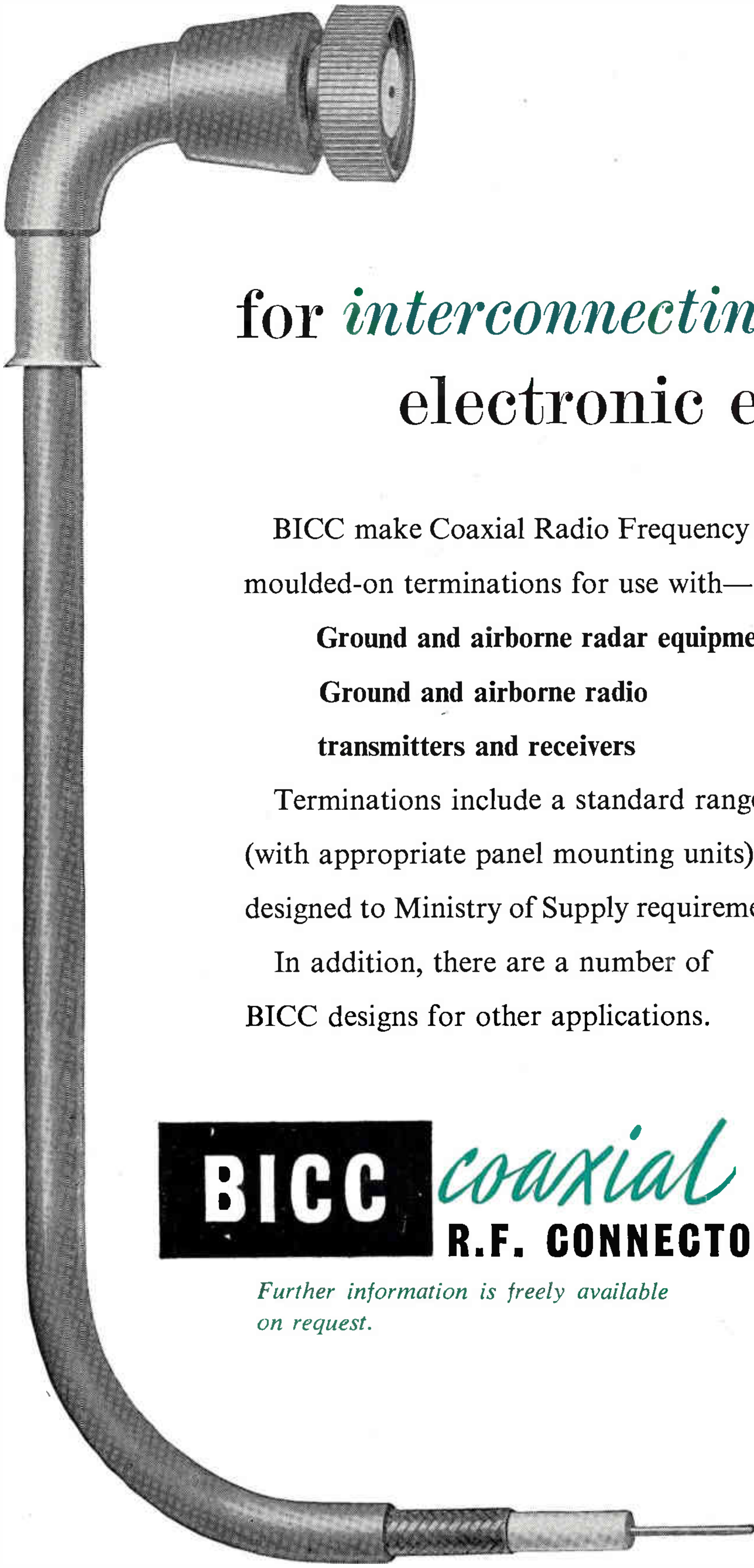
ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

