identical amplifiers with calibrated attenuators. Each amplifier is followed by a trigger circuit which generates a pulse at the point where the signal crosses the zero axis. These pulses are then passed to the metering circuit where they determine the relative duty-cycles of two valves forming a flip-flop circuit. The differential of these duty-cycles is indicated on the meter as phase measurement. The instrument is relatively insensitive to waveform distortion. The accuracy of phase measurement is given as ± 2° (100 c/s–20 kc/s) and better than ± 5° over the rest of the frequency range.
Airmec Ltd.,
High Wycombe, Bucks.

Hydrogen Bell Furnace

Small electric furnaces are specially designed and manufactured for the electronic industry by Royce Electric Furnaces Ltd.

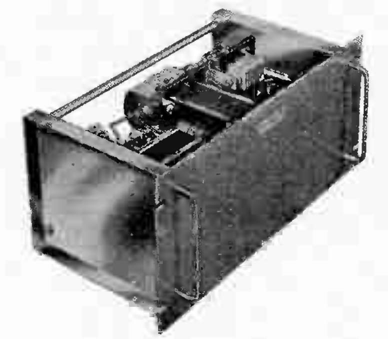
The furnace illustrated has a work space of 5 in. diam. × 13 in. high, and is suitable for operation at temperatures up to 1,000°C. Heating elements are of 80/20 nickel chromium and operate from the low-voltage secondary of a step-down transformer, permitting the use of a heavy-section element.



For operation with protective atmospheres, heat-resisting steel gas-tight retorts are provided. Gas is introduced through the base.
Royce Electric Furnaces Ltd.,
Sir Richards Bridge,
Walton-on-Thames, Surrey.

Rack-Mounted Stabilizers

Series BMVR and TCVR distortionless a.c. automatic voltage stabilizers are now available from Claude Lyons in open rack-



mounted form suitable for incorporation in customers' equipment. Prices and electrical characteristics are stated to be the same as for cabinet-mounted versions.

The TCVR is recommended by the makers where speedy response to mains variations is essential. Its correction rate is specified as 40 volts per second.
Claude Lyons Ltd.,
Valley Works, Ware Road, Hoddesdon, Herts.

Silicon Junction Rectifiers

The range of 0·5-A and 1-A silicon junction rectifiers manufactured by S.T.C. has been extended by adding a 400-V p.i.v. type to the 0·5-A series and three new types with p.i.vs. of 200 V, 300 V, and 400 V to the 1-A series. Both 0·5-A and 1-A types are now available with p.i.v. ratings of 50 V, 100V, 150 V, 200 V, 300 V and 400 V.

The photograph shows 1-A rectifiers. These must be provided with a cooling fin of at least 1 in. × 1 in., but the current rating at high temperatures can be increased

if larger fins are used. The rectifiers are hermetically-sealed by glass-to-metal seals and provided with a threaded mounting stud.
Standard Telephones & Cables Ltd.,
Rectifier Division,
Edinburgh Way, Harlow, Essex.

Solvent for Synthetic Resins

De-Solv 292 is described as a powerful solvent which makes possible the reclamation and salvaging of components sealed in epoxy or polyester resin casings. By immersion and soaking in De-Solv 292, the components may be readily removed.
Oxley Developments Co. Ltd.,
Ulverston, Lancs.

Abstracts and References

COMPILED BY THE RADIO RESEARCH ORGANIZATION OF THE DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH AND PUBLISHED BY ARRANGEMENT WITH THAT DEPARTMENT

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publisher concerned.*

	Page A		Page A
Acoustics and Audio Frequencies	33	**Measurements and Test Gear**	42
Aerials and Transmission Lines	33	**Other Applications of Radio and Electronics**	43
Automatic Computers	34	**Propagation of Waves**	43
Circuits and Circuit Elements	35	**Reception**	44
General Physics	37	**Stations and Communication Systems**	44
Geophysical and Extraterrestrial Phenomena	38	**Subsidiary Apparatus**	46
Location and Aids to Navigation	39	**Television and Phototelegraphy**	46
Materials and Subsidiary Techniques	39	**Transmission**	46
Mathematics	42	**Valves and Thermionics**	47
Miscellaneous	48		

ACOUSTICS AND AUDIO FREQUENCIES

534.22-14-8 : 546.212 **657**

The Measurement of the Velocity of Sound in Doubly Distilled Water Containing Differing Amounts of Gas and at Various Temperatures.—H. Markgraf. (*Hochfrequenztech. u. Elektroakust.*, March 1957, Vol. 65, No. 5, pp. 169–173.)

534.23 : 621.396.677.3 **658**

Arrays with Constant Beam Width over a Wide Frequency Range.—D. G. Tucker. (*Nature, Lond.*, 7th Sept. 1957, Vol. 180, No. 4584, pp. 496–497.) A linear array of omnidirectional transducers is considered. Directivity is achieved in all planes containing the line of the array by the synthesis of directional patterns from elementary $(\sin x)/x$ patterns using delay lines to produce these patterns at various angles of deflection from the normal. See also 336 of February (Berman & Clay).

534.232 **659**

The Directivity and Impedance of Artificially Compensated Cylindrical Acoustic Transducers.—M. Federici. (*Ricerca sci.*, June 1957, Vol. 27, No. 6, pp. 1826–1838.) Sound radiation from a source consisting of cylindrical elements vibrating radially with variable phase is investigated (see also 3469 of 1955). By suitably arranging and phasing a number of transducer elements sound can be beamed in the direction of the cylinder axis. The characteristics of this type of source are calculated. The acoustic impedance of a cylinder of infinite length is expressed as the ratio of two Hankel functions of the first order.

534.844 : 681.84.087.7 **660**

Reverberation Chambers for Broadcasting and Recording Studios.—M. Rettinger. (*J. audio Engng Soc.*, Jan. 1957, Vol. 5, No. 1, pp. 18–22.) The design and construction of reverberation chambers are considered.

534.861 : 534.784 **661**

Acoustic Conditions for Broadcast Transmission of Speech.—I. Malecki. (*NachrTech.*, June 1957, Vol. 7, No. 6, pp. 267–273.) The conditions in a transmission channel which are necessary for the correct reproduction of the acoustic structure of speech are determined, and the influence of acoustic conditions at the microphone on transmission quality are investigated. The importance of allowing for differences in language characteristics is considered.

621.395.623.7 **662**

Mechanical Crossover Characteristics in Dual-Diaphragm Loudspeakers.—A. B. Cohen. (*J. audio Engng Soc.*, Jan. 1957, Vol. 5, No. 1, pp. 11–17.) The role of diaphragm shape and material in controlling mechanical crossover characteristics is discussed. Frequency response curves which show the effect of an auxiliary diaphragm are given for conical and curvilinear diaphragms.

621.395.623.7 : 537.523.3 **663**

The Corona-Wind Loudspeaker.—G. Shirley. (*J. audio Engng Soc.*, Jan. 1957, Vol. 5, No. 1, pp. 23–31.) An experimental model is described and illustrated, and problems arising in the construction of a practical loudspeaker, with particular reference to commercial production, are discussed. See also 659 of 1956 (Tombs) and *Electronics*, 1st July 1957, Vol. 30, No. 7, pp. 198, 200 (Tombs et al.).

621.395.625.3 **664**

Tape Storage Problems.—F. Radocy. (*J. audio Engng Soc., Jan.* 1957, Vol. 5, No. 1, pp. 32–35.) The reduction of adverse effects in storing magnetic recording tapes is discussed.

AERIALS AND TRANSMISSION LINES

621.315.212 : 621.372.54.001.2 **665**

The Design of T-Stub Lines.—P. Vielhauer. (*NachrTech.*, June 1957, Vol. 7, No. 6, pp. 241–243.) The dimensions of a T-stub line for use in vestigial-sideband filters are calculated.

621.372.2 **666**

Synthesis of Lumped-Parameter Precision Delay Line.—E. S. Kuh. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1632–1642.) The two parts of the problem of designing delay lines are (*a*) to provide the required time delay and bandwidth with the least complicated network and (*b*) to have a good time response. The network obtained is a tandem connection of a low-pass ladder and an all-pass bridge structure.

621.372.2 : 621.372.43 667

A New Wide-Band Balun.—W. K. Roberts. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1628–1631.) This balun has low loss and excellent impedance characteristics over a 3 : 1 frequency band, without any adjustments. The bandwidth increase is obtained by the use of a $\lambda/4$ transmission-line section which is placed inside one of the balanced arms.

621.372.8 : 621.372.2 668

A Method for Calculating Propagation Constants in Waveguides with Imperfectly Conducting Walls.—L. N. Loshakov. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 8–11.) An approximate method and a numerical example of its application are given. See also 1273 of 1957.

621.372.831 669

Designing Tapered Waveguide Transitions.—B. J. Migliaro. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 183–185.) "A procedure for designing well-matched transitions between different waveguides where the guide wavelength varies along the taper. Simple graphical method employs data presented here for double-ridge-to-rectangular transitions. Procedure may also be used for rectangular-to-rectangular or ridge-to-ridge tapered transitions with corresponding accuracy."

621.372.85 670

Some Waveguides with Discontinuous Structure.—M. Jouguet. (*C. R. Acad. Sci., Paris*, 17th July 1957, Vol. 245, No. 3, pp. 297–298.) The propagation of e.m. waves in a waveguide having thin insulating sheets at right angles to and parallel with its axis is considered for negligible and appreciable dielectric losses.

621.396.67 : 621.372.43 : 621.397.62 671

A Diplexer Two-Set Coupler.—H. Harris. (*Radio TV News*, Sept. 1957, Vol. 58, No. 3, pp. 69, 176.) This is a network for supplying two television receivers from a common aerial, and is the electrical equivalent of two $\lambda/4$ transmission lines.

621.396.67.029.62 : 621.3.015.1 672

V.H.F. Voltage Distribution in Communal Aerial Installations.—A. Fiebranz. (*Nachrichtentech. Z.*, July 1957, Vol. 10, No. 7, pp. 349–356.) Approximation formulae, curves and details of other aids are given to assist in the installation of distribution systems for blocks of flats, etc.

621.396.677 : 523.16 : 523.72 673

Tests on a Model of an Aerial for Use in Radio Astronomy.—G. C. Corazza & G. Francini. (*Ricerca sci.*, June 1957, Vol. 27, No. 6, pp. 1777–1786.) Radiation diagrams were obtained for a 1 : 15 scale model of an array of helical aerials for solar observations at 200 Mc/s. Tests on the model were made at 3 kMc/s; the equipment is briefly described.

621.396.677.029.63 : 621.397.62 674

Television Aerials for Bands IV and V.—F. R. W. Strafford. (*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 11–13.) The requirements of television aerials suitable for these frequency bands are discussed and attention is drawn, with the aid of performance figures taken at 654 Mc/s, to the advantages of the corner reflector with 'bow-tie' dipole.

621.396.677.3 : 534.23 675

Arrays with Constant Beam Width over a Wide Frequency Range.—Tucker. (See 658.)

621.396.677.71 : 621.397.61 676

The Aerial of the Television Transmitter Feldberg/Schwarzwald.—H. Mack. (*Nachrichtentech. Z.*, July 1957, Vol. 10, No. 7, pp. 356–361.) Design and constructional details of a two-section slotted-cylinder aerial structure about 105 ft. high, on top of a 135-ft. tower.

621.396.677.73 677

Ridge Vane Antenna Provides Constant Beam Width.—W. A. Scanga. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 196, 198.) A pair of vanes, extending from the open end of a ridge waveguide, act as an end-fire aerial which has good wideband properties.

621.396.677.8 : 551.578 678

The Effect of Atmospheric Precipitation on the Electrical Properties of Wire-Mesh Surfaces.—V. K. Paramonov. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 12–20.) The effect of icing on the reflecting properties of mesh surfaces is considered. Formulae are derived for determining the coefficient of transmission through a mesh, the conductors of which are covered with a uniform layer of ice. Experimental results are given, confirming the correctness of the method proposed.

AUTOMATIC COMPUTERS

681.142 679

A Multipurpose Electronic Switch for Analogue Computer Simulation and Autocorrelation Applications.—N. D. Diamantides. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 197–202. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 680

An Error Analysis of Electronic Analogue Computers.—V. A. Marsocci. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 107–212. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 681

A Wide-Band Multiplier using Crystal Diodes.—M. E. Fisher. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 580–585.) The principle of quarter-square multiplication is used, the squaring being performed by networks of biased Ge diodes in the feedback paths of standard computing amplifiers. These networks have phase shifts of less than 1° at 50 kc/s and the errors of the multiplier are less than 0·3% of maximum output.

681.142 682

A New Storage Element Suitable for Large-Sized Memory Arrays—the Twistor.—A. H. Bobeck. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1319–1340.) Use is made of the Wiedemann effect whereby a torsion applied to a magnetic wire shifts the preferred direction of magnetization into a helical path. Three modes are suggested in using the effect for storage cells: (*a*) the coincidence of circular and longitudinal magnetic fields inserts information into the wire as a polarized helical magnetization; (*b*) operation similar to the coincident-current toroid with the wire acting as its own sensing winding; (*c*) the wire is not twisted but the screw-sense of the flux path is related to the current polarities. Equations relating to the switching performance of a twistor are derived and a description of an experimental 320-bit array is given.

681.142 683

The IBM 705 EDPM Memory System.—R. E. Merwin. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 219–223. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 : 537.227 684

A New Type of Ferroelectric Shift Register.—J. R. Anderson. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 184-191. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 : 538.221 685

Ferroresonant Circuits for Digital Computers.—C. B. Newport & D. A. Bell. (*J. Brit. Instn Radio Engrs*, Nov. 1957, Vol. 17, No. 11, pp. 619–630.) An analysis of bistability in ferroresonant circuits, and the practical application to high frequency operation (present limit 1 Mc/s). Circuits are shown for a shift register, 3-stage binary counter and 2-input logical adder.

681.142 : 621.314.7 686

Logic Circuits for a Transistor Digital Computer.—G. W. Booth & T. P. Bothwell. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. EC-5, No. 3, pp. 132–138. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 255.)

681.142 : 621.314.7 687

Junction-Transistor Switching Circuits for High-Speed Digital Computer Applications.—G. J. Prom & R. L. Crosby. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 192–196. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 : 621.317.729 688

An Electrolytic Tank as an Analogue Computing Machine for Factorizing High-Degree Polynomials.—S. K. Ip. (*Quart. J. Mech. appl. Math.*, Aug. 1957, Vol. 10, Part 3, pp. 369–384.) By means of simple measurements all the roots of a polynomial as high as the 16th degree can be located to within about 4%.

681.142 : 621.373.43 689

Pulse Generator and High-Speed Memory Circuit.—Z. Bay & N. T. Grisamore. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. EC-5, No. 4, pp. 213–218. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 574.)

681.142 : 621.398 **690**

A Simple Shaft Digitizer and Store.—A. Tiffany. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 568–574.) A shaft position encoder of simple construction for analogue-to-digital conversion is described.

CIRCUITS AND CIRCUIT ELEMENTS

621.318.57 : 621.318.134 **691**

Some Applications of Square-Loop Ferrite Cores to Telecommunication Switching.—W. Six & R. A. Koolhof. (*Philips Telecommun. Rev.*, Sept. 1957, Vol. 18, No. 3, pp. 105–124.) Slightly abridged version of paper published in *Proc. Instn elect. Engrs*, Part B, July 1957, Vol. 104, Supplement No. 7, pp. 491–501.)

621.318.57 : 621.385.83 **692**

High-Speed Gating Circuit using the E80T Beam-Deflection Tube.—L. Sperling & R. W. Tackett. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 59–63. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.372 : 621.3.018.41 **693**

The Approximation of Special Frequency Laws by means of Tables of Frequency Functions.—H. Dobesch. (*Hochfrequenztech. u. Elektroakust.*, March 1957, Vol. 65, No. 5, pp. 159–164.) An approximation method facilitating the synthesis of two-pole and quadripole networks is described.

621.372.2 : 512.831 **694**

Admittance, Impedance and Scattering Matrices.—G. C. Corazza & F. Serracchioli. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, No. 3, pp. 336–342.) The concepts and their interrelation are defined.

621.372.4/.5 : 621.317.729.1 **695**

Analysis and Synthesis of Electrical Circuits by means of an Electrolyte Tank.—M. L. D'Atri & U. Pellegrini. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, No. 3, pp. 305–326.) Experimental procedure for solving network problems is described.

621.372.412 : 549.514.51 **696**

Effects of X-Ray Irradiation on the Frequency/Temperature Behaviour of AT-Cut Quartz Resonators.—A. R. Chi. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1524–1529.) Results for resonators fabricated from natural quartz, synthetic quartz grown on several different cuts of seed plates and synthetic quartz containing Al or Ge impurity, are reported.

621.372.412.088.33 : 621.317.3 **697**

Tolerances of Quartz Crystals for Filters and their Measurement.—C. Kurth & R. Miczynski. (*NachrTech.*, June 1957, Vol. 7, No. 6, pp. 244–249.) Methods for measuring series resonance and crystal inductance are indicated.

621.372.5 : 621.375.13 **698**

Negative Impedances and Gyrators.—J. Gensel. (*NachrTech.*, June 1957, Vol. 7, No. 6, pp. 249–256.) Survey of characteristics and classification of the various types of impedance inverter and gyrator. Their realization in the form of transistor circuits is discussed. See e.g. 2867 of 1955 (Bogert).

621.372.5.016.35 **699**

Stability Criteria for Linear Systems and Realizability Criteria for *RC* Networks.—A. T. Fuller. (*Proc. Camb. phil. Soc.*, Oct. 1957, Vol. 53, Part 4, pp. 878–896.) A new set of stability criteria in linear systems is derived; about half of the Hurwitz criteria can be neglected when certain of the coefficients of the characteristic equation are positive. The conditions for realizability of *RC* networks are closely related to the stability and aperiodic criteria and are given in the form of polynomial coefficients.