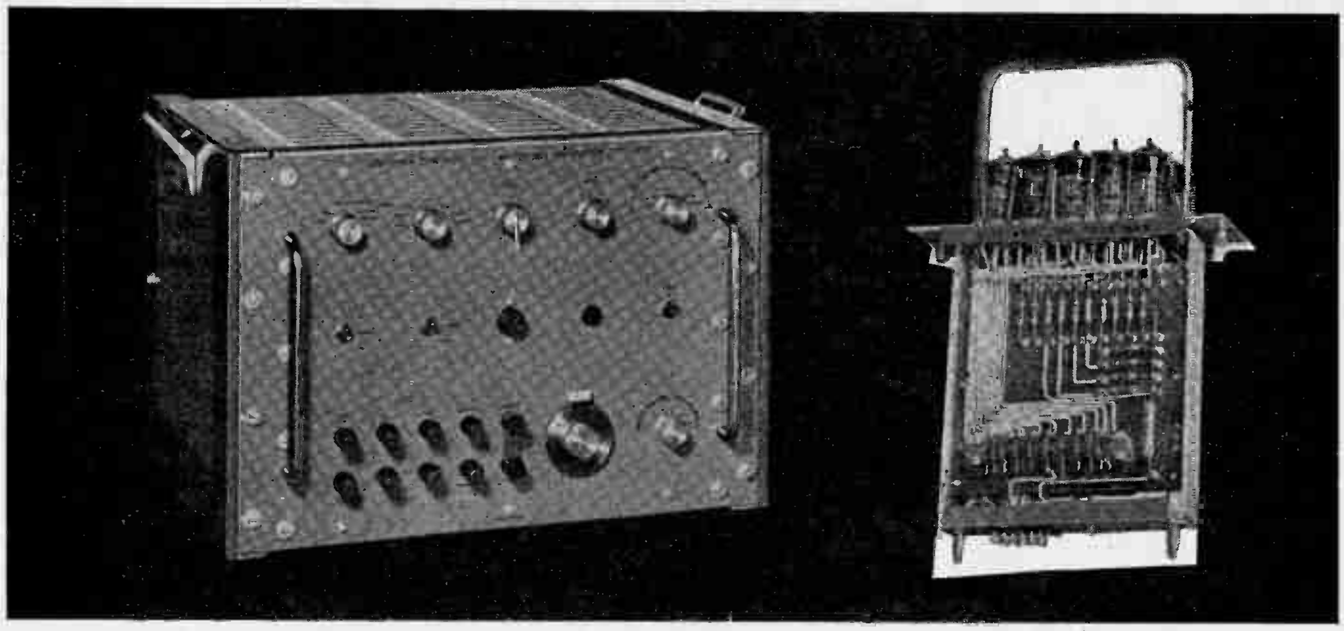
Three shillings and sixpence

MARCH 1959 Vol 36 *new series* **No 3**



Unchallenged

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



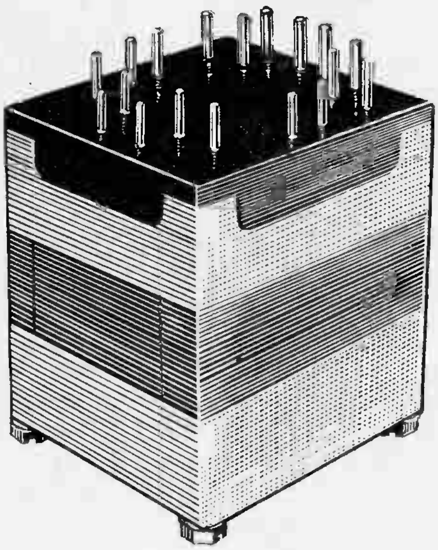
Still the most flexible supply of test signals for testing servo mechanisms and automatic controllers

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

- ★ *Sine waves from 500 cycles/sec. down to one cycle every 2,000 secs. (33 minutes)*
- ★ *Pulses and square waves with rise time of 10 microseconds*
- ★ *Ramp functions with a linear rise variable from 1 millisecond to 1,000 seconds*
- ★ *Over 30 different waveforms available*
- ★ *Voltage variable between 100 microvolts and 150 volts peak-peak*
- ★ *Load current up to 5 milliamps peak*

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

SERVOMEX CONTROLS LIMITED, CROWBOROUGH SUSSEX. CROWBOROUGH 1247



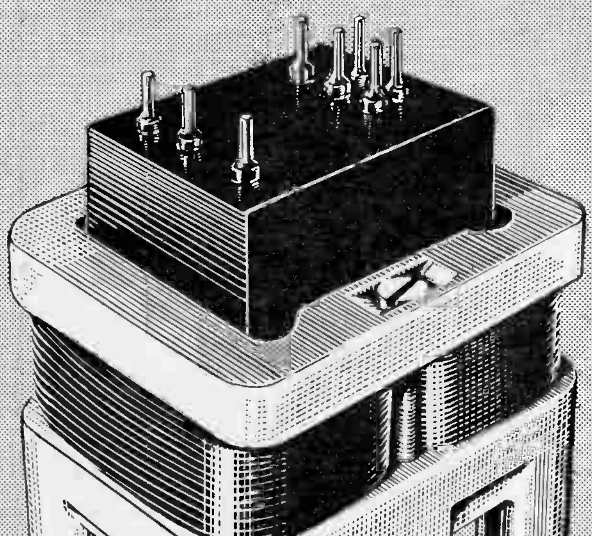
R120 and R130 laminated series.

FORTH SERIES

TRANSFORMERS AND CHOKES

smaller in size – lighter in weight

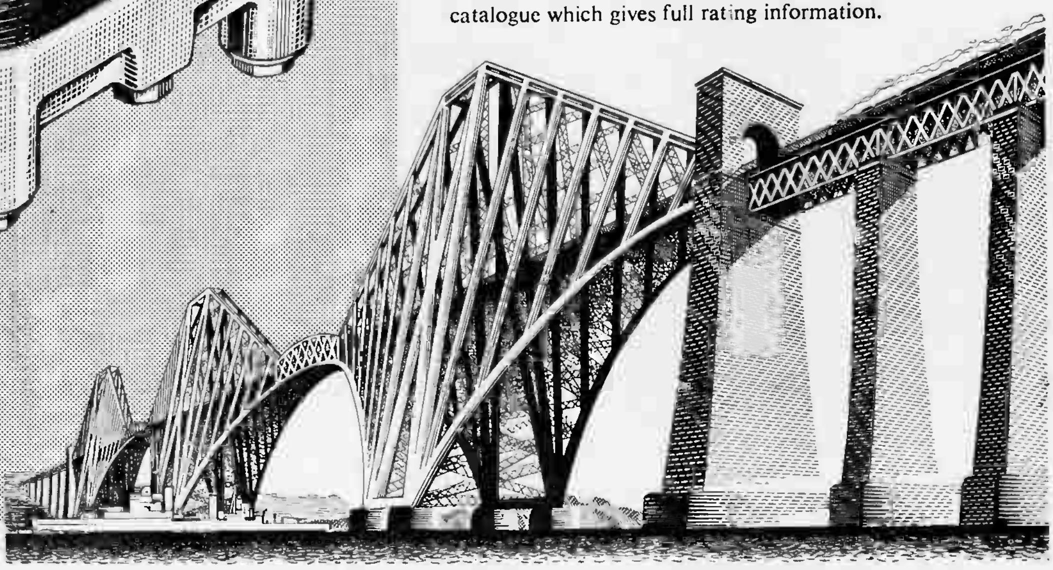
R200 'C' Core series.



FERRANTI

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

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



FERRANTI

FERRANTI LTD · FERRY ROAD · EDINBURGH 5

ES/T41

Captured on the screen for leisurely examination or for photography —that's the way the new Solartron Infinite Persistence Oscilloscope tackles a transient.

Two channels for correlation of interdependent phenomena, full-screen sweep in a few microseconds or a hundred seconds, push-button erasing and stabilised brilliance of display—these are just a few reasons why the Solartron IPO is quickly taking the lead in visual instrumentation of 'once-in-a-lifetime' and transient phenomena.

No need now to waste film waiting for non-recurrent signals, every frame exposed to your IPO screen will be significant—a certain step forward in your investigations.

From the specification, here are salient features which at once denote high sensitivity and an extremely wide field of application.

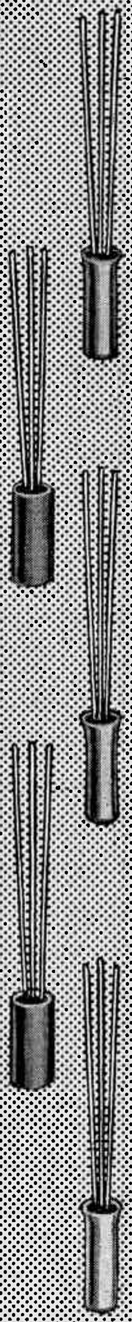
- ★ D.C. to 1 Mc/s bandwidth (3 db)
- ★ Maximum sensitivity 10 mV/centimetre
- ★ T.B. delay continuously variable to 100 Sec.
- ★ Identical 'X' and 'Y' input channels
- ★ Direct time and voltage calibration
- ★ Triggering in these modes:
  - (*a*) Repetitive operation
  - (*b*) Single-shot with 'lock-out'
  - (*c*) Push-button with 'lock-out'
  - (*d*) Delayed, using internal generator
  - (*e*) From signal, after adjustable delay period.

SOLARTRON

Write now for literature.

THE SOLARTRON ELECTRONIC GROUP LTD.

Thames Ditton, Surrey, England · Telephone: Emberbrook 5522 · Telegrams and Cables: Solartron, Thames Ditton · International Telex: 23842 (Solartron T.Dit.)



You can now specify S.T.C. Transistors for Entertainment and Industrial Applications

ENTERTAINMENT TYPES

(a)	TS.7	TS.8	...	...small signal amplifiers and frequency changers
(b)	TS.9	TS.17	...	...higher power amplifiers up to several hundred milliwatts
(c)	TS.13	TS.14	...	...low frequency applications

INDUSTRIAL TYPES

(a)	TK.23A		...	...low frequency telephone and telegraph carrier systems
(b)	TK.20B		...	...switching circuits (computers)
	TK.25B		...	...switching circuits 8 Mc/s and above
(c)	TK.21B	TK.24B		
	TK.26B	TK.27B		...switching circuits and/or small signal amplification
(d)	TK.40A		...	...amplifier and oscillator. Audio frequencies for power of several hundred milliwatts

First in Europe to produce transistors, S.T.C. are now making available to industry a range previously only used in their own equipment. Their long experience of components manufacture coupled with the latest production techniques has resulted in yet another high-quality product.

Send for literature on the available types.

Standard Telephones and Cables Limited

Registered Office: Connaught House, Aldwych, London, W.C.2

TRANSISTOR DIVISION: FOOTSCRAY · SIDCUP · KENT

ADCOLA
(Regd. Trade Mark)

Soldering Instruments and Equipment

Comprehensive Range of Models

P.V.C. Cable Strippers

Solder Dipping Pots

Supplied in ALL VOLT RANGES

A PRODUCT FOR PRODUCTION

RADIO, TV
RADAR
ELECTRONICS
TELECOMMUNICATIONS
ETC.

(Illustrated)

Protective Shield List No. 68

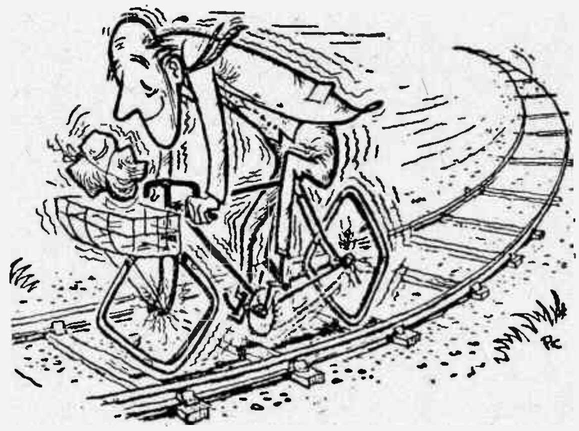
$\frac{3}{16}$" Detachable Bit Model List No. 64

Traditional British Quality and Workmanship

ADCOLA PRODUCTS LIMITED

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

VIBRATION



TESTING ?

Today's rapid increase in the speeds of Guided Missiles, Aircraft and various types of machinery has made the Engineer acutely aware of vibration.

Full scale testing is often expensive, though ultimately necessary, yet it is possible to assimilate the conditions under which parts will work by inducing controlled vibration and modifications can then be made at the design stage. In this field, W. Bryan Savage Ltd. have been actively engaged for the past twenty years and are one of the leaders in the manufacture of Power Amplifiers and Vibrators.

- If you have any problems regarding vibration, consult our Technical Dept.

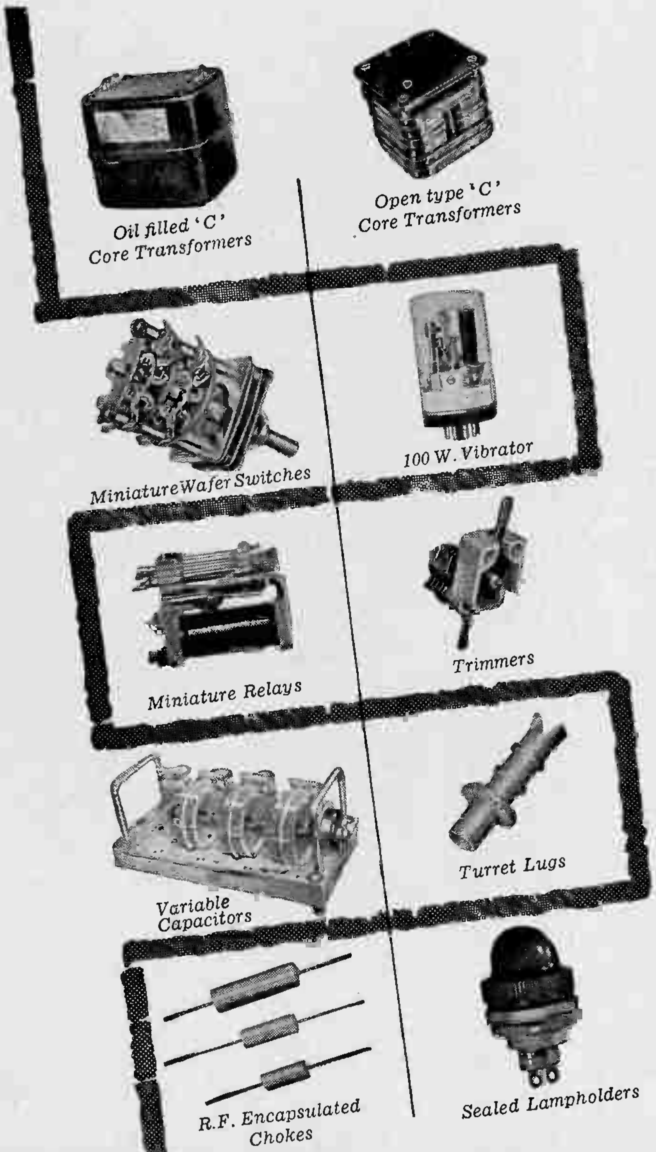
W. BRYAN SAVAGE LTD.