621.372.54 **700**

Basic Properties and Characteristics of a Synchronous Filter.—N. K. Ignat'ev. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 59–71.) The operation of integrating devices used for improving signal/noise ratio is discussed. A simple circuit with a long open-circuited line as store is analysed.

621.372.54 **701**

The Development of a New Method of Circuit Analysis in Ladder Networks.—L. F. Coker. (*Commun. & Electronics*, May 1957, No. 30, pp. 158–160.) The method outlined is applied to networks consisting of tandem-connected L sections. Calculations are simpler than in conventional methods.

621.372.54 **702**

Explicit Formulas for Tschebyscheff and Butterworth Ladder Networks.—L. Weinberg. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1155–1160.) A new set of simple formulae has been found for the element values in a Tchebycheff or Butterworth ladder network, which apply when the degree of the denominator of the transfer function is odd and the reflection coefficient has zeros alternating in the left and right half-planes.

621.372.54 + 621.373.5] : 621.314.7 **703**

***RC* Filters and Oscillators using Junction Transistors.**—N. Sohrabji. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 606–608.)

621.372.54 : 621.376.3 : 621.3.018.78 **704**

The Distortion of Frequency-Modulated Oscillations caused by *RC* Networks. General Remarks on Distortion Factors of Filters not Accurately Tuned to the Carrier, which have a Skew-Symmetric Phase Characteristic and a Mirror-Symmetric Amplitude Characteristic.—E. G. Woschni. (*Hochfrequenztech. u. Elektroakust.*, March 1957, Vol. 65, No. 5, pp. 165–169.)

621.372.54 : 621.396.621 **705**

Mutual Correlation of Fluctuation-Type Interference at the Output of Frequency Filters.—M. V. Maksimov. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 28–38.) The method is discussed of determining the interference correlation function for two filters forming the loads of a two-frequency-channel receiving system with or without detector.

621.373.029.6 : 538.569.4 **706**

Proposal for a Solid-State Radio-Frequency Maser.—J. Itoh. (*J. phys. Soc. Japan*, Sept. 1957, Vol. 12, No. 9, p. 1053.) A brief comment on conditions which are necessary to construct a maser using Zeeman levels of the pure quadrupole spectrum.

621.373.029.64 : 538.569.4 **707**

Characteristics of the Beam-Type Maser: Part 1.—K. Shimoda. (*J. phys. Soc. Japan*, Sept. 1957, Vol. 12, No. 9, pp. 1006–1016.) Velocity distribution of molecules in a beam-type maser is analysed by an approximate method. The average velocity of the molecules is much less than the most probable velocity when the power level and focusing voltage are low. The relations between amplitude, focuser voltage and frequency agree well with experimental results to be detailed in Part 2.

621.373.029.64 : 538.569.4 **708**

Solid-State Oscillator for Microwave Frequencies.—(*Engineer, Lond.*, 8th March 1957, Vol. 203, No. 5276, p. 389.) An experimental solid-state device using gadolinium ethyl sulphate, an ionically bound paramagnetic salt, as the active element is described. See 2108 of 1957 (Scovil et al.).

621.373.4 : 537.525 **709**

The Influence of a High-Frequency Gas Discharge on the Frequency of a Self-Excited Short-Wave Oscillator Stage.—E. Häusler. (*Z. angew. Phys.*, Feb. 1957, Vol. 9, No. 2, pp. 60–66.) Continuing earlier investigations [381 of 1955 (Häusler & Koch)] further tests were made with a 20–50-Mc/s oscillator under various conditions of discharge, coupling and magnetic field. Characteristic curves indicating a linear relation between frequency and anode voltage are obtainable. The interpretation of the observations is discussed.

621.373.42 **710**

Polyphase Oscillators.—A. S. Gladwin. (*Electronic Radio Engr*, Jan. 1958, Vol. 35, No. 1, pp. 16–24.) The frequency stability and freedom from harmonics is greater for single-phase than for symmetrical polyphase oscillators.

621.373.42 **711**

Phase Generator has Resistive Shifter.—G. E. Pihl. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 175–177.) The generator operates over a frequency range 20 c/s–20 kc/s and supplies a pair of sinusoidal voltages with an accurately known phase relation that is continuously adjustable over 360°.

621.373.42 **712**

Designing Oscillators for Greater Stability.—S. N. Witt, Jr. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 180–182.) Methods of improving the frequency stability of an oscillator which can be divided into an amplifier and a feedback network are described and examples are given.

621.373.421 **713**

Simultaneous Pulled Oscillations in a Triode Oscillator Incorporating Two Oscillatory Circuits. — Abd El-Samie Mostafa. (*Commun. & Electronics*, May 1957, No. 30, pp. 120–127.) Simultaneous oscillation at two frequencies is possible and owing to mutual pulling both frequency and amplitude modulation occur. A physical explanation of the pulling of a nonlinear oscillator by an external input signal is given. See also 398 of February (Feist).

621.373.43 **714**

Second-Order Nonlinear Systems.—L. Sideriades. (*J. Phys. Radium*, May 1957, Vol. 18, No. 5, pp. 304–311.) In these systems, where time is not explicitly shown, the four-space is split into a displacement subspace and a velocity subspace, the two being related by a hypercone in a 1–1 transformation. Singularities of integral curves and certain shock phenomena are discussed, and the analysis is applied to two types of multivibrator. See also 1369 of 1957.

621.373.43 : 681.142 **715**

Pulse Generator and High-Speed Memory Circuit.—Bay & Grisamore. (See 689.)

621.373.431.1 **716**

Cathode-Coupled Flip-Flop.—T. G. Clark. (*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 24–27.) A reliable design procedure.

621.373.431.1 **717**

Analysis of Cathode-Coupled Free-Running Multivibrator.—D. C. Sarkar. (*Indian J. Phys.*, Aug. 1957, Vol. 31, No. 8, pp. 431–439.) A quantitative equivalent-circuit analysis valid for frequencies where the interelectrode capacities can be neglected, gives results in good agreement with experiment.

621.373.431.1 : 621.318.57 **718**

High-Speed Flip-Flops for the Millimicrosecond Region.—Z. Bay & N. T. Grisamore. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. EC-5, No. 3, pp. 121–125. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 254.)

621.373.431.1 : 621.318.57 : 621.314.7 **719**

Transient Analysis of Second-Order Flip-Flops. L. M. Vallese. (*Commun. & Electronics*, May 1957, No. 30, pp. 161–166.) A mathematical analysis of the trigger sensitivity and switching speed of flip-flop circuits whose action can be represented by a second-order differential equation. The method is applied to a point-contact emitter-input transistor flip-flop.

621.373.5 : 621.317.361.089.68 **720**

A Sawtooth Crystal Calibrator.—E. L. Campbell. (*QST*, July 1957, Vol. 41, No. 7, pp. 22–24.) Circuit and construction details of a sawtooth generator triggered from a 100-kc/s crystal and giving detectable harmonics up to 50 Mc/s.

621.373.52 **721**

Feedback Coupling in Circuits with Crystal Triodes.—Ya. K. Trokhimenko. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 46–53.) Circuit design methods are described and the formulae derived are tabulated.

621.374.32 **722**

Binary-Decimal Counter Operates at 10 Mc/s.—D. E. Cottrell. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 186–189.) A design of counter which produces one output pulse for every ten input pulses is described. It uses logical 'and' gates in a feedback loop.

621.374.32 : 621.3.083.8 **723**

Statistical Fluctuations and Optimum Transfer Functions of a Counting Rate Meter.—H. Maier-Leibnitz. (*Z. angew. Phys.*, Feb. 1957, Vol. 9, No. 2, pp. 57–60.) Equations are derived to obtain the optimum response conditions, and suitable circuits are described.

621.375.2 : 621.317.733 **724**

An Amplifier for A. C. Bridges.—A. H. Allan, J. R. Gabriel & B. H. Robinson. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 597–599.) "The requirements of an amplifier for a.c. bridges working to an accuracy of one part in ten thousand or better, from power frequencies to audio frequencies, are discussed. An amplifier to meet, in part, these requirements, is described, and figures are given for its performance."

621.375.2.029.63 **725**

A U.H.F. Wide-Band Amplifier.—J. Kason. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 600–602.) Description of the design of a 500-Mc/s amplifier with bandwidth about 40 Mc/s for program distribution.

621.375.23 **726**

A Feedback Circuit Equivalence.—A. W. Keen. (*Electronic Radio Engr*, Jan. 1958, Vol. 35, No. 1, pp. 8–12.) Transfer-network representations and their transformations are used to show that the feedback in a bootstrap amplifier with shunt feedback may be considered either positive or negative without inconsistency.

621.375.3 **727**

Dynamic Core Behaviour and Magnetic-Amplifier Performance.—L. A. Finzi & D. L. Critchlow. (*Commun. & Electronics*, May 1957, No. 30, pp. 229–240.) Discrepancies between theoretical analyses and measurements of magnetic-amplifier performance are attributed to excessively simplified representations of core behaviour. Experimentally acquired knowledge of core flux characteristics is used to predict the operation of amplifier systems.

621.375.4 : 621.3.018.756 **728**

Reduction of the Distortion of Pulse Fronts in Transistor Video Amplifiers.—T. M. Agakhanyan. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 54–58.) Methods are described for reducing the distortion of leading edges caused by diffusion and other effects in the transistor base region. Compensating circuits are analysed and phase and amplitude response curves are given.

621.375.4 : 621.314.7 **729**

Compensation for Changes in Base to Emitter Voltage with Temperature.—P. Tharma. (*Mullard tech. Commun.*, May 1957, Vol. 3, No. 24, pp. 106–109.) An economical method of compensating power-transistor circuits consists in including in the emitter circuit metallic resistors having small positive temperature coefficients.

621.375.4.024 **730**

D.C. Amplifier using Transistors and a Silicon Bridge Modulator.—K. Holford. (*Mullard tech. Commun.*, June 1957, Vol. 3, No. 25, pp. 126–137.) Matched Si junction diodes are used in a bridge modulator to convert the input to a.c., which is then amplified by transistors and detected. The design is suitable for measuring d.c. inputs of 2 mV and $0 \cdot 2\ \mu$A or more.

621.375.427 **731**

Transistor Class-B Push-Pull Stages.—L. H. Light. (*Mullard tech. Commun.*, May 1957, Vol. 3, No. 24, pp. 98–101.) The relative merits of single-ended and symmetrical output stages are discussed. In general the single-ended circuit offers considerable advantages.

621.375.427 **732**

Feedback Arrangements in Transformerless Push-Pull Output Stages.—L. H. Light. (*Mullard tech. Commun.*, May 1957, Vol. 3, No. 24, pp. 102–105.) Different methods of applying negative feedback to single-ended and symmetrical class-B circuits (731 above) are discussed.

621.375.9 : 538.569.4 : 621.396.822 **733**

Measurement of Noise in a Maser Amplifier.—L. E. Alsop, J. A. Giordmaine, C. H. Townes & T. C. Wang. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1450–1451.) Measurements on an NH_3-beam maser whose cavity and input and output loads were cooled to near liquid-nitrogen temperature, yielded a noise figure of $-2 \cdot 0$ dB based on room temperature, as compared with a theoretical figure of $-2 \cdot 3$ dB.

621.375.9 : 538.569.4.029.6 **734**

Computation of Noise Figure for Quantum - Mechanical Amplifiers.—M. W. P. Strandberg. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1483–1484.) An expression is derived in terms of the physical quantities of the electromagnetic structure and of the paramagnetic salt used.

621.375.9 : 538.569.4.029.6 **735**

Theory of a Three-Level Maser.—Javan. (See 764.)

621.375.9 : 538.569.4.029.6 **736**

Gain Bandwidth and Noise in Maser Amplifiers.—A. E. Siegman. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, 1737–1738.) A theoretical study reveals that the optimum gain-bandwidth product of the two-port maser is only half that of the circulator maser for the same basic cavity. The optimum noise figure is the same if some bandwidth is sacrificed in the two-port maser. Practical considerations may, however, make the latter competitive.

621.375.9 : 538.569.4.029.6 : 621.396.822 **737**

Experimental Determination of the Noise Figure of an Ammonia Maser.—J. P. Gordon & L. D. White. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1728–1729.) An outline of the technique of measurement is given.

GENERAL PHYSICS

530.145 : 533.15 **738**

Modified WKB Approximation for Bothe's Differential Equation in Diffusion Theory.—H. Joos & P. L. Ferreira. (*Ann. Acad. bras. Sci.*, 31st March 1957, Vol. 29, No. 1, pp. 9–22. In English.)

533.6.011 **739**

Equipartition of Energy and Local Isotropy in Turbulent Flows.—M. S. Uberoi. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1165–1170.) Homogeneous turbulence was produced experimentally and experiments showed that the turbulence became isotropic at a faster rate than equipartition of energy occurred. Other experiments where approximately isotropic turbulence was subjected to deformation indicated that even at high Reynolds number the deformation in a shear flow can cause anisotropy. The connection of the investigation with turbulent flows in general is discussed.

535.215 **740**

Barrier-Layer Photo-e.m.f. of Photo-electric Elements with Dyestuffs.—I. A. Karpovich & A. T. Vartanyan. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Nov. 1957, Vol. 117, No. 1, pp. 57–60.) Tests were made on photo elements consisting of dye-stuffs deposited on a quartz plate clamped between semi-transparent electrodes of Pt, Au or Rh. Three different types of assembly were transversely illuminated; in most cases a positive photo-e.m.f. was obtained.

537.122 **741**

On the Nature of the Electron.—J. L. Salpeter. (*Proc. Instn Radio Engrs, Aust.*, June 1957, Vol. 18, No. 6, pp. 183–193; *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1588–1598.) "In this paper the concept of the electron as a fundamental particle of modern physics is discussed in relation to Pauli's exclusion principle, wave mechanics, the uncertainty principle, and relativity."

537.222.6 **742**

The Charge Density near a Sharp Point on a Conductor.—R. Cade. (*Proc. Camb. phil. Soc.*, Oct. 1957, Vol. 53, Part 4, pp. 870–877.) The idea that surface charge tends to infinity at convex sharp points on conductors and to zero at concave points is investigated on electrostatic principles. It is found that the problem has not been solved; a new attempt is made using potential-theory methods from which a fairly general solution is obtained.

537.311.33 : 538.6 **743**

Evaluation of Transport Integrals for Mixed Scattering and Application to Galvanomagnetic Effect.—A. C. Beer, J. A. Armstrong & I. N. Greenberg. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1506–1513.) The Johnson-Whitesall evaluations of the conductivity integrals for mixed scattering have been extended to allow their application to the high-mobility semiconductors. Applications in the analysis of Hall effect, Corbino magnetoresistance and thermomagnetic phenomena as functions of magnetic field are illustrated.

537.312.62 : 530.145.6 **744**

Interaction between Waves and Electrons, with some Remarks on Superconductivity.—L. Brillouin. (*J. Phys. Radium*, May 1957, Vol. 18, No. 5, pp. 331–336.) Discussion of interactions occurring between elastic waves and free electrons in metal, particularly below the Debye temperature.

537.52 **745**

Silent Electric Discharge at Low Frequency in Air, using Insulating Electrodes.—D. P. Jatar & H. D. Sharma. (*C. R. Acad. Sci., Paris*, 22nd July 1957, Vol. 245, No. 4, pp. 414–416.) Note on the Joshi effect and the variation of discharge current as a function of applied voltage, using beeswax electrodes in dry air irradiated by light.

537.525 : 537.222 **746**

Three-Dimensional Potential Well.—H. B. Williams. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1451–1452.) The possibility of generating potential wells having depths of the order of thousands of electron volts is discussed.

537.525 : 621.3.011.3 **747**

Variation of Inductance by Dielectrics, in particular Plasmas.—R. Seitner. (*Z. angew. Phys.*, Feb. 1957, Vol. 9, No. 2, pp. 66–68.) A qualitative interpretation of the frequency changes in oscillatory circuits which are coupled to electrodeless gas discharges. See e.g. 709 above.

537.533 **748**

Analysis of Multivelocity Electron Beams by the Density-Function Method.—A. E. Siegman. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1132–1138.) Mathematical techniques useful in the solution and interpretation of the density-function equations as applied to multivelocity electron-beam problems are presented with some conclusions about the nonconservative nature of signal or noise propagation along a multivelocity beam. The paper serves as a background for a detailed calculation of noise propagation through the gun region of an electron beam (749 below).

537.533 : 621.396.822 **749**

Density-Function Calculations of Noise Propagation on an Accelerated Multivelocity Electron Beam.—A. E. Siegman, D. A. Watkins & Hsung-Cheng Hsieh. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1138–1148.) The propagation of noise fluctuations through the low-voltage multivelocity region immediately in front of the potential minimum of a diode or electron beam has been computed. Despite the absence of any conventional loss elements, two noise parameters or measures of the noise fluctuations when the beam is passed through a multivelocity region are significantly different at the output from those at the input. Changes are such as to lower significantly the minimum noise figure of a microwave valve.

537.533.74 **750**

The Elastic Scattering of Electrons.—R. Zouckermann. (*J. Phys. Radium*, Feb. 1957, Vol. 18, No. 2, pp. 133–137.) An examination of basic theory, in particular the Rutherford formula, with reference to experimental results.

537.533.8 **751**

Theory of Secondary Electron Emission by High-Speed Ions.—E. J. Sternglass. (*Phys. Rev.*, 1st Oct. 1957, Vol. 108, No. 1, pp. 1–12.) A new theoretical treatment shows that the yield of secondaries is proportional to the rate of energy loss of the incident particles and is independent of work function, conductivity and other bulk properties of the metal. The theory explains the experimental observations and its application to general problems of electron escape and capture is discussed.

537.533.8 **752**

Estimate of the Time Constant of Secondary Emission.—A van der Ziel. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1216–1217.) Energy considerations allow a simple estimate to be made of the time constant of secondary emission.