designers and manufacturers of vibrators for industry

17 STRATTON STREET, LONDON, W.1.

Telephone: Grosvenor 1926

DaS359ERE



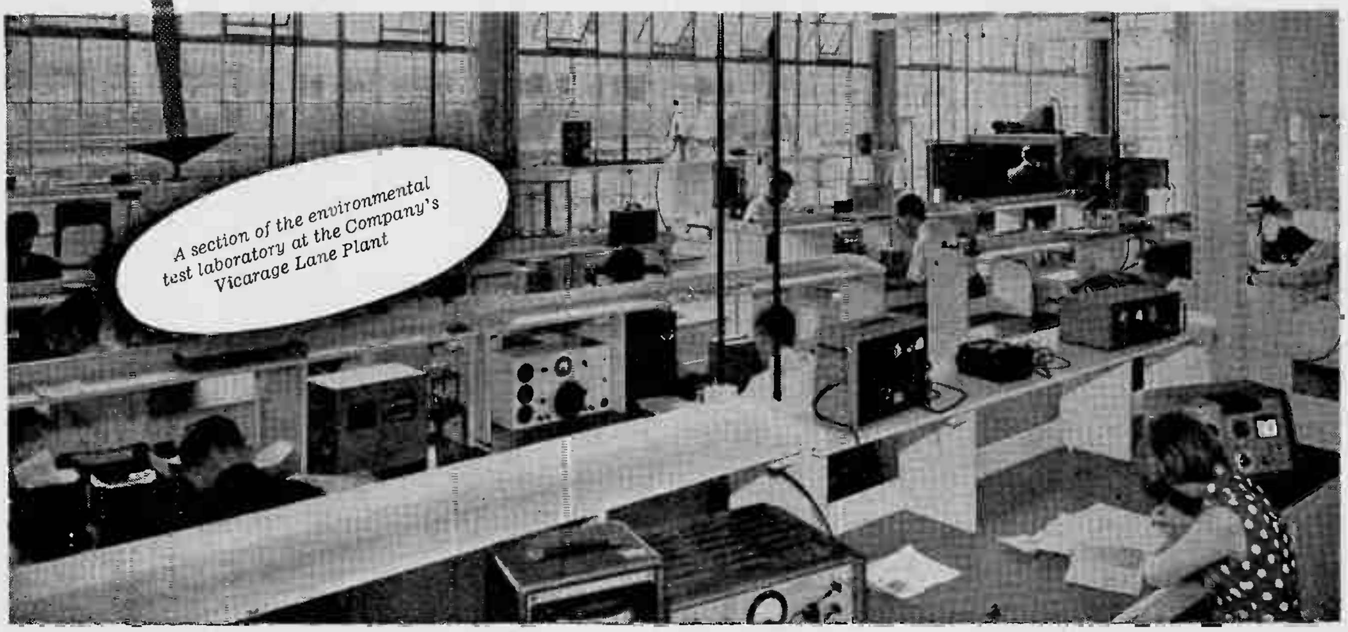
A lot depends on

Plessey

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

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

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



COMPONENTS GROUP | THE PLESSEY COMPANY LIMITED · ILFORD · ESSEX

TELEPHONE: ILFORD 3040

Overseas Sales Organisation: Plessey International Limited · Ilford · Essex · Telephone: Ilford 3040

CP2

Heathkit GENERAL PURPOSE 5" SCREEN OSCILLOSCOPE KIT

FOR LABORATORY, PRODUCTION OR SERVICE USE

SPECIAL FEATURES

- ★ 5" flat-face C.R. tube
- ★ Gold-plated printed-circuit board
- ★ 'Y' bandwidth: 3 c/s to 5 Mc/s
- ★ 'Y' sensitivity: 10 mV per cm
- ★ 'X' bandwidth: 1 c/s to 400 kc/s
- ★ 'X' input impedance: 30 MΩ, shunted 31 pF
- ★ Time base: 10 c/s to 500 kc/s

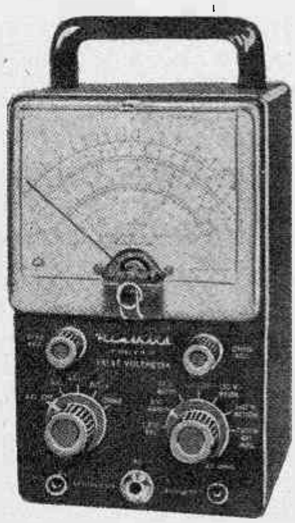
Using only components of the highest possible quality, the versatile model 0-12U Oscilloscope has many features hitherto exclusive to very much more expensive instruments . . . wide bandwidth, stabilised power supply, automatic lock-in synchronising circuit, unusually large, easy-to-read, flat-face screen, exceptionally wide range time base . . . etc. The use of printed circuit boards, together with preformed cable harness greatly simplifies wiring and assembly and guarantees consistency. The extremely comprehensive manual gives clear "step-by-step" instructions together with large pictorial diagrams.

MODEL 0-12U OSCILLOSCOPE KIT

£34 . 15 . 0

DELIVERED FREE, U.K.

You get an excellent instrument for much less money when you buy a Heathkit and the simple 'step-by-step' instructions make it so easy for anyone to build. Deferred terms available.



MODEL V-7A

VALVE VOLTMETER KIT

£15 . 14 . 0

DELIVERED FREE, U.K.

Because of its accuracy, reliability and overall quality, this instrument is outselling all other valve voltmeters. The precision and quality of the components used in this VVM cannot be duplicated at this price elsewhere. Its attractive appearance and performance will make you proud to own it. A large 4½" panel meter, clearly calibrated for all ranges, is used for indication. Front panel controls comprise a rotary function switch, a rotary range selector switch, zero-adjust and ohms-adjust controls. The voltage divider circuit uses 1% precision resistors. A printed-circuit board is employed for most of the circuitry, cutting assembly time and eliminating the possibility of wiring errors. It also ensures duplication of laboratory instrument performance.