537.56 **753**

Ionization by Positive Ions.—H. B. Gilbody & J. B. Hasted. (*Proc. roy. Soc. A*, 11th June 1957, Vol. 240, No. 1222, pp. 382–395.) A method of measuring the ionization cross-section of atoms is described in which electrons are collected from single collisions of an ion beam passing through a gas at low pressures. Measurements are given for twenty-three cases over energy ranges approximately 5–40 keV and 100 eV–3 keV.

537.56 : 538.69 **754**

Plasma Oscillations in a Steady Magnetic Field: Circularly Polarized Electromagnetic Modes.—T. Pradhan. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1222–1227.) The propagation of e.m. waves in the direction of the field is considered. By taking account of the thermal motion of electrons, results are derived which differ markedly from those of the Appleton-Hartree electromagneto-ionic theory near the cyclotron resonance frequency.

538.11 **755**

Notes on the Ground State of Antiferromagnetism.—O. Nagai. (*J. phys. Soc. Japan*, Aug. 1957, Vol. 12, No. 8, pp. 978.) The conclusions reached by earlier investigators are compared with those which result from the spin-wave theory [see e.g. 726 of 1956 (Marshall)]. The discrepancies are briefly discussed.

538.3 **756**

General Solutions of Equations of a Classical Nonconservative Electromagnetism.—P. Gautier. (*C. R. Acad. Sci., Paris,* 1st July 1957, Vol. 245, No. 1, pp. 45–47.)

538.312 **757**

Electromagnetic Potentials in a Heterogeneous Nonconducting Medium. —A. Nisbet. (*Proc. roy. Soc. A,* 11th June 1957, Vol. 240, No. 1222, pp. 375–381.) For e.m. fields in a stationary nonconducting medium (isotropic or anisotropic), the dielectric constant and permeability of which are given point functions, the general theory of representations in terms of scalar and vector potentials and of Hertzian potentials is developed.

538.566 : 538.221 **758**

Tensor Theory of Gyromagnetic Power.—Y. Le Corre. (*J. Phys. Radium,* May 1957, Vol. 18, No. 5, pp. 312–317.) General linear relations for non-absorbing media are derived. The influence of symmetry upon tensor equations and the effect of magnetic anisotropy on e.m. wave propagation is examined.

538.566 : 621.384.622 **759**

Electromagnetic Waves in Nearly Periodic Structures.—E. Wild. (*Quart. J. Mech. appl. Math.,* Aug. 1957, Vol. 10, Part 3, pp. 322-341.) "The theory of simple harmonic electromagnetic waves in a nearly periodic structure of the type used in linear accelerators is discussed by an expansion in a series of appropriately defined transmission modes of the field in a unit cell of the structure. The possibility of the expansion being assumed, it is shown that the coefficients can be expressed as integrals of the field over the input or output apertures of the unit cell. The modification to the corresponding solution in the strictly periodic structure can be calculated by perturbation theory. In the first approximation the structure can be represented by a series of four-terminal networks, and the voltages and currents in these can be defined in such a way that the relevant parameters vary smoothly in the neighbourhood of resonance."

538.569.4 : 538.2 : 53.082.5 **760**

Optical Detection of Magnetic Resonance in Alkali Metal Vapour.—W. E. Bell & A. L. Bloom. (*Phys. Rev.,* 15th Sept. 1957, Vol. 107, No. 6, pp. 1559–1565.) The apparatus is described together with experimental conditions under which signals have been observed. A possible application of the technique is the measurement of weak magnetic fields.

538.569.4 : 538.22 **761**

Heterodyne Detection of Free Precession in Nuclear Magnetic Resonance. —H. Benoit & R. Klein. (*C. R. Acad. Sci., Paris,* 8th July 1957, Vol. 245, No. 2, pp. 155–157.) Details of a method applied to proton resonance in solutions.

538.569.4 : 538.22 **762**

Narrowing Effect of Dipole Forces on Inhomogeneously Broadened Lines.—S. Geschwind & A. M. Clogston. (*Phys. Rev.,* 1st Oct. 1957, Vol. 108, No. 1, pp. 49–53.) Observations of the effect in Mn ferrite and yttrium iron garnet are described. The theory for cases in which the spatial period of the inhomogeneity is large compared to atomic distances and is either short or comparable to sample size is discussed, and its applications are considered.

538.569.4 : 621.372.826 : 537.226 **763**

Dielectric Rod Waveguide Cells for Microwave Spectroscopy.—E. B. Brackett, P. H. Kasai & R. J. Myers. (*Rev. sci. Instrum.,* Sept. 1957, Vol. 28, No. 9, pp. 699–702.) The calculation of the distribution of microwave power around a quartz rod excited in the dipole mode enabled rod sizes to be selected for cells covering the range 17–60 kMc/s. Teflon windows were found to be satisfactory. See also 3843 of 1957 (Costain).

538.569.4.029.6 : 621.375.9 **764**

Theory of a Three-Level Maser.—A. Javan. (*Phys. Rev.,* 15th Sept. 1957, Vol. 107, No. 6, pp. 1579–1589.) A complete theory is discussed for a gaseous system with an extension to paramagnetic solids.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 **765**

An Attempt to Detect Linearly Polarized Radio Emission from the Galaxy.—J. M. Thomson. (*Nature, Lond.,* 7th Sept. 1957, Vol. 180, No. 4584, pp. 495–496.) A 7·5-m paraboloidal reflector was used at a frequency of 159·5 Mc/s with receiver bandwidth 4 Mc/s. Plots of position angle of the plane of polarization against time are given for regions near 05 h and 18 h and show that its variation is in the direction to be expected from the ionospheric data.

523.16 **766**

Relative Intensities of the Four Principal R.F. Sources Observed at a Wavelength of 22 cm; Note on R.F. Source *Sagittarius A.*—G. Westerhout. (*C. R. Acad. Sci., Paris,* 1st July 1957, Vol. 245, No. 1, pp. 35–38.) Some 55 r.f. sources have been located during observations of the Milky Way region made by means of the new 25-m radio telescope at Dwingeloo, Netherlands. Special studies have been made of five of these in Cass A, Cygn A, Taur A, Virg A, and Sgtr A, and the equivalent temperatures have been determined.

523.16 : 523.72 **767**

The Effects of Incomplete Resolution on Surface Distributions Derived from Strip-Scanning Observations, with Particular Reference to an Application in Radio Astronomy.—S. F. Smerd & J. P. Wild. (*Phil. Mag.,* Jan. 1957, Vol. 2, No. 13, pp. 119–130.) The limitations of one-dimensional scanning methods for measuring surface distribution, particularly that of radio brightness over the sun's disk, are discussed. See e.g. 2607of 1955 (O'Brien & Tandberg-Hanssen).

523.16 : 523.72 : 621.396.677 **768**

Tests on a Model of an Aerial for Use in Radio Astronomy.—Corazza & Francini. (See 673.)

523.16 : 523.841.11 **769**

Radio Emission from the Remnants of the Supernovae of 1572 and 1604.—J. E. Baldwin & D. O. Edge. (*Observatory,* Aug. 1957, Vol. 77, No. 899, pp. 139–143.) Measurements at 1·9 m λ are compared with previous radio and optical observations [see e.g. 2708 of 1956 (Mills et al.)].

523.165 **770**

The Cosmic Radiation and Solar Terrestrial Relationships.—J. A. Simpson. (*Ann. Géophys.,* July-Sept. 1955, Vol. 11, No. 3, pp. 305–329. In English.) Paper presented at the Assembly of the Union Géodésique et Géophysique Internationale, Rome, September 1954, reviewing the properties of cosmic radiation and its association with solar and geophysical processes, and including a description of experimental methods developed to study low-energy cosmic-ray particles.

523.5 **771**

The Incidence of Meteor Particles upon the Earth.—A. A. Weiss. (*Aust. J. Phys.,* Sept. 1957, Vol. 10, No. 3, pp. 397–411.) Radio echo rates for both shower and sporadic meteors, measured at Adelaide with 27-Mc/s c.w. equipment, are applied to the calculation of the incident flux above limiting brightnesses in the region $M_R < +7{\cdot}5$. Dependence of ionizing probability on velocity is discussed, and the fluxes and densities agree reasonably well with independent evaluations from visual meteor rates.

523.53 **772**

A Prediction of a Meteor Orbital Period.—E. G. Bowen. (*Observatory,* June 1957, Vol. 77, No. 898, pp. 99–102.) The probable orbital period of a meteor shower, observed on 5th–6th December 1956, is predicted from examination of snowfall records in Japan, assuming the suggested correlation between precipitation and occurrence of meteor showers to be real.

550.385 **773**

Analysis of the Sinusoidal Variation of Magnetic Declination of 11th April 1954.—K. Burkhart & E. Selzer. (*Ann. Géophys.,* July–Sept. 1955, Vol. 11, No. 3, pp. 353–368.) Analysis of magnetograms of 14 European stations suggests the development of a counter-clockwise vortex-like perturbation in the ionosphere bordering on the southern limits of the auroral zone and lasting less than 2 h.

551.5 : 621.396.96 **774**

Radar in the Rain.—(*Electronic Radio Engr,* Jan. 1958, Vol. 35, No. 1, pp. 13–15.) An outline of radar investigations into cloud structure.

551.508.822 **775**

Radiosonde Trials at Payerne, 1956. —A. H. Hooper. (*Met. Mag., Lond.,* Feb. 1957, Vol. 86, No. 1016, pp. 33–36, plates.)

Report of trials in Switzerland in May and June 1956 in which 14 types of radiosonde from various countries were compared.

551.510 776

Investigation of the Upper Layers of the Atmosphere.—V. V. Mikhnevich & I. A. Khvostikov. (*Izv. Ak. Nauk S.S.S.R., Ser. geofiz.*, Nov. 1957, No. 11, pp. 1393–1409.) Measurements of composition, pressure, density and temperature, up to a height of 90 miles, were made by means of containers which collected samples of air after being automatically released from rockets. Experimental data are tabulated and some results obtained in the U.S.A. and U.S.S.R. compared. 78 references.

551.510.535 777

Ionospheric Drifts at Brisbane.—M. J. Burke & I. S. Jenkinson. (*Aust. J. Phys.*, Sept. 1957, Vol. 10, No. 3, pp. 378–386.) Measurements over a 2-year period show that speeds are less than those observed at higher altitudes. The E-region 12-h and 8-h solar harmonics show large seasonal phase changes in the northward component; the phase for the 12-h northward component in summer is in fair agreement with higher latitudes. The F-region 12-h harmonic seasonal change is greater in the eastward than in the northward component.

550.510.535 778

Measurement of Ionospheric Winds.—P. Lejay. (*C. R. Acad. Sci., Paris*, 17th July 1957, Vol. 245, No. 3, pp. 253–257.) Experiments made at Domont, France, by the three-receiver method are described. The electron displacements deduced from observations correspond to some extent with those necessary to account for the daily variations in the earth's magnetic field.

551.510.535 : 551.593.9 779

Some Possible Relations between the Nightglow and the Ionosphere.—P. St. Amand. (*Ann. Géophys.*, Oct.-Dec. 1955, Vol. 11, No. 4, pp. 450–460. In English.) "Ionospheric data from Stanford, California are compared with nightglow data from Cactus Peak, California. The nocturnal variation of the brightness of the green line, 5 577, of [OI] is found to be similar to the variation of height of the F region. The nocturnal variation of the brightness of the red lines 6 300–6 364 of [OI] is found to be similar to that of the electron density of the F region. The possibility that the red lines are produced by radiation following dissociative recombination of electrons with O_2^+ ions is discussed." Paper presented at the Assembly of the Union Géodésique et Géophysique Internationale, Rome, September 1954.

551.510.535 : 621.396.11 780

The Reflection Coefficient of Ionospheric Layers.—G. Pillet. (*C. R. Acad. Sci., Paris*, 17th July, 1957, Vol. 245, No. 3, pp. 335–338.) Experiments made at 2·1 and 3·4 Mc/s show that the reflection coefficient which is nearly unity for the F layer falls to a very low value for the E layer and that loss of energy due to reflection from the latter is of the same order as the nondeviative absorption.

551.510.535 : 621.396.11.029.62 781

The Scattering of Radio Waves of Very High Frequency in the Ionosphere.—Ionescu. (See 890.)

551.594.22 : 621.396.969 782

Radar Echoes from Inter-stroke Processes in Lightning.—F. J. Hewitt. (*Proc. phys. Soc.*, 1st Oct. 1957, Vol. 70, No. 454B, pp. 961–979.) Radar echoes at 50 cm λ were used to investigate inter-stroke lightning processes. The most intense activity of junction streamers is between heights of 4–7 km; these junction streamers may sometimes grow in height up to 10 km although their activity decreases with height; the lowest level at which they occur remains constant throughout flashes and their horizontal extent is about 1½ to 2 km. Echoes from these junctions always decrease before the onset of a flash.

551.594.5 783

Theory of the Auroral Spectrum.—D. R. Bates. (*Ann. Géophys.*, July-Sept. 1955, Vol. 11, No. 3, pp. 253–278. In English.) Paper presented at the Assembly of the Union Géodésique et Géophysique Internationale, Rome, September 1954. The main luminescence observed is due to (*a*) inelastic collisions made by the incident particles and by the ejected electrons, (*b*) degradation of the ultraviolet radiation emitted and scattering of solar radiation, (*c*) thermal processes such as dissociative recombination. 81 references.

551.594.5 : 551.593.9 784

The Work of Soviet Scientists on the Luminescence of the Night Sky and Aurorae.—B. A. Bagaryatskiĭ. (*Izv. Ak. Nauk S.S.S.R., Ser. geofiz.*, Nov. 1957, No. 11, pp. 1410–1417.) A review of the research carried out in the last 30 years with particular reference to the I.G.Y., in a network of stations all over the U.S.S.R. 118 references.

551.594.6 785

A Method for Interpreting the Dispersion Curves of Whistlers.—L. R. O. Storey. (*Canad. J. Phys.*, Sept. 1957, Vol. 35, No. 9, pp. 1107–1122.) An analytical solution to the problem of determining the dispersion of a whistler from a known electron density in the outer atmosphere is formulated. The results are then applied to the inverse problem of deducing the variation in electron density from the dispersion curve. A numerical example is given.

551.594.6 : 621.396.663 786

Calibration of Narrow-Sector Radiogoniometers for Atmospherics.—Carbenay. (See 788.)

LOCATION AND AIDS TO NAVIGATION

534.88-14 787

Underwater Acoustic Echo-Ranging.—J. W. R. Griffiths & A. W. Pryor. (*Electronic Radio Engr*, Jan. 1958, Vol. 35, No. 1, pp. 29–32.) Measurements having shown the reverberation power spectrum to be similar to that of the emitted pulse, the effects of pulse duration and receiver bandwidth on reverberation level and peak-signal reverberation ratio are examined theoretically. The pulse should be as short as possible and the product bandwidth × pulse duration should be approximately unity.

621.396.663 : 551.594.6 788

Calibration of Narrow-Sector Radiogoniometers for Atmospherics.—F. Carbenay. (*C. R. Acad. Sci., Paris*, 17th July 1957, Vol. 245, No. 3, pp. 298–300.) An extension of the method of calibrating omnidirectional recording receivers with reference to signal flux and effective aerial height. See 1451 of 1957, 1651 of 1951 and back references.

621.396.933 : 621.395.625.3 789

Multiple-Track Tape Recorders in Air Traffic Control.—K. Heidelauf. (*Nachrichtentech, Z.*, July 1957, Vol. 10, No. 7, pp. 344–348.) Description of German 14-track equipment.

621.396.96 : 551.5 790

Radar in the Rain.—(See 774.)

621.396.96 : 621.316.726.078 791

Transistorized A.F.C. uses Triangular Search Sweep.—H. H. Hoge & D. L. Spotten. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 178–179.) Circuit details of a radar a.f.c. system.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 792

Generalities about Outgassing at Room Temperature.—R. Geller. (*Le Vide*, May/June 1957, Vol. 12, No. 69, p. 194. In French and English.)

533.5 793

Outgassing at Room Temperature of Materials under Vacuum.—R. Barré, R. Geller & G. Mongodin. (*Le Vide*, May/June 1957, Vol. 12, No. 69, pp. 195–201. In French and English.) Work done in connection with the construction of the synchrotron at Saclay is described. The overall outgassing of a standard vacuum installation may be calculated on the hypothesis that most metals used are comparable to steel. Characteristics obtained for rubber are shown.

533.5 : 621.385.032.213.13 794

Practical Experiences with a Vacuum-Tight Sealing Mechanism for the Repeated Use of Activated Oxide Cathodes.—H. Fetz & K. Schiefer. (*Z. angew. Phys.*, Jan. 1957, Vol. 9, No. 1, pp. 13–14.) The equipment described is used in investigations of cathode sputtering processes. The life of the cathode is prolonged by protecting the cathode from contact with the oxygen of the atmosphere every time the sputtering vessel is opened. An interpretation is given of experimental results observed in tests of a BaO cathode.

533.5 : 621.389 795

Mass Spectrometer for Leak Detection operating with a Mixture containing a Small Percentage of Helium.—H. Warmoltz & H. A. M. de Grefte. (*Le Vide*, May/June 1957, Vol. 12, No. 69, pp. 202–207.)

535.215 : 546.32-1 796

The External Photoelectric Effect of the Alkali Metals: Part 1.—H. Thomas. (*Z. Phys.*, 12th Jan. 1957, Vol. 147, No. 4, pp. 395–418.) The dependence of the photoelectric effect on the thickness of the metal film, and the range of photoelectrons in K are investigated. Measurements indicate a bulk effect in the wavelength range 289–578 mμ. A satisfactory interpretation of results is obtained on the basis of plasma theory. 47 references.

535.215 : 546.32-1 797

The External Photoelectric Effect of the Alkali Metals: Part 2.—H. Mayer & H. Thomas. (*Z. Phys.*, 12th Jan. 1957, Vol. 147, No. 4, pp. 419–441.) The spectral distribution of the quantum yield was investigated for thin films and heavier deposits of K. Experimental results can be satisfactorily interpreted by theory based on a photoelectric bulk effect which is also applicable to the spectral distribution obtained with solid K. 47 references. Part 1: 796 above.

535.215 : 546.32-1 798

The External Photoelectric Effect of Alkali Metals: Part 3.—S. Methfessel. (*Z. Phys.*, 12th Jan. 1957, Vol. 147, No. 4, pp. 442–464.) The energy distribution of photoelectrons emitted from K and Cs films of various thicknesses was measured. The photoeffect is found to be mainly a bulk effect. 38 references. Part 2: 797 above.

535.215+535.376] : 546.472.21 799

Anisotropy in Electroluminescence and Conductivity of Single Crystals of ZnS.—A. Lempicki, D. R. Frankl & V. A. Brophy. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1238–1239.) The marked anisotropy observed is considered to arise from stacking faults in the direction of the *c*-axis.

535.215 : 546.59 800

Images of Strained Metal Samples, obtained with a Photoemission Microscope.—R. Bernard, C. Guillaud & R. Goutte. (*J. Phys. Radium*, May 1957, Vol. 18, No. 5, pp. 327–330.) The crystalline structure of pulled gold samples after etching by ionic bombardment were examined. Areas of maximum strain showed an increased brightness which could be due to a local increase of photoelectric emission.

535.215 : 546.682.19 : 621.383.4 801

Photoelectric Effects in InAs at Room Temperature.—C. Hilsum. (*Proc. phys. Soc.*, 1st Oct. 1957, Vol. 70, No. 454B, pp. 1011–1012.) A note of the photoconductive and photoelectromagnetic effects in InAs, showing the spectral sensitivity of a photocell constructed by etching.

535.215 : 546.816.231 802

Optical Sensitization of Photoconductors of the Lead Salt Group.—V. Schwetzoff. (*C. R. Acad. Sci., Paris*, 8th July 1957, Vol. 245, No. 2, pp. 149–152.) A residual increase of conductivity σ and variation of relative sensitivity $\Delta\sigma/\sigma$, are observed after subjecting photoconductors at low temperature (77°K) to visible light. The effect is confirmed for PbS, PbSe and PbTe.

535.37 803

Sulphur Vacancy Emission in ZnS Phosphors.—N. T. Melamed. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, p. 1727.) An emission band centred on 3 950 Å is attributed to sulphur vacancies rather than interstitial silver activator.

535.37 804

The Luminescence of Alkali Vanadates.—H. Gobrecht & G. Heinsohn. (*Z. Phys.*, 10th Jan. 1957, Vol. 147, No. 3, pp. 350–360.) Results of an experimental investigation of four different vanadates are compared and an interpretation of the luminescence mechanism is given.

537.226/.227 805

Dielectric and Thermal Study of Tri-glycine Sulphate and Tri-glycine Fluoberyllate.—S. Hoshino, T. Mitsui, F. Jona & R. Pepinsky. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1255–1258.) The dielectric constants ϵ_b show a pronounced anomaly at the Curie temperatures of 48°C and 70°C for the sulphate and the fluoberyllate respectively. The dielectric constants ϵ_a and ϵ_c are practically temperature-independent. Measurements of the specific heat as a function of temperature yield values of entropy change $\Delta S = 0 \cdot 48$ and $1 \cdot 17$ cal/mole degree respectively.

537.226/.227 : 546.431.824-31. 806

Preparation of Thin Single Crystals of Barium Titanate.—J. T. Last. (*Rev. sci. Instrum.*, Sept. 1957, Vol. 28, No. 9, pp. 720–721.) Single crystal samples as thin as $1 \cdot 5\,\mu$ were prepared using phosphoric acid as an etchant and Teflon-silicone tape as a support.

537.226/.227 : 546.431.824-31 807

Domain Effects in Polycrystalline Barium Titanate.—E. C. Subbarao, M. C. McQuarrie & W. R. Buessem. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1194–1200.) The indirect observation of domain changes under electrical and mechanical stresses was undertaken principally by means of X-ray reflections and dimensional changes. Observations were made of the changes as a function of time.

537.226/.227 : 546.431.824-31 808

Effect of Iron-Group Ions on the Dielectric Properties of $BaTiO_3$ Ceramics.—T. Sakudo. (*J. phys. Soc. Japan*, Sept. 1957, Vo. 12, No. 9, p. 1050.) The Curie point of these ceramics falls with the addition of Fe or Ni ions but is hardly affected by Mn or Co. For Cu, the Curie point rises; this is briefly discussed.

537.226.2 809

The Existence Domain of Complex Dielectric Constant of Binary Mixture.—F. Irie. (*Ann. Phys., Lpz*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 31–40. In English.) Extension of existing theory and notes on its applications.

537.226.2/.3 : [547.261+547.262 810

The Dielectric Properties of Methyl and Ethyl Alcohols in the Wavelength Range 3 cm–52 cm.—E. H. Grant. (*Proc. phys. Soc.*, 1st Oct. 1957, Vol. 70, No. 454B, pp. 937–944.) The dielectric behaviour of both alcohols can be described in terms of a principal dispersion region at cm λ together with a subsidiary dispersion region occurring at mm λ.

537.227 811

Ferroelectricity in Glycine Silver Nitrate.—R. Pepinsky, Y. Okaya, D. F. Eastman & T. Mitsui. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1538–1539.) Ferroelectricity is observed below −55°C. At −195°C the spontaneous polarization is $0 \cdot 55 \times 10^{-6}$ C/cm² and the coercive field is 740 V/cm. There are no dielectric anomalies between 4°K and −55°C.

537.227 812

Ferroelectricity of Dicalcium Strontium Propionate.—B. T. Matthias & J. P. Remeika. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, p. 1727.) Preliminary results are reported.

537.311.31 : 538.63 813

Magnetoresistance in Metals.—D. K. C. Macdonald. (*Phil. Mag.*, Jan. 1957, Vol. 2, No. 13, pp. 97–104.) Report of experimental measurements on plate-shaped specimens of Na and Rb, and on a cylindrical specimen of Na.

537.311.31 : 538.63 814

New Type of Oscillatory Magnetoresistance in Metals.—J. Babiskin & P. G. Siebenmann. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1249–1254.) The magnetoresistive properties of a thin sodium wire have been studied at 1°K in transverse magnetic fields up to 60 000 G. The oscillations observed were not of the de Haas-van Alphen type: they are interpreted as being caused by surface scattering.

537.311.33 815

Theory of Impurity-Centre Electrons: Part 2—Nonradiative Transitions.—G. Helmis. (*Ann. Phys., Lpz.*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 41–54.) Further application of the method given in Part 1 (3770 of 1956).

537.311.33 816

The Effect of Free Electrons on Lattice Conduction at High Temperatures.—R. Stratton. (*Phil. Mag.*, March 1957, Vol. 2, No. 15, pp. 422–424.) Phonon-electron scattering is considered for the range above the Debye characteristic temperature θ. See 2017 of 1956 (Ziman) for lattice temperatures below θ.

537.311.33 817

Variation of Mobility with Electric Field in Nondegenerate Semiconductors.—M. S. Sodha. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1266–1271.) By assuming a Maxwellian distribution of electron velocities at normal temperatures, in a semiconductor having low impurity concentration, it is shown that the net zero-field mobility decreases monotonically with decreasing 'impurity' mobility.

537.311.33 **818**

Semiconductor Compounds Open New Horizons.—A. Coblenz. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 144–149.) In this survey inorganic and organic semiconducting compounds are distinguished from conventional semiconductors, and their characteristics, unusual properties and applications are discussed. Tabulated data on a large number of compounds are given.

537.311.33 **819**

Thermo-compression Bonding.—(*Bell Lab. Rec.*, Sept. 1957, Vol. 35, No. 9, p. 336.) A combination of heat and pressure is used to provide a firm bond between various soft metals and clean single-crystal semiconductor surfaces.

537.311.33 **820**

Cleaning Semiconductor Components.—(*Bell Lab. Rec.*, Sept. 1957, Vol. 35, No. 9, p. 337.) A continuous water washing system is described for removing all water-soluble materials that remain after etching. The effectiveness of the washing process is monitered automatically.

537.311.33 : 538.63 **821**

Magnetoresistance of Holes in Germanium and Silicon with Warped Energy Surfaces.—J. G. Mavroides & B. Lax. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1530–1534.) Experimental directional magnetoresistance effects and the variation of magnetoresistance with magnetic field are interpreted in terms of calculated values of the magnetoresistance coefficients.

537.311.33 : 546.23 : 535.3 **822**

Absorption of Light in Se Near the Band Edge.—W. J. Choyke & L. Patrick. (*Phys. Rev.*, 1st Oct. 1957, Vol. 108, No. 1, pp. 25–28.) Photovoltaic measurements, at temperatures 80°K–440°K with photons of energy 1·6 eV–2·0 eV, show that the transitions are indirect at the band edge, requiring the absorption or emission of a phonon. The energy gap for hexagonal Se is deduced and the band edges for hexagonal and amorphous Se are compared.

537.311.33 : 546.24 **823**

Growth of Tellurium Single Crystals by the Czochralski Method.—T. J. Davies. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1217–1218.)

537.311.33 : 546.28 **824**

Threshold Energy for Electron-Hole Pair Production by Electrons in Silicon.—A. G. Chynoweth & K. G. McKay. (*Phys. Rev.*, 1st Oct. 1957, Vol. 108, No. 1, pp. 29–34.) Measurements of the reverse bias necessary for the onset of multiplication in silicon *p-n* junctions of various widths lead to a value of $2 \cdot 25 \pm 0 \cdot 10$ eV for the threshold energy for electron-hole pair production by energetic electrons. An apparent slight variation of threshold energy with crystallographic direction is noted and the ionization rate is found to be greater for electrons than for holes. The maximum phonon drag opposing the acceleration of an electron up to the threshold energy by a parabolic field distribution is equivalent to a field of $5 \cdot 2 \times 10^4$ V/cm.

537.311.33 : 546.281.26 **825**

Intrinsic Electrical Conductivity in Silicon Carbide.—J. H. Racette. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1542–1544.) Measurements on *n*-type hexagonal single crystals are described. The band gap at absolute zero is $3 \cdot 1 \pm 0 \cdot 2$ eV on the assumption of an intrinsic conductivity variation of the form $\sigma =$ constant $\times \exp(-\Delta E/2kT)$.

537.311.33 : 546.289 **826**

Processes of Preparation of Germanium Single Crystals.—T. Niimi, H. Baba, N. Ogawa, K. Furusho & C. Tadachi. (*Rep. elect. Commun. Lab., Japan*, May 1957, Vol. 5, No. 5, pp. 5–9.) Accepted production methods are outlined and some special techniques are described which have have been developed to solve problems in the processes of reduction, zone melting and preparation of single crystals.

537.311.33 : 546.289 **827**

Free Bonds ou the Clean Surfaces of Germanium Single Crystals.—A. Kobayashi & S. Kawaji. (*J. phys. Soc. Japan*, Sept. 1957, Vol. 12, No. 9, p. 1054.) Field effect patterns have been observed on clean and oxidized surfaces. The density of the fast states is estimated at $1 \cdot 7 \times 10^{13}$ cm^{-2} eV for a clean surface and $2 \cdot 4 \times 10^{12}$ cm^{-2} eV for an oxidized surface.

537.311.33 : 546.289 **828**

Detection of Both Vacancies and Interstitials in Deformed Germanium.—J. N. Hobstetter & P. Breidt, Jr. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1214–1215.) Experimental evidence supporting the view that interstitials as well as vacancies form during deformation of Ge, was developed from studies of the effect of small compressions on the electrical conductivity of specimens of various initial conductivities.

537.311.33 : 546.289 : 534.2-8 **829**

Acousto-electric Effect in *n*-Type Germanium.—W. Sasaki & E. Yoshida. (*J. phys. Soc. Japan*, Aug. 1957, Vol. 12, No. 8, p. 979.) The flow of acoustic energy through a Ge crystal has been measured over a range of ultrasonic frequencies. The results confirm the prediction made by Parmenter (2281 of 1953) concerning the interaction of electrons and acoustic waves. See also 1039 of 1957 (Weinreich).

537.311.33 : 546.289 : 539.16 **830**

Effect of Irradiation on the Hole Lifetime of *N*-Type Germanium.—O. L. Curtis, Jr, J. W. Cleland, J. H. Crawford, Jr, & J. C. Pigg. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1161–1165.) The minority-carrier lifetime in *n*-type Ge is extremely sensitive to irradiation by fast neutrons and γ rays. The simple dependence of recombination rate on irradiation received, permits prediction of the expected decrease in lifetime in a known radiation field. For the same change in carrier concentration, change in lifetime is much smaller when produced by γ radiation than by fast neutrons.

537.311.33 : 546.431-31 **831**

Fundamental Absorption of Barium Oxide from its Reflectivity Spectrum.—F. C. Jahoda. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1261–1265.) Results compare favourably with those obtained previously by Zollweg (2024 of 1955) from transmission measurements with thin films.

537.311.33 : 546.682.19 : 535.3 **832**

Optical Absorption in *p*-Type Indium Arsenide.—F. Stern & R. M. Talley. (*Phys. Rev.*, 1st Oct. 1957, Vol. 108, No. 1, pp. 158–159.) An absorption peak, observed on the long-wavelength side of the intrinsic absorption edge, is attributed to transitions between the light- and heavy-hole bands and a smaller peak at 0·055 eV to lattice absorption.

537.311.33 : 546.682.86 : 537.312.9 **833**

Piezoresistive Effect in Indium Antimonide.—F. P. Burns & A. A. Fleischer. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1281–1282.) " Room-temperature measurements on the variation of resistivity of pure InSb with hydrostatic and uniaxial stress were made to determine the piezoresistive and elastoresistive coefficients of pure InSb. The results are consistent with a spherical conduction-band model."

537.311.33 : 546.817.221 : 621.314.7 **834**

Transistor Action on Natural Galena Surface after Heat Treatment with H_2S.—J. N. Das & P. V. Khandekar. (*Z. Phys.*, 10th Jan. 1957, Vol. 147, No. 3, pp. 271–276. In English.) Report of experiments on *n*-type galena crystals. Transistor action and photovoltaic effects were observed. See also 2560 of 1956 (Bhide et al.).

537.312.62 **835**

Superconducting Alkaline Earth Compounds.—B. T. Matthias & E. Corenzwit. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, p. 1558.)

537.323 **836**

Friedel Theory of Thermoelectric Power Applied to Dilute Magnesium Alloys.—E. I. Salkovitz, A. I. Schindler & E. W. Kammer. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1549–1552.)

537.324 : 539.23 **837**

Juxtaposition Thermocouples in the form of Thin Films.—A. Aron. (*C.R. Acad. Sci., Paris*, 1st July 1957, Vol. 245, No. 1, pp. 48–50.) Report of measurements on thermocouple systems, one element of which is a film of Ag or Al on pyrex glass, and the other a film of Cu_2O, Te or antimony oxide deposited on an Al or Ag film.

537.582 : 546.883 **838**

Thermionic Emission from a Planar Tantalum Crystal.—H. Shelton. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1553–1557.) For a clean [211] surface the work function is $4 \cdot 352 \pm 0 \cdot 01$ V and the emission constant A is 120 A/cm^2 °K^2. Results for contaminated surfaces are also given.

538.22 **839**

The Magnetic Properties of Iron Selenide Single Crystals.—K. Hirakawa. (*J. phys. Soc. Japan*, Aug. 1957, Vol. 12, No. 8, pp. 929–938.) Large single crystals of Fe_7Se_8 and Fe_3Se_4 were prepared and their magnetic properties investigated below the Curie point. A change in the direction of easy magnetization occurs for the former but not for the latter crystal.

538.221 **840**

Magnetic Domain Patterns on Single-Crystal Iron Whiskers.—R. V. Coleman & G. G. Scott. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1276–1280.) These have been investigated using the Bitter powder technique. Whiskers with axes along the [111] and [100] directions have a very simple domain structure in the unmagnetized state.

538.221 **841**

The Investigation of the Magnetic Reversal Process in High-Coercivity Alnico by means of the Powder Pattern Technique.—W. Andrä. (*Ann. Phys., Lpz.*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 10–18.) Continuation of the work of Kussmann & Wollenberger (3805 of 1956).

538.221 **842**

Remarks on Zener's Classical Theory of the Temperature Dependence of Magnetic Anisotropy Energy.—R. Brenner. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1539–1541.) The temperature dependence of the first anisotropy constant for Ni is calculated by averaging the local anisotropy over a Langevin function. Reasonably good agreement with the experimental curve is obtained for $T/\theta_c > 0{\cdot}3$.

538.221 **843**

Anhysteretic Remanent Magnetization of Ferrimagnetics.—F. Rimbert. (*C. R. Acad. Sci., Paris*, 22nd July 1957, Vol. 245, No. 4, pp. 406–408.) Remanence characteristics of Fe_3O_4 and α-Fe_2O_3 samples subjected to a continuous field superimposed on a decreasing alternating field are shown.

538.221 **844**

Evidence for Subgrains in MnBi Crystals from Bitter Patterns.—W. C. Ellis, H. J. Williams & R. C. Sherwood. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1215–1216.)

538.221 : 539.23 : 53.087.63 **845**

Magnetic Writing on Thin Films of MnBi.—H. J. Williams, R. C. Sherwood, F. G. Foster & E. M. Kelley. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1181–1184.) The domain structure of thin films ($\approx$1 000Å) of MnBi and their capability of storing information magnetically has been studied. Information can be read optically using the Faraday effect. The films have a uniaxial direction of easy magnetization normal to their surfaces and retain magnetization after saturation along the direction in spite of the large demagnetizing factor. High optical contrast between writing and background can be obtained.