This multi-function VVM will measure AC voltage (RMS) AC voltage (Peak-to-Peak) DC voltage and resistance. There are 7 AC (RMS) and DC voltage ranges of 1.5, 5, 15, 50, 150, 500 and 1,500. In addition there are 7 Peak-to-Peak AC ranges of 0-4, 14, 40, 140, 400, 1,400 and 4,000. Input impedance is 11 megohms. Seven ohm-meter ranges provide multiplying factors of X1, X10, X100, X1,000, X10K, X100K, and X1 megohm. Centre scale resistance readings are 10, 100, 1,000, 10K, 100K ohms, 1 megohm and 10 megohms. A centre-zero scale dB range is also provided. The battery and test leads are included in the kit. Shipping weight is 7 lb.

OTHER MODELS INCLUDE:		delivery free U.K.
DX-40U	Amateur Transmitter, 40w. to aerial	£29.10.0
S-88	Hi-Fi Stereo Amplifier, 8w. per channel	25. 5.6
S-33	Junior Stereo Amplifier, 3w. per channel	11. 8.0
SS-1	Hi.Fi Bookcase/Pedestal Speaker System	10. 9.0
UXR-1	Transistor Portable Receiver (in leather case)	17.17.0

FOR FURTHER DETAILS, PLEASE WRITE TO:

DAYSTROM LTD., DEPT. ER3, GLOUCESTER, ENGLAND

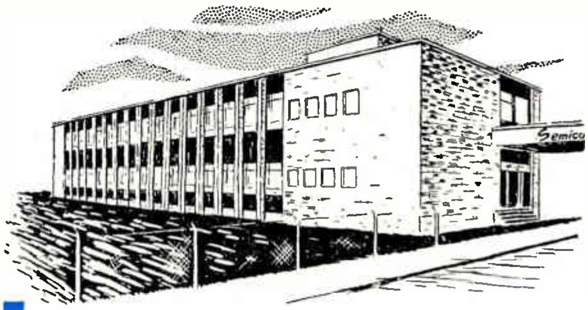
A member of the Daystrom Group, Manufacturers of

THE LARGEST-SELLING ELECTRONIC KITS IN THE WORLD

Semiconductors Limited

Applications Laboratory

making new equipment possible—existing equipment better . . .

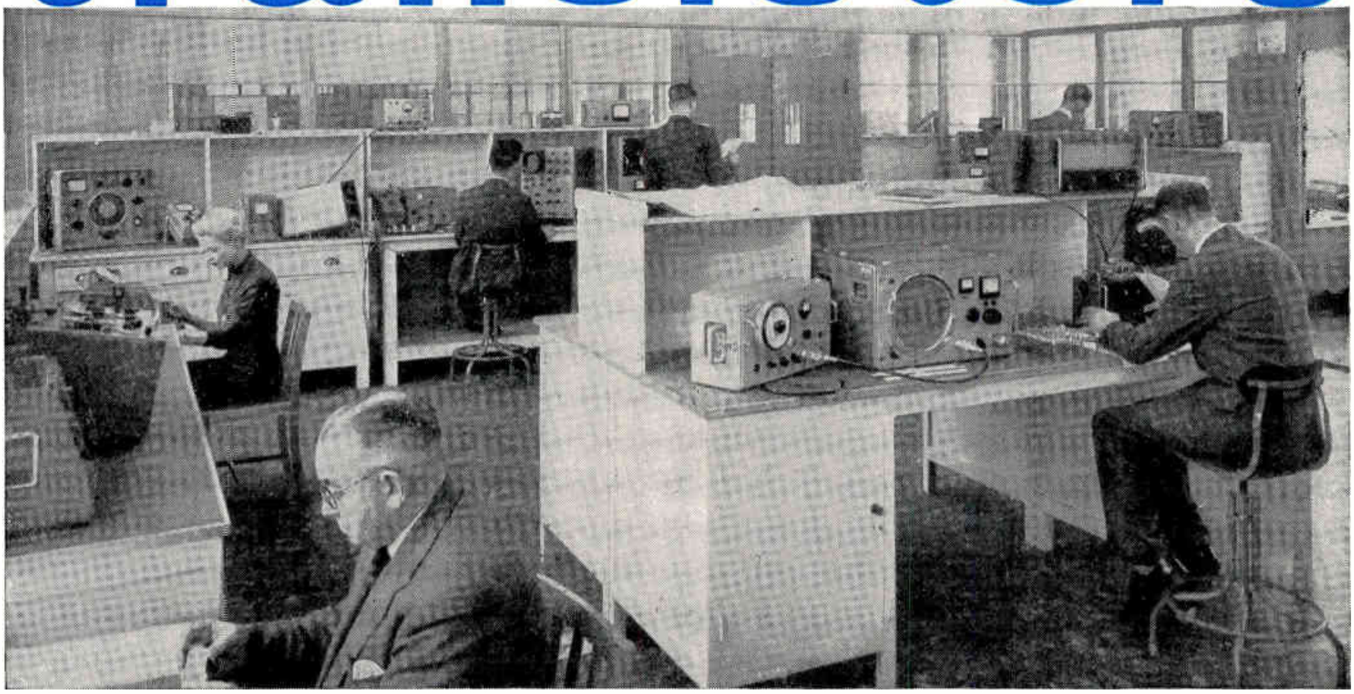


The Semiconductors Limited Transistor Applications Laboratory, with greatly improved facilities, is now installed in a new building adjacent to the Swindon factory. This will considerably increase its capacity to assist manufacturers in the correct selection and use of Semiconductors transistors. You are invited to take advantage of this service, either directly or through your Semiconductors representative.

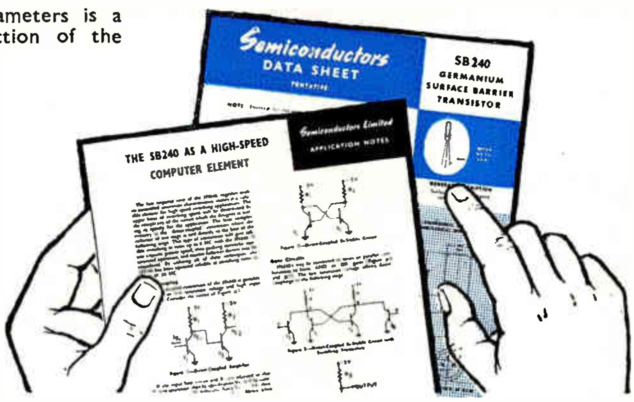
In addition to providing direct assistance to manufacturers, the Applications Laboratory is continuously investigating new approaches to existing problems.

For immediate assistance on your transistor problems telephone the Applications Laboratory SWINDON 6421.

transistors



The evaluation of transistor high-frequency parameters is a continuing study in this specially-equipped section of the Transistor Applications Laboratory.



Semiconductors Limited

CHENEY MANOR · SWINDON · WILTSHIRE

Telephone: Swindon 6421/4 Telegrams: Semicon, Swindon

** NEW APPLICATION NOTES AND DATA SHEETS ARE CONSTANTLY BEING ISSUED*
ARE YOU ON THE MAILING LIST?

SC.12

The best 75 volt stabilisers in the world

BRITISH SERVICES PREFERRED TYPE

M8225/CV4080

The high performance of the Mullard stabiliser 75C1 has led to the recent adoption of its Special Quality equivalent M8225/CV4080 by the British Services as their Preferred 75-volt stabiliser. The M8225/CV4080 is tested for specialised applications in which conditions of extreme shock and vibration are encountered.

- **Wide Current Range . . .**
  2 to 60 milliamps
- **Small Regulation Voltage . . .**
  Less than 9 volts
- **High Stability . . .**
  Typical variation in burning voltage less than ± 2% in any 10,000 hours of operation.

GENERAL PURPOSE TYPE

75C1

The 75C1 is the best 75 volt stabiliser available in the world for general purpose use in industry and communications. It has the same electrical characteristics as the M8225/CV4080 and like this British Services Preferred valve provides an exceptional *combination* of long life, stability and good regulation.

Full data is readily available from the address below.

MULLARD LIMITED
MULLARD HOUSE
TORRINGTON PLACE
LONDON · W.C.1
TEL: LANGHAM 6633

Mullard
GOVERNMENT AND INDUSTRIAL VALVE DIVISION

Mullard

MVT355a

ELECTRONIC & RADIO ENGINEER

- *incorporating* WIRELESS ENGINEER

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

CONTENTS

VOLUME 36 NUMBER 3 MARCH 1959

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 25210* : 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*

No. 12

Permanent Magnets

Design Advisory Service

Magnetic Attraction Applications – 1

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

The earliest known effect of permanent magnets is their ability to attract ferrous objects.

The attraction or holding power of a magnet under ideal conditions can be calculated from the basic formula :—

$$\text{Force in dynes} = \frac{B^2A}{8\pi}$$

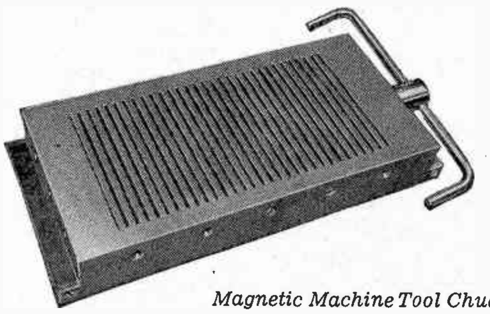
or

$$\text{Force in lb.} = \frac{B^2A}{11{,}263{,}000}$$

where A = *area of magnetic pole faces in cm²*
and B = *corresponding flux density in gauss.*

Under normal conditions joints, tolerances, misalignments and leakage flux rapidly reduce the theoretical pull; therefore the calculated value should only be used as a guide. One factor which must always be observed is that the magnet under the most severe conditions of open circuit should be of sufficient length to prevent self-demagnetisation —i.e. the maximum value of H should not exceed H_d.

This advertisement deals briefly with some of the many industrial applications using magnetic attraction, such as Machine Tool Chucks, Relays, Industrial Filtration, Door Catches, etc.



Magnetic Machine Tool Chuck

Machine Tool Chucks

One of the most useful applications of permanent magnets is in machine tool chucks where steel articles of a form extremely difficult to clamp, are held firmly in position for machining operations.

There are many types of magnetic chuck designed for various purposes but generally with relatively wide pole pitch spacing. The chuck illustrated has the advantage of small pole spacing and is particularly suitable for small or thin articles. It consists of thin 'Magnadur' blocks with mild steel pole plates assembled in sandwich form giving alternate poles 1/8″ apart.

The attractive force to iron and steel objects of not less than 1/16″ in thickness is approximately 130 lb./sq. in. The objects are released by moving the lower section of the chuck one pole pitch to short circuit or cancel the flux in the whole chuck.

Industrial Filtration Equipment

It is well known that one of the principal causes of wear in machinery is the presence of abrasive matter in the lubricating oil. A certain amount of the contamination to the oil can be prevented by careful design and dust covers but minute particles of steel can only be removed by the use of permanent magnets.

Relays and Thermostats

For current carrying relays and thermostats, contacts should open and close with a snap action and contact pressure must be sufficient to prevent chatter and arcing. It is in applications such as these that the attractive force of a magnet can be used to supply the necessary minimum contact pressure and also the desired degree of snap action.

Magnetic Fishing Tool

The tool shown is used for recovering broken rock drills or bits of iron or steel which accidentally get into deep boreholes. The one illustrated is of 14″ diameter, and uses a 'Ticonal' magnet capable of lifting over two tons. *Photograph by courtesy of D. F. J. Burns Co., Ltd.*

Magnetic Door Catches

Magnetic door catches can be designed to be extremely small, efficient and inexpensive. As an example, a 'Magnadur' magnet 0.89″ x 0.59″ x 0.18″, when fitted between mild steel pole plates, is capable of holding an armature with a force of between four and five pounds.

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

Mullard

'TICONAL' PERMANENT MAGNETS
'MAGNADUR' CERAMIC MAGNETS
FERROXCUBE MAGNETIC CORES

MULLARD LIMITED, COMPONENT DIVISION, MULLARD HOUSE, TORRINGTON PLACE, W.C.1. LANgham 6633
MC273

VOLUME 36 NUMBER 3 **ELECTRONIC & RADIO ENGINEER**

MARCH 1959 *incorporating* *WIRELESS ENGINEER*

Transistors Again

ELSEWHERE in this issue we include a letter criticizing our Editorial of November 1958 together with our comments upon it. Our critics' view appears to be that a competent transistor-circuit designer, who is able to tackle all classes of work, must have a good knowledge of semiconductor theory, and we agree with them.