538.221 : [621.318.124 + 621.318.134 **846**

Magnetic Anisotropy of Cobalt Ferrite ($Co_{1.01}Fe_{2.00}O_{3.62}$) and Nickel Cobalt Ferrite ($Ni_{0.72}Fe_{0.20}Co_{0.08}Fe_2O_4$).—H. Shenker. (*Phys. Rev.*, 1st Sept. 1957, Vol. 107, No. 5, pp. 1246–1249.) Measurements of the first magnetic anisotropy constant K_1 are described. The results are similar to those for metallic ferromagnetic materials for which the temperature variation of K_1 is expressed in the form $K_1(T)/K_1(0) = \exp(-\alpha T^2)$.

538.221 : 621.318.134 **847**

Ferrimagnetic Resonance of Erbium Garnet at 9 400 Mc/s.—J. Paulevé. (*C. R. Acad. Sci., Paris*, 22nd July 1957, Vol. 245, No. 4, pp. 408–411.) Experimental results were obtained for temperatures from 4°K to 530°K. The compensation temperature is 84°K. See also 3214 of 1957.

538.224 : 546.3-1'56'47 **848**

The Magnetic Susceptibility of α and β Brass.—B. G. Childs & J. Penfold. (*Phil. Mag.*, March 1957, Vol. 2, No. 15, pp. 389–403.) Measurements of magnetic susceptibility were made at 77° and 300° K on a series of Cu-Zn alloys. The results are discussed with reference to the theory proposed by Henry & Rogers (2824 of 1956).

538.632 **849**

Hall Effect in Titanium, Vanadium, Chromium and Manganese.—S. Foner. (*Phys. Rev.*, 15th Sept. 1957, Vol. 107, No. 6, pp. 1513–1516.) Measurements have been made at room temperature with fields up to 30 000 oersteds. The effect is linear with magnetic field and positive.

538.632 : 546.32-1 : 539.23 **850**

Hall Effect and Conductivity Measurements on Thin Films of Potassium.—W. Cirkler. (*Z. Phys.*, 12th Jan. 1957, Vol. 147, No. 4, pp. 481–498.) Measurements were made for a thickness range of 30–2 000 Å. Experimental results are compared with those obtained theoretically.

549.514.5 : 539.185.9 **851**

Effect of Rapid Neutrons on some Physical Constants of Crystalline Quartz and Vitreous Silica.—G. Mayer & J. Gigon. (*J. Phys. Radium*, Feb. 1957, Vol. 18, No. 2, pp. 109–114.)

621.315.612 + 621.318.124 + 621.318.134 **852**

Some New Electrical and Magnetic Ceramics.—G. Campbell. (*J. sci. Instrum.*, Sept. 1957, Vol. 34, No. 9, pp. 337–348.) Well known insulating and conducting types of ceramics are briefly discussed and the ferroelectric and ferromagnetic types are examined in some detail. Their applications are described in transducers, dielectric amplifiers, storage devices and microwave components, and including future developments in 'solid' circuits.

621.315.614.6 : 621.317.335.3.029.64 **853**

Dielectric Anisotropy of Paper at 3 400 Mc/s. Influence of Humidity.—R. Servant & J. Cazayus-Claverie. (*C. R. Acad. Sci., Paris*, 29th July 1957, Vol. 245, No. 5, pp. 509–511.) Extension of earlier measurements at 9 350 Mc/s [3140 of 1956 (Servant & Gougeon)] gives similar results; strong birefringence accompanied by rectilinear dichroism.

621.357.53 **854**

Conductive and Resistive Coatings.—R. J. Phair. (*Bell Lab. Rec.*, Sept. 1957, Vol. 35, No. 9, pp. 331–335.) Details are given of the preparation of conductive layers by the use of resins and lacquers pigmented with metals or carbon. These are applied like paint to give thin resistive films in complex patterns.

MATHEMATICS

518.2 **855**

Mathematical Tables—a Bibliography.—C. R. Sexton. (*Product Engng*, July 1957, Vol. 28, No. 7, pp. 183–194.) An annotated list of mainly American publications.

MEASUREMENTS AND TEST GEAR

529.786 + 621.3.018.41(083.74) **856**

Standards of Time and Frequency.—L. Essen. (*Research, Lond.*, June 1957, Vol. 10, No. 6, pp. 217–224.) The mean solar second is used for current purposes and is uniform to approximately ± 5 parts in 10^9. A quartz clock used at the National Physical Laboratory since June 1955 and calibrated by a caesium atomic standard to an accuracy within ± 1 part in 10^{10} provides a basis for monitoring MSF standard-frequency transmissions. Astronomical standards will still be required for preserving the continuity of time measurements over long periods.

621.3.018.41(083.74) : 621.396.666 **857**

Fade-Cancelling Zero-Beat Indicator for Reception of Standard Radio Frequencies.—R. J. Blume. (*Rev. sci. Instrum.*, Sept. 1957, Vol. 28, No. 9, pp. 703–708.) A simple circuit is described which enables a local secondary frequency standard to be set quickly to a definite zero beat with a received s.w. standard-frequency signal, without the usual uncertainty due to amplitude fading of the signal. The bandwidth of the device may be made much less than 1 c/s.

621.3.089.6 : 061.6 **858**

Electronic Calibration Centre of the National Bureau of Standards.—(*Engineer, Lond.*, 31st May 1957, Vol. 203, No. 5288, pp. 852–853.) Description of the centre under construction at Boulder, Colorado, which will provide greatly expanded facilities for calibration services at all frequencies in general use.

621.317.2 : 621.373.42 **859**

Simplified General-Purpose Signal Generator.—M. W. Kirby. (*Short Wave Mag.*, July 1957, Vol. 15, No. 5, pp. 243–244.) A low-cost circuit using two valves: one as a cathode-coupled oscillator, allowing wide frequency coverage by means of coil switching, the other as modulator.

621.317.2 : 621.373.52 : 621.397.62 **860**

A Transistorized TV Bar Generator.—T. G. Knight. (*Radio TV News*, Sept. 1957, Vol. 58, No. 3, pp. 48–49.) A self-contained television pattern generator for checking the linearity of deflection circuits. It operates over a range 48–64 Mc/s.

621.317.3 : 621.314.7 **861**

Accurate Measurement of r_c and x_0 for Transistors.—M. A. Melehy. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1739–1740.)

621.317.3 : 621.372.412.088.33 **862**

Tolerances of Quartz Crystals for Filters and their Measurement.—Kurth & Miczynski. (See 697.)

621.317.335.3.029.63/.64 **863**

Method of Measurement to Determine the Complex Dielectric Constant at Wavelengths from 8 to 80 cm and Temperatures to −150°C.—H. K. Ruppersberg. (*Z. angew. Phys.*, Jan. 1957, Vol. 9, No. 1, pp. 9–13.) The method described is one of input impedance measurement on a coaxial line filled with the liquid under test. Special arrangements are made to counter the effects of low temperature, including the use of a short-circuit plunger (1198 of 1957). The method is particularly suitable for lossy materials; errors range from 1 to 5%. A modification of another method [2160 of 1956 (Lueg & Ruppersberg)] is outlined which is suitable for solid disk-shaped specimens.

621.317.335.3.029.64 **864**

Measurement of the Complex Dielectric Constant of Very-High-Dielectric-Constant Materials at Microwave Frequencies.—I. Bady. (*Commun. & Electronics*, May 1957, No. 30, pp. 225–228.) Two methods of measurement are described, one for low-loss and the other for high-loss specimens. A short-circuited waveguide is used in both cases, but greater sensitivity is obtained than with conventional methods.

621.317.335.3.029.64 : 546.217 **865**

Microwave Refractometer Cavity Design.—A. W. Adey. (*Canad. J. Technol.*, March 1957, Vol. 34, No. 8, pp. 519–521.) The operation of the instrument involves the comparison of the resonance frequency of a sealed reference cavity with that of a cavity exposed to the atmosphere. Limitations of the instrument due to the flushing time of the ventilated cavity are avoided in a design described in which 67% of the end area is cut away without affecting the cavity Q. Operation is in the TE_{011} mode at 9·1 kMc/s.

621.317.335.3.029.64/.65 **866**

Measurement of the Complex Dielectric Constant of Liquids at Centimetre and Millimetre Wavelengths.—A. G. Mungall & J. Hart. (*Canad. J. Phys.*, Sept. 1957, Vol. 35, No. 9, pp. 995–1003.) A free-space method is described for the measurement of absorption and reflection coefficients. Results obtained for methyl and ethyl alcohol at 13 and 9 mm λ agree with those previously quoted by other authors.

621.317.351 : 621.316.825 **867**

The Measurement of the Dynamic Characteristics of Thermistors.—G. Barzilai. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, No. 3, pp. 343–348.) Oscillograms of voltage and current are given which were obtained in tests made at frequencies between 0·02 and 5 c/s. The method described, in which a double-beam c.r.o. is used, permits the direct observation of waveforms at the thermistor terminals.

621.317.39 : 531.74.07 **868**

Design, Performance and Application of the Vernier Resolver.—G. Kronacher. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1487–1500.) Description of an angle transducer consisting of a variable-coupling transformer with the primary winding on a rotor and the secondary winding on a stator. Repeatability within ±3 seconds of shaft rotation is possible.

621.317.39 : 534.154 **869**

Vibration Measurements.—J. T. Broch. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 575–579.) Principles of the technique and a particular $BaTiO_3$ accelerometer are described.

621.317.616 : 621.373.4.029.3 **870**

Audio-Frequency Sweep Generator.—R. Graham. (*Radio TV News*, Aug. 1957, Vol. 58, No. 2, pp. 63–65, 140.) The unit covers a range 30 c/s–20 kc/s in one sweep, three times a second. A crystal-controlled oscillator beats with a variable-frequency oscillator controlled by a reactance valve and a 3-c/s sawtooth waveform.

621.317.725.029.6 **871**

A Millivoltmeter for Ultra-high Frequencies.—C. C. Eaglesfield. (*Electronic Engng*, Dec. 1957, Vol. 29, No. 358, pp. 603–604.) The technique of measurement involves the use of a modulated v.h.f. source, the detection of the signal by a crystal diode probe and the subsequent amplification and measurement of the modulation.

621.317.729.1 **872**

Note on the Measurement of Gradient in the Electrolyte Tank.—A. Lepschy, U. Pellegrini & A. Ruberti. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, No. 3, pp. 327–335.) The suitability of various types of probes for plotting gradients in electrolyte tanks is discussed. A four-electrode type appears preferable; test results obtained with it are given and a specially designed probe holder is described.

621.317.77 **873**

Extended-Angular-Range Direct-Reading Phase Meter.—S. Bigelow & J. Wuorinen, Jr. (*Rev. sci. Instrum.*, Sept. 1957, Vol. 28, No. 9, pp. 713–717.) A direct-reading pulse-position comparison instrument is described which enables angles from 540° lagging to 540° leading to be measured to ±1° for input amplitudes from 0·1 to 100 V.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.2 : 621.396.9 **874**

Precise Measurement of Distance by Microwaves.—R. Hammond. (*Instrum. Practice*, Aug. & Sept. 1957, Vol. 11, Nos. 8 & 9, pp. 828–831 & 942–945.) A description of the 'Tellurometer' giving results of a series of operational tests. See 3250 of 1957.

531.76 : 621.374.3 **875**

One-Tenth-Microsecond, Multichannel Chronograph.—L. E. Bollinger. (*Aust. J. Instrum. Tech.*, Aug. 1957, Vol. 13, No. 3, pp. 97–104.) Circuit details and description of an instrument for measuring the velocity of combustion waves, using optical or insulated probes and Schmitt trigger, oscillator and scaling circuits.

534.23-8 : 620.179.1 **876**

Automatic Ultrasonic Inspection.—H. W. Taylor. (*J. Brit. Instn Radio Engrs*, Nov. 1957, Vol. 17, No. 11, pp. 649–661.) A description of flaw-detection equipment developed to replace earlier manual inspection.

621.384.613 **877**

The Development of Iron-Free Betatrons with an Operating Frequency of 2·5 and 8·0 kc/s.—G. Hentze. (*Ann. Phys., Lpz.*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 55–81.) Design details of a 2·5-kc/s and an 8-kc/s pulsed betatron incorporating air-cored inductors are given.

621.385.833 **878**

The Computation of Rotationally Symmetrical Potential Fields in Electron Lenses.—F. Lenz. (*Ann. Phys., Lpz.*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 82–88.) A modification of Regenstreif's theory (2500 of 1951) for the three-electrode lens is discussed and approximation methods are compared.

621.387.422 **879**

Boron Trifluoride Counters.—J. J. Beauval, S. Dousson & P. Prugne. (*Le Vide*, May/June 1957, Vol. 12, No. 69, pp. 208–214.) A description of the techniques used at Saclay for detecting neutrons.

621.398 : 681.142 **880**

A Simple Shaft Digitizer and Store.—Tiffany. (See 690.)

655.3.024 : 621-523.8 **881**

Colour Printing.—(*Electronic Radio Engr*, Jan. 1958, Vol. 35, No. 1, pp. 26–28.) Description of an electronic method of colour correction applied in block-making.

PROPAGATION OF WAVES

621.396.11 **882**

A Note on the Propagation of the Transient Ground Wave.—J. R. Wait. (*Canad. J. Phys.*, Sept. 1957, Vol. 35, No. 9, pp. 1146–1151.) Formulae are derived showing how idealized pulses, represented by ramp, step and pulse functions, are modified by propagation.

621.396.11 : 523.5 **883**

The Forward-Scattering of Radio Waves from Overdense Meteor Trails.—C. O. Hines & P. A. Forsyth. (*Canad. J.*

Phys., Sept. 1957, Vol. 35, No. 9, pp. 1033–1041.) An approximate formula has been obtained by using a simplified working model. The received power is found to vary as the square root of the initial line density of electrons and the transition between underdense and overdense trails occurs at the same value of charge density as in the backscatter case.

621.396.11 : 523.5 : 621.396.43 **884**

Some Airborne Measurements of V.H.F. Reflections from Meteor Trails.—J. P. Casey & J. A. Holladay. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1735–1736.) The probability of simultaneous reception of meteor bursts at separated receivers was investigated using one fixed receiver and one installed in an aircraft.

621.396.11 : 523.72 **885**

On the Effect of Solar Ultra-radiation on Radio Propagation Conditions on 23rd February 1956.—B. Beckmann, P. Dietrich & H. Salow. (*Nachrichtentech. Z.*, July 1957, Vol. 10, No. 7, pp. 329–334.) Report of observations carried out under conditions favourable for assessing the effects of high-energy corpuscular radiation. Results of measurements obtained in various parts of the northern hemisphere are analysed and discussed.

621.396.11 : 551.510.52 **886**

The Possible Transmission Band for Long-Range Tropospheric Propagation.—V. N. Troitski. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 3–7.) The distortion of transmission is considered on the assumption that the atmosphere is anisotropic and that the horizontal inhomogeneities of permittivity are smaller than the vertical inhomogeneities. Formulae are derived for determining the band for distortionless transmission. The effect of the directivity of aerials on the possible transmission band is analysed. See also 1216 of 1957.

621.396.11 : 551.510.535 **887**

Brief Outline of Modern Concepts on the Propagation of Radio Waves in the Ionosphere.—Ya. L. Al'pert. (*Izv. Ak Nauk S.S.S.R., Ser. geofiz.*, Nov. 1957, No. 11, pp. 1418–1430.) The influence on wave propagation of earth curvature and heterogeneity of surface is discussed. The effect of ionospheric inhomogeneity and propagation in the troposphere is considered. 56 references.

621.396.11 : 551.510.535 **888**

Back-Scatter Soundings: an Aid to Radio Propagation Studies.—A. F. Wilkins & E. D. R. Shearman. (*J. Brit. Instn Radio Engrs*, Nov. 1957, Vol. 17, No. 11, pp. 601–616.) A comprehensive account of the radar technique for studying ionospheric propagation. Theoretical and practical aspects are treated. Results obtained at Slough at frequencies between 10 and 26 Mc/s are discussed; seasonal variations of echo patterns, very-long-range scattering, and errors due to aerial beam width are included. The utility of the rotating-aerial system is stressed.

621.396.11.029.62 **889**

V.H.F. Propagation Measurements in the Rocky Mountain Region.—R. S. Kirby, H. T. Dougherty & P. L. McQuate. (*Trans. Inst. Radio Engrs*, July 1956, No. PGVC-6, pp. 13–19. Abstract, *Proc. Inst. Radio Engrs*, Sept. 1956, Vol. 44, No. 9, p. 1213.)

621.396.11.029.62 : 551.510.535 **890**

The Scattering of Radio Waves of Very High Frequency in the Ionosphere.—T. V. Ionescu. (*C. R. Acad. Sci., Paris*, 29th July 1957, Vol. 245, No. 5, pp. 520–522.) V.h.f. transmission over distances greater than 1 000 km [see 243 of 1956 (Bailey et al.)] can be explained on the basis of the large number of negative oxygen ions at heights between 50 and 90 km having natural periods of vibration similar to the wavelengths used.

621.396.11.029.62 : 551.510.535 **891**

N.B.S. Equatorial-Region V.H.F. Scatter Research Program for the I.G.Y.—K. Bowles & R. Cohen. (*QST*, Aug. 1957, Vol. 41, No. 8, pp. 11–15.) Scattering from elongated centres in the F region over a 2 580-km path centred on the magnetic equator is to be attempted. Reception by amateurs at greater ranges is suggested.

621.396.11.029.62 : 621.396.82 **892**

The Occurrence of E_s and F_2 Skip in the 30–50 Mc/s Mobile Band.—E. W. Allen. (*Trans. Inst. Radio Engrs*, July 1956, No. PGVC-6, pp. 39–42.) Possible interference in the 30–50-Mc/s band allocated for mobile service is discussed, considering skip-distance/frequency curves and field strength data from earlier reports. For a typical 250-W base station with aerial 200 ft. above ground, interference via the F_2 layer from a similar station 2 200 miles away may reduce the median service radius from 60 to 12 miles. Interference via the E_s layer from a similar station 1 000 miles away may reduce the radius to 36 miles.