Where we differ is about the proper way for the valve-circuit designer to make a start with transistors. Our critics consider he should start with semiconductor theory. We hold that he should ignore this for the time being and start right away designing transistor circuits. We think that the paper referred to in our November Editorial will be of enormous help to him in this. We also think that, with his valve-circuit experience, he will find no serious difficulty in designing a great many satisfactory transistor circuits. These circuits will not necessarily be like valve circuits. The application of proper design methods to transistor, or any other, characteristics will lead to appropriate circuits.

We do not say that he will never encounter difficulty. He will certainly do so when he becomes ambitious and it is then that he will need semiconductor theory to help him. He will realize it and be ready to undertake the quite considerable task of acquiring an understanding of it. He will be greatly helped in this, however, by the knowledge of transistor characteristics which he has gained in his simple design work.

It seems to us illogical to make semiconductor theory the starting point. A theory is an explanation of observed phenomena in terms of more fundamental concepts and both explanation and concepts are invented by man.

It seems to us unreasonable to expect anyone to grasp and understand concepts which have been invented to explain phenomena without first being familiar, not necessarily with all, but at least with some of the phenomena themselves.

Saturable-Transformer Switches

APPLICATION TO MAGNETIC-DRUM HEAD SELECTION

By Brian D. Simmons, A.M.I.E.E., A.M.Brit.I.R.E., M.I.R.E.*

SUMMARY. *Saturable magnetic cores may be used for non-linear switching. A switch which uses cores in a balanced saturable transformer is considered. The on-off impedance ratio obtained by suitable design is adequate to switch magnetic-recording heads for reading and writing operations. Details are given of the design and construction of a saturable transformer used in a selection system, in which a large number of magnetic-drum heads share common read and write amplifiers. Circuits are described for the operation of a matrix selection system and, in particular, a simple arrangement using transistors which switches heads in a few tens of microseconds.*

Magnetic drums are widely used for the storage of digital information in data-processing systems. They are able to store a large quantity of information in a permanent form by means of magnetized cells which occupy a relatively small space on the surface of the drum. Access time to any bit written on a track is the time taken for the drum to make one revolution and this is usually several tens of milliseconds. As relatively few magnetic cells are inspected or have their magnetized state changed at any one instant of time, it is common practice to share a number of heads, associated with the magnetic tracks, with a few common amplifiers. When this is done, a means must be found of switching a particular head to the amplifier. The period during which switching is taking place represents lost processing time and it is essential therefore that, in all systems which handle a mass of data at a high rate, the switching time must be minimized. In all but the simplest uses of a drum (e.g., as a permanent library) it is necessary to be able not only to read information from the tracks, but to write new information and modify that already recorded. The switches used to select heads must be capable of connecting them to either reading or writing amplifiers. The levels of the writing and reading signals may differ by as much as 60–80 dB.

Several methods of achieving a head-switching network have been described and are in common use. Relay and cross-bar selectors[1,2] can provide economic switch networks and give very good on-to-off ratios. A low-resistance metallic connection is provided in the selected path and all other heads are well isolated. They are relatively slow in operation, however, and contacts are not as reliable as one would wish. Selection systems using diode gates and thermionic valves or transistors, although fast, still suffer from limitations of impedance matching and reliability. Separate selection systems are frequently required to connect heads for the reading and writing operations[3]. This article describes a simple method of selection, based on a switch using a saturable transformer constructed from tape-wound cores of high-permeability material. The advantages associated with using iron-cored switches for this application are largely self-evident.

(a) They can be made faster in operation than relay switches
(b) are inherently robust and reliable
(c) can be designed to match almost any impedances
(d) provide complete electrical isolation between head and amplifier
(e) should handle both read and write signals adequately
(f) require only the one control circuit for both operations
(g) are modest in power consumption.

Magnetic-drum heads and core circuits are compatible and the overall advantages weigh heavily in favour of the adoption of such a system of head selection, provided the design is realizable.

An investigation was made of the available types of core materials, which could be used as the basis for design of a magnetic switch, capable of giving sufficient change in on-to-off impedance ratio to permit its use for the selection of magnetic-drum heads. This impedance ratio determines the change in output voltage from a correctly-terminated saturable transformer as it is switched, and is referred to as the switching ratio. Having obtained a satisfactory design, a switching network was then considered which would be capable of selecting one of a large number of heads for connection to common read or write amplifiers, using a minimum amount of control apparatus.

Principle of Operation

The simplest type of saturable-transformer switch shown in Fig. 1 (a) consists of a closed core with windings for control, excitation and output. It should be noted that where a switch is used as a both-way device, the functions of the excitation and output windings can be

* *Siemens Edison Swan Research Laboratory.*

interchanged. An idealized hysteresis curve of a typical core of high initial permeability is shown in Fig. 1 (b). Switching is accomplished by passing through the control winding a direct current of sufficient magnitude to take the core into saturation beyond point B. When the control winding current is zero, the core has no residual flux and is at point A. If an alternating voltage is applied to the excitation winding, normal transformer action takes place and a voltage appears at the output winding. If a direct current is passed through the control winding, a constant magnetizing force is superimposed on the alternating one and provided the resultant current always maintains the magnetizing force beyond point B, then the core is kept saturated. In this state the core permeability is essentially unity, negligible coupling exists between excitation and output windings and zero output voltage appears across the output winding.

In practice, the ideal cannot be obtained and reference to Fig. 2, which shows a typical hysteresis loop, indicates that the switched voltage ratio must be a finite quantity. For many applications, however, ratios of between 10 and 100 are tolerable, so making the practical design realizable. If a signal consisting of unipolar pulses is applied to the excitation winding, they will trace out a

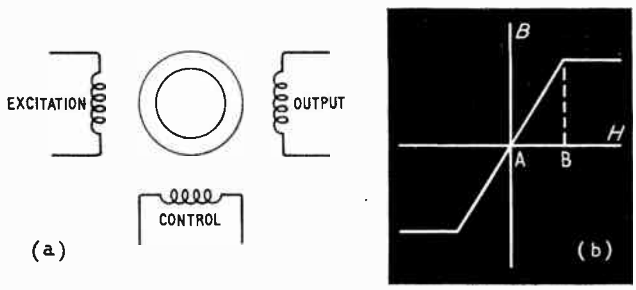


Fig. 1. *Basic saturable-transformer switch (a) and idealized B-H characteristics (b)*

minor hysteresis loop originating from the flux remanence point B_0. The size of the minor loop will depend on the amplitude of the input pulses (which will be small for reading and large for writing) and the degree of loading on the output winding, which gives rise to an opposing flux. When the core is saturated at B_s, the same change in H now produces hardly any change in B. The switching ratio of the transformer is governed by the ratio of the permeabilities of the core at B_0 and B_s, given by $\Delta B_0/\Delta H_0$ and $\Delta B_s/\Delta H_s$, which is simply μ_0/μ_s. The design problem largely resolves itself into finding a suitable core material which:

(a) has high incremental permeability at remanence
(b) saturates effectively at low levels of magnetizing force
(c) does not introduce significant hysteresis or eddy-current losses.