621.396.11.029.64 **893**

On the Radio Wave Propagation in a Stratified Atmosphere: Part 2.—R. Yamada. (*J. phys. Soc., Japan*, Sept. 1957, Vol. 12, No. 9, pp. 1022–1030.) The field produced by a microwave aerial is calculated assuming that the refractive index varies with height according to a second-order expression. The solution gives a series of rays reflected 1, 2, 3 or more times from the ground; these rays explain the 'duct' mode of propagation. A possible explanation is given for the deep fading associated with duct propagation. Part 1: 2083 of 1955.

621.396.11.029.64 : 621.3.018.7 **894**

Distortion in Tropospheric Scatter Propagation.—H. Bremmer. (*Philips Telecommun. Rev.*, Sept. 1957, Vol. 18, No. 3, pp. 137–154.) The Booker-Gordon theory is extended to transmitter currents that are not time-harmonic functions. The average intensity of the field scattered by a single 'blob' is first derived and the summation of the field contributions due to the total scattering volume leads to a 'convolution product'. This is shown to depend on the Fourier spectrum of the 'delay-time function' with which all distortion effects are connected. An approximate evaluation shows that the received field is equivalent to a small number of components with different amplitudes and delays. As a numerical example the distortion in f.m. frequency-division multiplex is found to depend on the relative amplitudes of the delayed signals.

RECEPTION

621.396.62 : 621.314.7 **895**

Midget Self-Contained Transistor Receiver.—S. F. Weber. (*R.S.G.B. Bull.*, Aug. 1957, Vol. 33, No. 2, pp. 66–68.) Circuit and construction details of a fixed-tuned medium-wave receiver incorporating a built-in ferrite rod aerial and a hearing-aid earpiece.

621.396.62 : 621.314.7 **896**

Transistorized Regenerative Receiver.—(*QST*, July 1957, Vol. 41, No. 7, pp. 36–37.) Circuit and constructional details of a two-transistor receiver suitable for the 80-, 40- and 20-m bands.

621.396.62 : 629.11 **897**

Car Radio Receiver Design.—J. C. Beckley. (*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 36–40.) A hybrid circuit for 12-V operation with transistor output.

621.396.621.54 **898**

The Interceptor.—C. W. Cragg. (*R.S.G.B. Bull.*, Aug. 1957, Vol. 33, No. 2, pp. 56–60.) Circuit details of a simple double-superheterodyne communication receiver for the amateur.

621.396.666 : 621.3.018.41(083.74) **899**

Fade-Cancelling Zero-Beat Indicator for Reception of Standard Radio Frequencies.—Blume. (See 857.)

621.396.823 **900**

Receiving Aerials and Industrial Interference.—V. V. Roditi & M. S. Gartsenshtein. (*Radiotekhnika, Mosk.*, Sept. 1956, Vol. 11, No. 9, pp. 21–27.) Methods are discussed for determining the effective height of indoor aerials and a coefficient of interference transfer as the main parameters determining the quality of radio reception in towns. Data are given on measurements of these quantities in some cities of the U.S.S.R., and the results obtained are analysed statistically.

STATIONS AND COMMUNICATION SYSTEMS

621.376.56 **901**

What Use is Delta Modulation to the Transmission Engineer?—F. K. Bowers. (*Commun. & Electronics*, May 1957, No. 30, pp. 142–147.) The signal/noise ratio and pulse rate of the delta modulation system are compared with those for p.c.m., and the coding and decoding arrangements are discussed. See also 3084 of 1955 (Zetterberg).

621.391 **902**

Information-Theory Impact on Modern Communications.—P. Mertz. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 431–437 ; *Elect. Engng, N.Y.*, Aug. & Sept. 1957, Vol. 76, Nos. 8 & 9, pp. 659–664 & 773–776.) The contributions made by Nyquist and Hartley and the concept of entropy are examined, and their influence on practical communications is outlined, with reference to facsimile, television and telephony systems.

621.391 **903**

Non-binary Error Correction Codes.—W. Ulrich. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1341–1388.) A theoretical study of the problem of correcting information which has become distorted by transmission via a noisy channel. Codes are derived for correcting any single unrestricted error in a message of arbitrary length and for correcting a number of errors in messages of restricted length, for an arbitrary number of different signals.

621.391 **904**

Shortest-Connection Networks and some Generalizations.—R. C. Prim. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1389–1401.) A consideration of the problem of interconnecting a given set of terminals with the shortest possible network of direct links. Simple graphical and computational methods are described.

621.391 : 621.396.822 **905**

Conditions for the Equivalence of the Statistical Properties of Radio Communication Systems with a Large Number of Random Parameters.—V. I. Siforov & Yu. B. Sindler. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Oct. 1957, Vol. 116, No. 6, pp. 956–958.) Short analysis concerning signal/noise ratio in radio relay and radiolocation systems. See also 1586 of 1957 (Sindler).

621.396.3 : 621.396.41 : 523.5 **906**

On the Influence of Meteor-Radiant Distributions in Meteor-Scatter Communication.—M. L. Meeks & J. C. James. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1724–1733.) An idealized distribution in which the radiants lie near the ecliptic is analysed and the results compared with previous calculations for a uniform radiant distribution. Some experimental data show evidence of a rather diffuse concentration of radiants near the ecliptic. A method for predicting the contributions of meteor showers to forward-scatter propagation is developed and applied to an example.

621.396.3 : 621.396.43 **907**

Storage Capacity in Burst-Type Communication Systems.—L. L. Campbell. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1661–1666.) Mean rate of transfer of information is derived, in terms of storage capacity, for known probability distributions of signal duration and interval between signals.

621.396.3 : 621.396.43 : 523.5 **908**

The Principles of JANET—a Meteor-Burst Communication System.—P. A. Forsyth, E. L. Vogan, D. R. Hansen & C. O. Hines. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1642–1657.) Propagation characteristics and design considerations of the system are surveyed and preliminary operating experience is summarized.

621.396.3 : 621.396.43 : 523.5 **909**

Bandwidth Considerations in a JANET System.—L. L. Campbell & C. O. Hines. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1658–1660.) "It is shown that the mean rate of transfer of information increases with bandwidth, for bandwidths in the range currently contemplated, in spite of the consequent decrease in the duty cycle. A system designed to maintain a constant signal/noise ratio by varying the bandwidth with received signal power is discussed, and its advantage over a fixed bandwidth system is calculated."

621.396.3 : 621.396.43 : 523.5 **910**

The Canadian JANET System.—G. W. L. Davis, S. J. Gladys, G. R. Lang, L. M. Luke & M. K. Taylor. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1666–1678.) The equipment is designed for use with double side-band a.m. v.h.f. radio links having 3-kc/s bandwidths. Standard 60-w.p.m. teletype machines are used, the instantaneous transmission rate (obtained automatically in both directions whenever a meteor reflection giving adequate signal/noise ratio is present) being 1 300 w.p.m. The messages are initially stored in parallel on paper tape in standard 5-digit form. Over the radio link they appear as a p.p.m. code having two pulse positions for each digit of the teletype code, together with a synchronizing tone. At the receiver the message is reconverted from p.p.m. to p.c.m. and stored in parallel on magnetic tape. The read-out mechanism, operating continuously at 60 w.p.m., removes the message from the store and converts it into a standard 7½-digit serial code to operate the printer. An average information rate of 60 w.p.m. with an error rate of 0·09% has been achieved when using 500-W transmitters with 5-element Yagi aerials, over a 600-mile path in Canada.

621.396.3 : 621.396.43 : 523.5 **911**

The Utility of Meteor Bursts for Intermittent Radio Communication.—G. F. Montgomery & G. R. Sugar. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1684–1693.) Transmission experiments at v.h.f. in a 100-kc/s band show that about half the signal bursts are distorted by multipath effects.

621.396.3 : 621.396.43 : 523.5 **912**

A Meteor-Burst System for Extended-Range V.H.F. Communications.—W. R. Vincent, R. T. Wolfram, B. M. Sifford, W. E. Jaye & A. M. Peterson. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1693–1700.) Describes equipment transmitting teletype or speech information over an 820-mile path. A frequency-shift system is used for teletype with a 2-kW transmitter and 3-element Yagi aerials, the information being sent at 600 w.p.m. during bursts. S.s.b. is used for speech, which is transmitted at five times normal rate in a band 16·5 kc/s wide with a power of 1 kW. Magnetic-tape storage is used at the receiver in both cases.

621.396.3 : 621.396.43 : 523.5 **913**

Analysis of Oblique-Path Meteor-Propagation Data from the Communications Viewpoint.—W. R. Vincent, R. T. Wolfram, B. M. Sifford, W. E. Jaye & A. M. Peterson. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1701–1707.) Characteristics such as duration, interval between usable signals, aerial direction effects, diurnal rate and duty cycle, and rate of signal decay are presented.

621.396.3 : 621.396.43 : 523.5 **914**

An Investigation of Storage Capacity Required for a Meteor-Burst Communications System.—R. A. Rach. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1707–1709.) A theoretical analysis of systems having storage either with or without simultaneous read-out.

621.396.3 : 621.396.43 : 523.5 **915**

On the Wavelength Dependence of the Information Capacity of Meteor-Burst Propagation.—V. R. Eshleman. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1710–1714.) The wavelength dependence of the information capacity of meteor-burst propagation is approximately $\lambda^{2.7}$ to be compared with $\lambda^{4.7}$ for the continuous forward-scatter. The advantages to which this leads are pointed out.

621.396.3 : 621.396.43 : 523.5 **916**
: 621.397.2

Experimental Facsimile Communication utilizing Intermittent Meteor Ionization.—W. H. Bliss, R. J. Wagner, Jr, & G. S. Wickizer. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1734–1735.) Preliminary results obtained over a 910-mile path at 40 Mc/s gave encouraging results ; one frame was transmitted per meteor burst.

621.396.3 : 621.396.43 : 621.396.812.3 **917**

Intermittent Communication with a Fluctuating Signal.—G. F. Montgomery. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1678–1684.) Analysis of a Rayleigh distribution of signal amplitudes suggests that intermittent operation when the signal is high gives greater average transmission rates than with continuous operation for the same average message error. Binary f.m. and ph.m. are considered.

621.396.41 : 621.318.57.01 **918**

A Network containing a Periodically Operated Switch Solved by Successive Approximations.—C. A. Desoer. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1403–1428.) The analysis of a basic system, such as is used in multiplex working, consisting of two reactive networks connected for a time τ with switching period T. The ratio τ/T is taken as 10^{-2}, considered to be as small as practicable. Examples are given of the application of the method which involves less work than the rigorous treatment of Bennett (*Trans. Inst. Radio Engrs*, March 1955, Vol. PGCT-2, No. 1, pp. 17–22).

621.396.41 : 621.396.65 : 621.372.55 **919**

Experimental Transversal Equalizer for TD-2 Radio Relay System.—B. C. Bellows & R. S. Graham. (*Bell Syst. tech. J.*, Nov. 1957, Vol. 36, No. 6, pp. 1429–1450.) A correction system, based on the echo principle of a transversal filter, for correcting residual gain and delay distortions in television relay systems. Directional couplers are used for tapping and controlling the leading or lagging echo voltages, required for correction purposes, which are applied in the pass-band 60–80 Mc/s. Some details of the assembly and typical field trials are given.

621.396.43 : 523.5 : 621.396.96 **920**

Directional Characteristics of Meteor Propagation Derived from Radar Measurements.—V. R. Eshleman & R. F. Mlodnosky. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1715–1723.) The geometrical correspondence between the radar and oblique-path detection of meteors is considered, and radar measurements of range and azimuth are used to determine the best directions in which to point the aerial beams on particular oblique paths for maximum duty cycle. For a N–S path the beams should be pointed west of the path at night and east of the path during the day. For an E–W path, north during the morning and south during the evening.

621.396.43 : 621.396.11 : 523.5 **921**

Some Airborne Measurements of V.H.F. Reflections from Meteor Trails.—Casey & Holladay. (See 884.)

621.396.65.029.64 **922**

New Microwave Repeater System using a Single Travelling-Wave Tube as both Amplifier and Local Oscillator.—H. Kurokawa, I. Someya & M. Morita. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1604–1611.) This system uses a minimum number of vacuum tubes and requires no a.f.c. Output power, frequency stability and crosstalk are considered. See also 1212 of 1956 (Sawazaki & Honma).

621.396.931 **923**

A Narrow-Band Experimental F.M. Mobile Telephone System.—W. A. Miller. (*Commun. & Electronics*, May 1957, No. 30, pp. 98–100.) Trials on an experimental split-channel system transmitting on 35·52 Mc/s and receiving on 43·52 Mc/s show that it is feasible to reduce frequency deviation from 15 kc/s to 7·5 kc/s to double the number of channels available.

SUBSIDIARY APPARATUS

621.311.6 : 621.373.52 **924**

The Balanced Transistor D.C. Convertor.—J. Noordanus. (*Philips Telecommun. Rev.*, Sept. 1957, Vol. 18, No. 3, pp. 125–136.) A theoretical analysis of operation is given.

621.314.63 **925**

Current/Capacitance Characteristics of Metal Rectifiers.—Y. Moriguchi & A. Okazaki. (*Proc. phys. Soc.*, 1st Oct. 1957, Vo. 710, No. 454B, pp. 991–999.) Report of measurements and discussion of the reverse-current/voltage and capacitance/voltage characteristics of metal rectifiers at room and liquid-air temperatures.

621.314.63 : 546.28 **926**

Silicon Rectifiers. — E. Nitsche. (*Elektronische Rundschau*, July 1957, Vol. 11, No. 7, pp. 197–199.) The characteristics are given of an experimental rectifier unit rated at 0·5 A and 650 V peak inverse voltage. It is hermetically sealed in a metal case of 7 mm diameter.

621.314.632.1 : 546.56-1 **927**

The Influence of the Copper Raw Material on the Properties of Copper-Oxide Rectifiers.—F. Eckart & C. Fritzsche. (*Ann. Phys., Lpz.*, 15th Nov. 1956, Vol. 19, Nos. 1/2, pp. 19–30.) The characteristics of Cu_2O rectifiers were investigated as a function of manufacturing conditions and degree of purity of the copper used. Results are tabulated and show that rectifier characteristics are considerably affected by the oxygen content of the copper.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2 : 621.396.3 : 621.396.43 : 523.5 **928**

Experimental Facsimile Communication utilizing Intermittent Meteor Ionization.—Bliss, Wagner & Wickizer. (See 916.)

621.397.5 : 621.395.625.3 **929**

Status of Video Tape in Broadcasting.—H. A. Chinn. (*J. Soc. Mot. Pict. Telev. Engrs*, Aug. 1957, Vol. 66, No. 8, pp. 453–458.) Experiences in the use of magnetic tape recording for television and some effects produced by faults in the recording equipment or tape are described.

621.397.611.2 **930**

Wide-Screen Television.—S. Rosin & M. Cawein. (*J. Soc. Mot. Pict. Telev. Engrs*, July 1957, Vol. 66, No. 7, pp. 404–406. Discussion, p. 406.) In the 'scanoscope' system described an aspect ratio of 8×3 instead of 4×3 is used. A special lens (see 931 below) compresses an 8×3 scene into 4×3, which is photographed, transmitted, and expanded electronically in the monitor.

621.397.611.2 : 771.35 **931**

Anamorphic Lens System.—S. Rosin. (*J. Soc. Mot. Pict. Telev. Engrs*, July 1957, Vol. 66, No. 7, pp. 407–409.) The optical design of the anamorphic 'scanoscope' lens is described and details of its application are given.

621.397.62 : 621.317.7 : 621.373.52 **932**

A Transistorized TV Bar Generator.—Knight. (See 860.)

621.397.62 : 621.396.67 : 621.372.43 **933**

A Diplexer Two-Set Coupler.—Harris. (See 671.)

621.397.62 : 621.396.677.029.63 **934**

Television Aerials for Bands IV and V.—Strafford. (See 674.)

621.397.62.029.63 **935**

Reception on Band V.—(*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 7–10.) An introduction to u.h.f. circuit techniques for reception on 650 Mc/s.

621.397.62.029.63 **936**

Band V on a Turret Tuner.—P. R. Stutz. (*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 14–16.) Adaptation of an existing band-I/band-III television tuner for u.h.f.

621.397.621 : 621.385.832.032.2 **937**

Wobbled Scanning with a New C.R.T.—(*Radio TV News*, Aug. 1957, Vol. 58, No. 2, pp. 52–53.) Scanning lines are merged by applying a 25-Mc/s oscillation to a split grid in the c.r. tube.

621.397.621.2 : 535.623 : 621.385.832 **938**

Low-Voltage Colour-Tube Gun Assembly with Periodic Focusing.—P. H. Gleichauf & H. Hsu. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 63–69. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.397.7(71 + 73) **939**

Television Station List.—M. I. Schiller. (*Radio Electronics*, Jan. 1957, Vol. 28, No. 1, pp. 82–83.) A list of U.S. and Canadian stations correct to 1st December 1956 giving call sign, location and channel number.

TRANSMISSION

621.396.61 : 621.373.42 **940**

The Design and Construction of a Drive Unit for Amateur Use.—N. Shires. (*R.S.G.B. Bull.*, Aug. 1957, Vol. 33, No. 2, pp. 61-65.) Constructional details of a unit providing a constant output level between 3·5 and 3·8 Mc/s for A1, A2, A3, F1 or F3-type transmissions.

621.396.61 : 621.376.22 **941**

Controlled-Carrier Constant Modulation.—(*Short Wave Mag.*, August 1957, Vol. 15, No. 6, pp. 298–299.) A modulation system in which a rectified a.f. voltage is used to modulate the screen grid of a transmitter power amplifier stage. A suggested circuit is shown.

621.396.61 : 621.376.222 **942**

Transmitter Cost Trimmed by Series Gate Modulator.—R. H. Baer. (*Electronics*, 1st Nov. 1957, Vol. 30, No. 11, pp. 167–169.) In this type of modulator power economy is obtained by varying the average carrier level in step with the modulation level.

VALVES AND THERMIONICS

621.314.63 **943**
A Theory of Voltage Breakdown of Cylindrical *P-N* Junctions, with Applications.—H. L. Armstrong. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 15–16. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.) See also 3681 of 1957 (Armstrong et al.).

621.314.63 : 537.311.33 **944**
Impedance of Bulk Semiconductor in Junction Diode.—T. Misawa. (*J. phys. Soc. Japan*, Aug. 1957, Vol. 12, No. 8, pp. 882–890.) The small-signal impedance is shown to have an inductive component, using the low-level solution. This reactance is small compared with the resistive component but it becomes comparable to the junction impedance as the injection level rises.

621.314.63 + 621.314.7](083.74) **945**
I.R.E. Standards on Graphical Symbols for Semiconductor Devices, 1957.—(*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1612–1617.) Standard 57 I.R.E. 21.S3.

621.314.632 : 546.289 **946**
Minority - Carrier Current across Metal/Germanium Rectifying Contacts.—G. Mesnard & A. Dolce. (*C. R. Acad. Sci., Paris*, 1st July 1957, Vol. 245, No. 1, pp. 42–44.) The minority-carrier current is calculated as a function of the applied voltage, taking account of variable lifetime and surface recombination velocities calculated on the basis of recombination centres.

621.314.632 : 546.289 **947**
The Relative Contribution of Majority and Minority Carriers to the Current across Metal/Germanium Rectifying Contacts.—G. Mesnard & A. Dolce. (*C. R. Acad. Sci., Paris*, 8th July 1957, Vol. 245, No. 2, pp. 152–155.) Electron and hole currents at the contact and at the limit of the space-charge zone are evaluated as a function of the applied e.m.f. The results are applicable to emitter and collector point contacts of transistors.

621.314.632 : 621.314.7 **948**
Small-Signal Wave Effects in the Double-Base Diode.—J. J. Suran. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 34–43. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.) See also 3562 of 1956.

621.314.7 **949**
Operation and Manufacture of Transistors.—A. J. Oliphant. (*R.S.G.B. Bull.*, Sept. 1957, Vol. 33, No. 3, pp. 107–111.) The basic theory of semiconductors is outlined and its application to transistors is discussed. Manufacturing techniques in producing transistors are briefly described and their present limitations are reviewed.

621.314.7 **950**
Low-Injection-Level Behaviour and Base Width Measurement in Junction Transistors.—D. Long. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1219–1220.)

621.314.7 **951**
The Junction Transistor as a Charge-Controlled Device.—J. J. Sparkes & R. Beaufoy. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1740–1742.) See 308 of January.

621.314.7 : 621.317.3 **952**
Accurate Measurement of r_c and x_o for Transistors.—M. A. Melehy. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1739–1740.)

621.314.7 : 621.318.57 **953**
The Effect of Collector Capacity on the Transient Response of Junction Transistors.—J. W. Easley. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 6–14. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.) See also 4044 of 1957 (Macdonald).

621.314.7 : 621.372.57 **954**
Tandem Transistors with the Properties of Thermionic Valves.—H. E. Hollmann. (*Hochfrequenztech. u. Elektroakust.*, March 1957, Vol. 65, No. 5, pp. 149–159.) The relation of junction transistors to space-charge and transit-time valves is defined in terms of duality. The tandem transistor is almost equivalent to a space-charge valve; some of its applications are discussed (see also 3001 of 1956).

621.314.7 : 621.385.4 **955**
The Tetrode Power Transistor.—J. T. Maupin. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 1–5. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, pp. 896–897.)

621.314.7.012.8 **956**
Equivalent Circuits for Junction Transistors.—L. E. Jansson. (*Mullard tech. Commun.*, June 1957, Vol. 3, No. 25, pp. 151–160.) Elements representing each major process in a junction transistor are assembled to form a complete equivalent circuit, which is then converted to a conventional two-generator grounded-base T circuit; other commonly used equivalent circuits are ultimately derived from this.

621.314.7.012.8 **957**
Base-Width Modulation and the High-Frequency Equivalent Circuit of Junction Transistors.—J. Zawels. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 17–22. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.383.4 : 546.682.86 **958**
Cooled Photoconductive Detectors using Indium Antimonide.—D. W. Goodwin. (*J. sci. Instrum.*, Sept. 1957, Vol. 34, No. 9, pp. 367–368.) "Improvement in the sensitivity of InSb photoconductive cells which operate at room temperature can be achieved under certain conditions by cooling cells to temperatures below ambient. Details of the performance of such cooled cells are given."

621.383.42 **959**
Resensitization of Selenium Photocells at Low Temperatures by the Action of the Near-Infrared.—G. Blet. (*J. Phys. Radium*, Feb. 1957, Vol. 18, No. 2, pp. 121–127.) See 607 of 1957.

621.383.42 **960**
Internal Resistance and Capacitance of Selenium Photocells at Low Temperatures.—G. Blet. (*J. Phys. Radium*, May 1957, Vol. 18, No. 5, pp. 297–303.) Forward and reverse conductances of normal and reactivated cells have been measured as a function of the applied potential difference in the range 0·001–5 V and at 66°–300° K. Observed variations extended over a range of 1–10^5. See also 1266 and 2951 of 1957.

621.385.029.6 **961**
Potential-Minimum Noise in the Microwave Diode.—A. E. Siegman & D. A. Watkins. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 82–86. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, pp. 897–898.)

621.385.029.6 **962**
Linear Beam-Tube Theory.—C. C. Wang. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 92–106. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 898.)

621.385.029.6 **963**
Space-Charge Effects in Klystrons.—W. E. Waters, Jr. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 49–58. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.385.029.6 **964**
Cylindrical Reflex Klystron with Lecher System as Oscillatory Circuit.—J. Koch. (*Z. angew. Phys.*, Jan. 1957, Vol. 9, No. 1, pp. 1–8.) The design and construction of a new type of klystron are described. Tests to determine the optimum arrangement of the control-electrode system are discussed. Measured values are in agreement with calculated parameters. A further development of this klystron providing facilities for wide-range frequency tuning appears feasible.

621.385.029.6 **965**
Some Special Magnetrons.—(*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 17–22.) A simple theory of operation and its application to voltage-tuned, minimum-voltage and spatial-harmonic magnetrons, and to the problem of scaling.

621.385.029.6 **966**
Stability of a Cylindrical Electron Beam in Nonsinusoidal Periodic Magnetic Focusing Fields.—D. C. Buck. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 44–49. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.) See 2546 of 1954 (Mendel et al.).

621.385.029.6 **967**
Travelling-Wave-Tube Gain Fluctuations with Frequency.—S. A. Cohen. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 70–78. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.385.029.6 **968**

Development of a Medium-Power L-Band Travelling-Wave Amplifier.—L. W. Holmboe & M. Ettenberg. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 78–81. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.385.029.6 **969**

Electron Bunching and Energy Exchange in a Travelling-Wave Tube.—S. E. Webber. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 87–91. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 898.)

621.385.029.6 **970**

The Gain and Bandwidth Characteristics of Backward-Wave Amplifiers.—M. R. Currie & D. C. Forster. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. ED-4, No. 1, pp. 24–34. Abstract, *Proc. Inst. Radio Engrs*, June 1957, Vol. 45, No. 6, p. 897.)

621.385.032.213.13 **971**

Donor Diffusion in Oxide Cathodes.—R. W. Peterson. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1176–1181.) "The activation of oxide cathodes by chemical impurities present in the base nickel is analyzed using classical diffusion theory and a model in which the concentration of donors in the oxide particles is controlled by alkaline earth metal adsorbed on the oxide particles. Strong surface adsorption of the alkaline earth metals on the oxide crystal surfaces is indicated."

621.385.032.265 **972**

Current and Velocity Fluctuations at the Anode of an Electron Gun.—E. V. Kornelsen, R. F. C. Vessot & G. A. Woonton. (*J. appl. Phys.*, Oct. 1957, Vol. 28, No. 10, pp. 1213–1214.) Measurements of the high-frequency noise-current fluctuations are reported.

621.385.1 + 621.314.7 **973**

Valves, Transistors and Efficiencies.—(*Wireless World*, Jan. 1958, Vol. 64, No. 1, pp. 41–44.) A simple theoretical discussion of the efficiencies obtainable from valves and transistors, used, for example, in a.f. output stages and d.c. converters.

621.385.1 : 537.525.92 **974**

The Determination of Plane Space-Charge Fields as well as those of Circular and Spherical Symmetry by means of Simple Resistance Networks with Additional Current Sources.—G. Čremošnik & M. J. O. Strutt. (*Z. angew. Math. Phys.*, 25th Sept. 1957, Vol. 8, No. 5, pp. 329–360.) Theoretical consideration of the application of a network analogue, with details of practical measurements. See also 3358 of 1957.

621.385.832 : 535.37 **975**

Luminescence Decrease of Phosphor Screens by Electron Burn.—K. H. J. Rottgardt. (*Elect. Commun.*, June 1957, Vol. 34, No. 2, pp. 130–135.) English version of 3420 of 1954.

621.385.832 : 621.396.963.3 **976**

Operation and Performance of the 6866 Display Storage Tube.—E. M. Smith. (*RCA Rev.*, Sept. 1957, Vol. 18, No. 3, pp. 351–360.) Description of a storage tube for direct viewing with an exceptionally bright display of radar-type information for periods as long as a minute. Principles of operation, important design features and performance characteristics are discussed.

621.385.832 : 621.397.62 **977**

A Thin Cathode-Ray Tube.—W. R. Aiken. (*Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1599–1604.) In this tube, only a few inches thick, the beam is injected parallel to one edge and caused to pass through two right-angle deflections; the first sends the beam into the region between the front and back tube surfaces, and the second turns it into the phosphor-coated front surface. See also 2484 of 1955 and 588 of 1957.

621.385.832 : 77 **978**

A Thin-Window Cathode-Ray Tube for High-Speed Printing with 'Electrofax'.—R. G. Olden. (*RCA Rev.*, Sept. 1957, Vol. 18, No. 3, pp. 343–350.) Constructional details are given and performance tests described; writing speeds of over 10 000 characters per second appear possible. See also 1132 of 1955 (Young & Greig).

621.385.832.032.2 : 621.397.621 **979**

Wobbled Scanning with a New C.R.T.—(See 937.)

621.385.832.032.269.1 **980**

The Influence of Anode Voltage Penetration on the Performance of Cathode-Ray-Tube Guns.—W. F. Niklas. (*J. Telev. Soc.*, Jan./March 1957, Vol. 8, No. 5, pp. 186–190.) Change of spot size with intensity modulation, caused by lens strength alteration and space charge influences, is reduced by a gun of new design, whose mechanism and performance are discussed.

621.385.832.032.36 : 621.397.621.2 : 535.623 **981**

Bilayer Bichromatic Cathode Screen.—C. Feldman. (*J. opt. Soc. Amer.*, Sept. 1957, Vol. 47, No. 9, pp. 790–794.) Screens consisting of superimposed thin transparent layers have been formed by vacuum deposition. A screen consisting of $Al/CaWO_4$-W (blue)/ZnS-Mn (yellow)/ glass supported in a 4-in. c.r. tube is considered in detail. The laws of colour mixture are found to be obeyed. The basic ideas of a multilayer chromatic system are briefly discussed.

621.387 **982**

Deionization in Gas Triodes and Tetrodes.—E. Knoop. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 126–132.) The effect of external operating conditions on the various parameters controlling deionization is investigated for a number of commercial-type gas-filled valves. See also 1542 of 1953.

621.387 **983**

A Particular Characteristic of Gas Triodes under Relaxation Conditions.—J. Lagasse, R. Lacoste & G. Giralt. (*C. R. Acad. Sci., Paris*, 22nd July 1957, Vol. 245, No. 4, pp. 412–414.) The minimum voltage across a thyratron in a relaxation oscillator can become very low and almost zero for certain values of the time constant of the anode circuit.

621.387 **984**

Pulse Firing Time and Recovery Time of the 2D21 Thyratron.—J. A. Olmstead & M. Roth. (*RCA Rev.*, June 1957, Vol. 18, No. 2, pp. 272–284.)

621.387 : 621.318.57 **985**

A Novel Cold-Cathode Tube.—D. W. Hill. (*A.T.E. J.*, April 1957, Vol. 13, No. 2, pp. 147–150.) A current-operated cold-cathode tube for use in electronic telephone exchanges is described; it has a movable trigger electrode of magnetic material.

621.387 : 621.318.57 **986**

Applications of a New Type of Cold-Cathode Trigger Tube.—K. F. Gimson & G. O. Crowther. (*Electronic Engng*, Oct.-Dec. 1957, Vol. 29, Nos. 356–358, pp. 462–468, 536–545 & 591–596.) Discussion of the general characteristics of trigger tubes and the design of the Type Z803U which has a highly stable ignition voltage. Its applications in timing, protection and counting circuits and in self-extinguishing circuits such as a relaxation oscillator are described.

621.387 : 621.396.822.029.6 **987**

Gas-Discharge Noise Tubes in the Range of High Discharge Admittances.—H. Schnitger. (*Nachrichtentech. Z.*, May 1957, Vol. 10, No. 5, pp. 236–240.) The range of noise generators can be extended to lower frequencies by coupling the gas discharge to a delay line. Parameters of the equivalent circuit are calculated and are confirmed by measurement. The calibration of hard-valve noise generators in the range 100–1 000 Mc/s by means of gas-discharge tubes is outlined.

621.387.032.212.3 **988**

Oxide-Coated Cathode for Cold-Cathode Discharge Tubes.—T. Imai & N. Mizushima. (*Rep. elect. Commun. Lab., Japan*, May 1957, Vol. 5, No. 5, pp. 10–15.) The effect of glow discharge on the cathode is investigated and methods of preparing a suitable cathode surface for practical use are described.

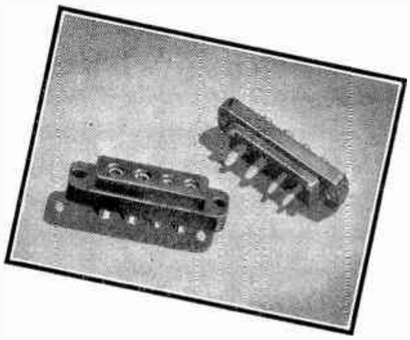
MISCELLANEOUS

538.56 : 621.37 **989**

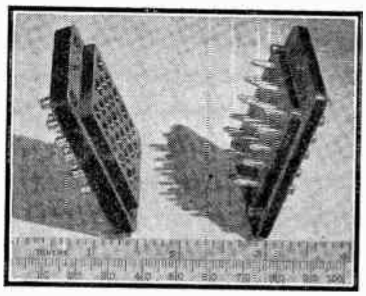
The Technique of Ultra-short Electromagnetic Waves since Heinrich Hertz.—F. W. Gundlach. (*Nachrichtentech. Z.*, July 1957, Vol. 10, No. 7, pp. 317–328.) A review article with 100 references.

001.891 : 621.396 **990**

Radio Research 1956: The Report of the Radio Research Board and the Report of the Director of Radio Research. [Book Review]—Publishers: H.M. Stationery Office, London, 1957, 47 pp., 3s. (*Nature, Lond.*, 28th Sept. 1957, Vol. 180, No. 4587, pp. 642–643.)

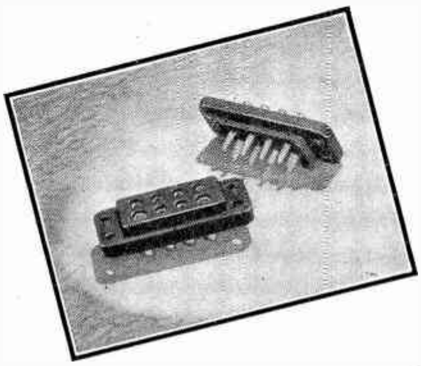


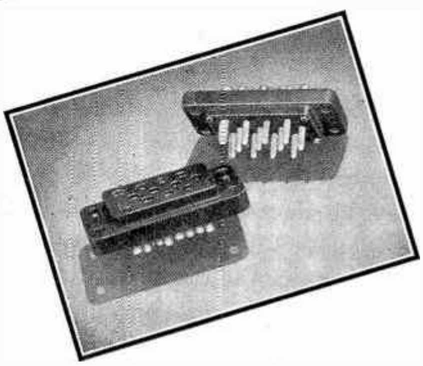
INTER-UNIT CONNECTORS WITH 4 TO 25 CONTACTS



"UNITORS"

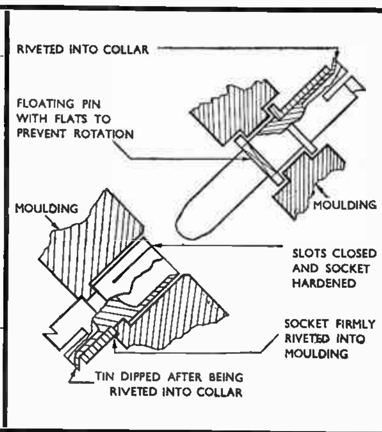
This popular range conforms to R.C.S. 321 and has Joint Service and A.R.B. approval. Pins are rated at three amperes, but each block has two larger pins rated at 10 amperes. All pins are fully floating. Each coupling consists of a block of plugs and another of sockets, arranged so as to be non-reversible. Bodies are moisture and tracking-resistant, being moulded from a nylon-filled phenolic material. Pins are of high-grade brass, silver plated. Sockets are of hardened beryllium copper; solder spills of both plugs and sockets are designed for ease of soldering. All plugs and sockets are numbered on both sides of the moulded body. The "Belling-Lee" "Unitor" is ideal for chassis mounting, or for flexible leads, when it is fitted with one of the range of die-cast light alloy covers or plug retainers.





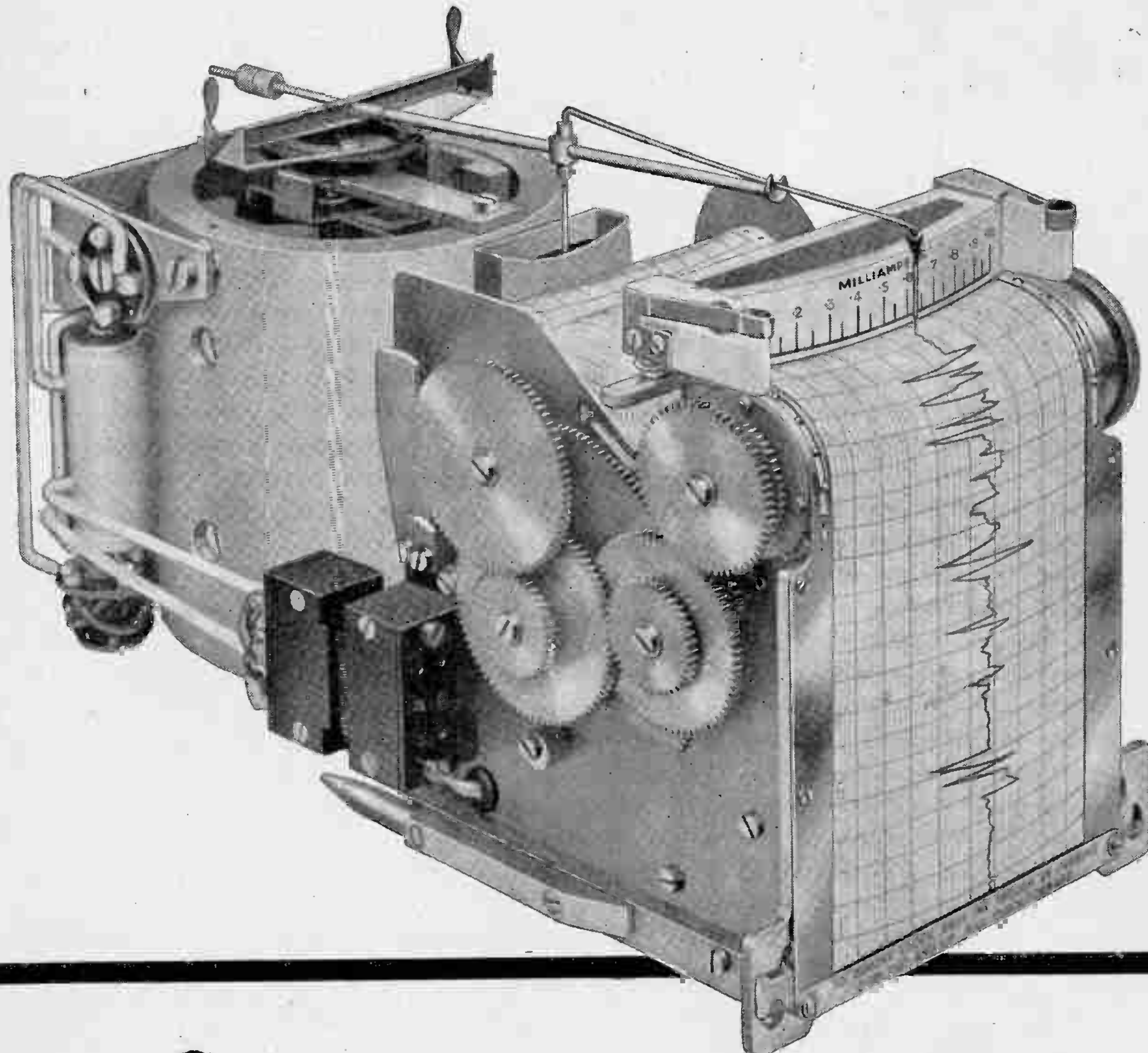
LIST NO.	PINS	BODY SIZE
L.653/P&S	4	1.437" x 0.406"
L.654/P&S	8	1.484" x 0.578"
L.655/P&S	12	1.500" x 0.843"
L.656/P&S	18	2.000" x 0.843"
L.1328/P&S	25 (4 large)	1.75" x 1.187"

Working volts: 500V. Peak, pin to pin
U.K. Pat. 649739
French Pat. 990017
Swiss Pat. 277799



BELLING & LEE LTD
GREAT CAMBRIDGE ROAD, ENFIELD, MIDDX., ENGLAND
Telephone: Enfield 3322 · Telegrams: Radiobel, Enfield

Inside story...



RECORD

See our Exhibits on Stand No. B9
ELECTRICAL ENGINEERS
A.S.E.E. EXHIBITION
Earls Court, March 25th—29th

GRAPHIC RECORDING INSTRUMENTS

are the fastest and most sensitive direct writing Recorders now available.

E.g. 0/100 microamperes, consumption 0·2 milliwatts
response time 0·4 second *
0/1·5 milliamperes, consumption 50 milliwatts,
response time 0·1 second *

*(To within 2% of indication)**

The secret of this remarkable performance lies in the generously proportioned movement which is designed to utilise every available cubic inch of a case which occupies the minimum of panel space.

A wide selection of other ranges can be quoted according to requirements, including Amperes, Volts, Watts, Vars, Frequency and R.P.M., also multi-range with Clip-on Transformer.

Also manufacturers of
"Cirscale" (reg. trade name)
Ammeters,
Voltmeters,
Wattmeters,
Frequency Indicators,
Power Factor Indicators,
and Rotary Synchroscopes,
Portable Testing Instruments,
Protective Relays,
"Cirscale"
Electric Tachometers, etc.

THE RECORD ELECTRICAL CO. LTD

" CIRSCALE WORKS," BROADHEATH, ALTRINCHAM, CHESHIRE

©8651-1

Offices at: Belfast, Birmingham, Bristol, Dublin, Glasgow, Leeds, London

ELECTRONIC & RADIO ENGINEER

CLASSIFIED ADVERTISEMENTS

Advertisement Rates. 6*d. per word, minimum charge* 6/-, *each paragraph charged separately. Remittances payable to Iliffe & Sons Ltd., Dorset House, Stamford St., S.E.*1. *Series discount of* 15% *is allowed for* 12 *consecutive insertions.*

Box Numbers *are available at an additional charge of* 2/- *to defray cost of registration and postage. Replies should be addressed Box No.* 0000, *c/o "Electronic & Radio Engineer", Dorset House, Stamford Street, London, S.E.*1.

Semi-Display *advertisements with centralized lines* £1/15/- *per single column inch—no additional charge for Box Numbers. The Publishers are not responsible for clerical or printers' errors although every care is taken to avoid mistakes.*

Press date for the April 1958 issue is first post 24th March 1958

SITUATIONS VACANT

A Senior Job for You with Hollerith Electronics

COMPUTER STORAGE PROBLEMS

WE recognise that this subject is among the outstanding problems in the electronic computer field. To work on it we need a senior circuit engineer to lead a team now exploring a novel technique using a magnetic drum.

THE essential qualification is experience in the magnetic recording field. The possession of an engineering degree is desirable.

IF you qualify, you will be fully supported by an enthusiastic group of fellow-workers, and will have at your disposal one of the best equipped and most modern electronics laboratories in the business machines field.

YOU will also have scope to develop your own ideas within the broad framework of our development programme.

TEAM MEMBERS: It is possible that some readers of this advertisement will be interested in the work mentioned and keen to make some contribution but may feel that they do not possess sufficient experience. We should still like to hear from them and to consider them (if otherwise suitable) for inclusion in the team.

ALL the positions, which are pensionable, carry good salaries commensurate with the work and experience required. Promotion, which is based on merit alone, can be rapid in this stable yet constantly growing company. Our laboratories are situated in pleasant, semi-rural surroundings at Stevenage, Herts. Help with housing can be given.

IF you are interested, write, with brief details about yourself, quoting Ref. No. C.S.9, to the Manager, Training and Personnel Division, The British Tabulating Machine Co., Ltd., 17 Park Lane, London, W.1. [1182

CHIEF DEVELOPMENT ENGINEER and DEVELOPMENT ENGINEERS

REQUIRED for work on development of Electronic and Electrical Measuring Instruments. Applicants for the position of Chief Development Engineer should be of degree standard and should have at least three years' experience in such development. The minimum qualification for the positions of Development Engineer is the Higher National Certificate. Applications are welcomed from men between the ages of 28 and 45 although more regard will be paid to qualifications than to age. Applications, giving full details of qualifications and experience should be addressed to:

Mr. L. R. Wratten
TAYLOR ELECTRICAL INSTRUMENTS, LTD.,
419/424 Montrose Avenue, Slough, Bucks

[1184

SENIOR Assistant Engineer required by Tanganyika Government Police Force for one tour of 30/36 months in first instance. Salary scale (including inducement pay) £1,383 rising to £1,566 a year. Gratuity at rate of 13½ per cent of total substantive salary drawn. Free passages. Liberal leave on full salary. Candidates, preferably not over 40 years of age, must have a wide knowledge of installation, running and maintenance of fixed and mobile radio communications equipment in the MF, HF and VHF categories, the erection and maintenance of lattice steel masts and towers, the installation, running and maintenance of low power and diesel generating equipment and a good knowledge of line equipment with teleprinter installation and associated practices. Candidates should have several years' experience and training up to degree or H.N.C. standard. Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters and full qualifications and experience and quote M2C/41982/EO. [1173

SITUATIONS VACANT

A Research Career with Hollerith Electronics

MAGNETIC CORE CIRCUITS

ARE you an electronics engineer looking for more scope for your ambitions? If so, you may qualify for a position with one of our expanding groups working on new ideas in the important field of magnetic circuits.

PREVIOUS direct experience in this particular field will not be regarded as essential, but if your experience does extend to pulse circuit techniques and you possess a qualification equivalent at least to an H.N.C., we may be able to offer you a senior position immediately. You will be given every opportunity to develop your own ideas within the area of the work being done.

A generous salary will be offered to the right man. The working conditions are ideal, in pleasant surroundings at Stevenage, Herts, and our electronics research laboratories are among the most modern and best equipped in the business machines and electronic computers industry. The position is pensionable and the prospects of advancement are excellent—with a stable, yet constantly growing company.

INTERESTED? Then write, with brief details about yourself (and quoting Ref. No. M.C.8), to The Manager, Training and Personnel Division, The British Tabulating Machine Co., Ltd., 17 Park Lane, London, W.1. [1180

SENIOR Technician required immediately for engineer's department, physics workshop. Knowledge of electronics, physics and general workshop technique desirable. Salary £600 × £25—£775 p.a., plus London weighting. Write, stating age, qualifications and experience to Chief Engineer, The Middlesex Hospital, W.1. [1177

DE HAVILLAND PROPELLERS, LIMITED, will be requiring Electronics Engineers in the next 2–3 months in connection with an advanced missile project. Experience of circuitry design and transistors is desirable but not necessarily essential. Engineers with a knowledge of power system, Electronics and servo systems, will also be considered. Applicants should write giving full details to the Personnel Manager (Ref. 34), de Havilland Propellers, Limited, Manor Road, Hatfield, Herts. [1179

SCIENTIFIC Officers required by Ministry of Supply at research and development establishments mainly in south of England, for work in physics, electronics, electrical mechanical or aeronautical engineering, applied mathematics, metallurgy or chemistry. Qualifications: first- or second-class honours degree in appropriate subject, or equivalent. Candidates should indicate fields of work in which interested. Starting salary in range £595 to £1,080 p.a. (superannuable) according to research experience. Rates for women somewhat lower but reaching equality in 1961. Forms from M.L. & N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting A.8/8A. [1174

INSTRUCTOR Grade I required by Posts and Telegraphs Department, Nigeria Federal Government, on contract for one tour of 12/24 months in first instance. Salary scale (including inducement addition) £1,536 rising to £1,674 a year. Outfit allowance £60. Free passages for officer and wife. Assistance towards children's passages and grant up to £150 annually towards maintenance in United Kingdom. Liberal leave on full salary. Candidates must be qualified training officers with experience in Carrier and Voice Frequency Telegraph equipment working on both line and radio. C. and G. Certs. in Telecomms. Principles III and Line Transmission I an advantage. Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2C/41922/EO. [1172

SITUATIONS VACANT

More scope for the Ambitious with Hollerith

TRANSISTOR COMPUTERS

WE are expanding our work in this field at our Electronics Research Laboratories, Stevenage, Herts. If you are a first-class circuit engineer we can offer you a first-class senior position on this project—and a good salary.

YOU should be up to at least H.N.C. standard, and you should have had experience of pulse circuits. It is also essential for you to possess the creative ability and the right kind of initiative to cope with the challenging nature of the problems encountered.

WE do not anticipate, however, that lack of direct experience with transistors will present any serious difficulty to the type of engineer we need.

THE working conditions are as good as we know how to make them—after fifty years of earning an excellent reputation for considerate employment, for sound products and for steady expansion. We offer full scope for ability—and advancement with us can be rapid.

THE position is pensionable. Assistance with housing can be given.

WE will tell you more about the possibilities of this interesting job if you will write (quoting Ref. No. T.C.9), to the Manager, Training and Personnel Division, the British Tabulating Machine Co., Ltd., 17 Park Lane, London, W.1. [1181

DRAUGHTSMAN. The G.E.C. Service Depot requires a junior draughtsman for small quantity production, electronics, etc. Applicants should have a knowledge of radio up to O.N.C. Standard. Apply to the Manager, G.E.C. Service Depot, Greycoat Street, Westminster, S.W.1. [1176

ENGINEER. The G.E.C. Service Depot has a vacancy for a radio and television field service engineer, to cover south London territory. Applicants should reside in the area, hold a clean driving licence and have had at least three years' experience of television field service. Apply to the Manager, G.E.C. Service Depot, Greycoat Street, Westminster, S.W.1. [1175

SERVICE

ASSEMBLY and Winding Capacity Electronic Equipment supervised by technicians experienced in modern technique applicable to computers, V.H.F. and high voltage oscillatory circuits. D.T.V. Contracts Dept., 134/136 Lewisham Way, New Cross, S.E.14. TIDeway 6666, Ext. 1. [1046

SATURABLE Reactors, Amplifiers and Automatic Control Equipment manufactured to customers' requirements. Let us develop and produce the complete equipment for you. Moderate charges, prompt service and delivery. Able Engineering Co., Ltd., 6 Singer Street Chambers, Singer Street, London, E.C.2. Telephone CLE 3695. [1183

PATENTS

PATENT No. 718956, Electronic Thermostat. The proprietors of Patent 718956 desire to enter into arrangements by way of licence or otherwise on reasonable terms to ensure its full development and commercial working in this country. Address inquiries to W. P. Thompson & Co., Chartered Patent and Trade Mark Agents, 12 Church Street, Liverpool, 1. [1178

Index to Advertisers

	PAGE
Adcola Products, Ltd.	8
Advance Components, Ltd.	11
Aero Research, Ltd.	15
Anderton Springs, Ltd.	36
Appointments Vacant	33, 34, 35
Bakelite, Ltd.	12
Belling & Lee, Ltd.	27
British Electric Resistance Co., Ltd.	8
British Insulated Callender's Cables, Ltd.	Cover ii
British Thomson-Houston Co., Ltd.	14, 33, 34
Brown, N. C., Ltd.	36
Bush Radio, Ltd.	34
C.J.R. Electrical & Electronic Development, Ltd.	33
De Havilland Propellers, Ltd.	34
Dubilier Condenser Co. (1925), Ltd.	29
Electrical Instruments (Hillington), Ltd.	18
Electronic Components	30
English Electric Valve Co., Ltd.	19
Ferranti, Ltd.	5
Foyle, W. & G., Ltd.	32
General Electric Co., Ltd.	2
Gilby-Brunton, Ltd.	32
Holiday & Hemmerdinger, Ltd.	36
Iliffe & Sons, Ltd.	32
Imperial Chemical Industries, Ltd.	9
Industrial Exhibitions	20
Institution of Electrical Engineers	33
Leland Instruments, Ltd.	Cover iv
Lewis Spring Co., Ltd., The	22
London Electric Wire Company and Smiths, Ltd.	18
Lyons, Claude, Ltd.	1
Magnetic Devices, Ltd.	17, 21
Marconi's Wireless Telegraph Co., Ltd.	34
Metal Gravure Co., Ltd.	32
Mullard, Ltd.	13, 26
Murex, Ltd.	22
Nagard, Ltd.	20
Newmarket Transistors, Ltd.	23
Parsonage, W. F., & Co., Ltd.	32
Pitman, Sir Isaac, & Son, Ltd.	33
Plessey Co., Ltd.	4, 16
Radiospares, Ltd.	3
Record Electrical Co., Ltd.	28
Salford Electrical Instruments, Ltd.	7
Shape Air Defence	34
Stratton & Co., Ltd.	10
Siemens Edison Swan, Ltd.	Cover iii
Sullivan, H. W., Ltd.	24
Technograph Electronic Products, Ltd.	31
Telephone Mfg. Co., Ltd.	6
Tinsley, H., & Co., Ltd.	30
Vactite Wire Co., Ltd., The	18

Printed in Great Britain for the Publishers, Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1, by Gibbs & Bamforth, Ltd., St. Albans. Distributed in U.S.A. by Eastern News Company, 306 West 11th Street, New York, 14.

Brion Leroux & Cie of Paris

have pleasure in announcing that their range of Sensitact Precision Measuring Relays and electrical meter movements are now available in the United Kingdom and British Commonwealth and Empire through LELAND LEROUX LIMITED, LONDON.

High sensitivity measuring relays with two adjustable contacts.

★ Moving-coil type movement for d.c. operation or with rectifier for alternating current of all frequencies.

★ Screened movement with patented spring suspension—standard octal valve base mounting.

★ Operating Power: from 0.01 micro-watts upwards. Accuracy better than 2 per cent.

★ Operating Response time: one to two-tenths of a second, depending on angular position of adjustable contacts.

★ Available as a polarized relay with centre zero, or as a minimum/maximum controller.

Automation and Communication Engineers are invited to write for details to:

LELAND INSTRUMENTS LTD.

Electronic Engineers & Consultants

TELEPHONE: VICTORIA 3243 (FIVE LINES)

22 & 23 MILLBANK · WESTMINSTER · LONDON S.W.1