Before considering suitable types of core materials, however, a further important aspect of the saturable transformer must be examined.

The simple circuit consisting of one core and three windings permits interaction between control flux and

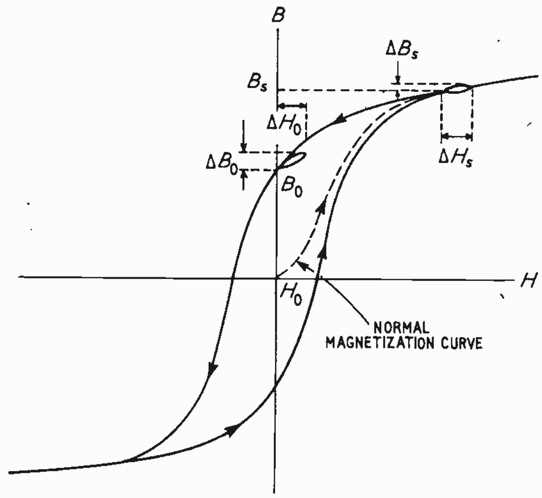
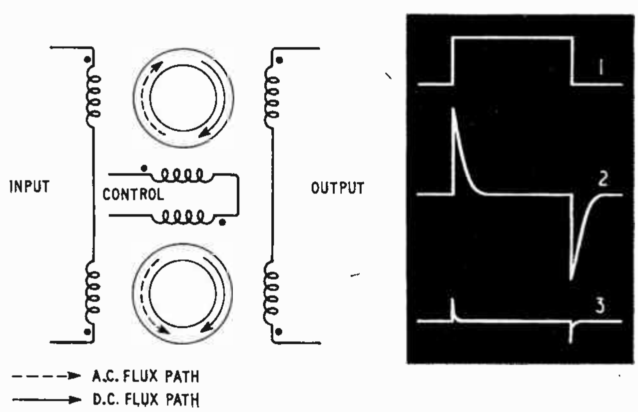


Fig. 2. *Typical hysteresis loop for core*

excitation flux. Rapid switching of control current is liable to introduce large transients into the control winding, which can cause spurious signals at the output and excitation windings. These signals may be of an amplitude sufficient to cause a change of state in the magnetic flux on a drum track and block or damage the read amplifiers. By choosing a balanced construction for the switch, as shown in Fig. 3, the control and excitation windings are adequately decoupled and no significant transients are produced in other windings.

However, to eliminate cross coupling completely, a very careful control of winding must be exercised during manufacture although, in practice, this extreme has not been found necessary as a small amount of break-through can be tolerated. A balanced construction is commonly employed in magnetic-amplifier design[4] and consists of winding the control turns on the centre limb of a three-legged core and splitting the excitation and output windings each into two halves, which are then wound on the two outer limbs and connected in series-aiding. The signal flux produced by the two halves of

Fig. 3. *Balanced saturable-transformer switch minimizes switching transients; 1. Control current switching waveform; 2. Break-through into a.c. windings on unbalanced switch; 3. Reduced break-through on balanced switch*



the excitation winding links round the outer core path and cancels in the centre leg, while the control flux divides and produces equal and opposite voltages in the two halves of the excitation and output windings. The excitation flux in practice aids the control flux in one outer limb, while opposing it in the other, and this inherently causes a slight difference voltage in the two halves of the split windings. When using toroidal cores the excitation and output windings are divided in a similar manner, one half wound on one core, and one half on the other. The cores are then placed together before the control winding is added to link both cores. The halves of the split windings are so arranged that the direction of induced flux in the two cores is different. As in the case of the three-limbed core construction, decoupling between control and other windings is thereby achieved.

Applications

Before investigating the choice of a suitable core material and the design of a saturable-transformer switch, the likely application of such a device in the design of systems is considered. As a both-way switch for the selection of recording heads, the saturable transformer is attractive, especially if it can be arranged so that a relatively small amount of control apparatus is required to effect switching.

Whenever a number of switches must be controlled, it is possible usually to arrange them in a matrix, so that control is effected by the coincidence of signals on both ordinates of a uniplanar array. A considerable economy in control circuits is possible using this technique, especially as the number of switches in the matrix group increases. Systems have been described, which use elementary magnetic amplifiers as switches for decoding signals and provide a means of controlling read-out information to a visual display[5,6].

By doubling the bias ampere-turns and winding two control coils on the transformer cores a half-current selection system can be constructed using a matrix, in which only $2n$ control circuits are required to control n^2 switches. An arrangement is shown in Fig. 4, where 100 drum-heads share common read and write amplifiers, one at a time. Two additional saturable-transformer switches, controlled from a bistable trigger, change over the common lead from one amplifier to the other. In this example, all the 100 output windings are connected in series and the 99 windings of the unswitched saturable-transformer output windings present an impedance that is several times that of the load. However, if the selection is done in groups of 10 switches and each output from the 10 groups is combined in another group of 10, then for the addition of another 10 switches, the attenuation of a wanted signal is reduced to 6 dB. The array of switches then resembles the familiar relay contact tree as shown in Fig. 5. A group switch is wired in series with the output windings of each group of 10 drum-head switches and is opened by a Y-ordinate control signal. The switch in whose control windings both the X- and Y-ordinate control currents coincide is the one which closes and connects its associated drum head to the common lead. Attenuation through the network is minimized by designing each switching stage approaching the common lead, for a progressively higher impedance match. Unwanted signals from all groups, except that in which the selected head is located, are suppressed considerably more by having two open switches in tandem in their paths to the common lead. A further reduction in the number of control elements is possible if a system of coding is adopted for the matrix selection leads, so that the presence and absence of signals coincident on a number of leads is used to define every one of the selection states which can be expressed using the code. For example, a 15×15 switch matrix is controlled by 8 leads only in groups of 4 on both ordinates, using binary coding and the additive-current selection principle.

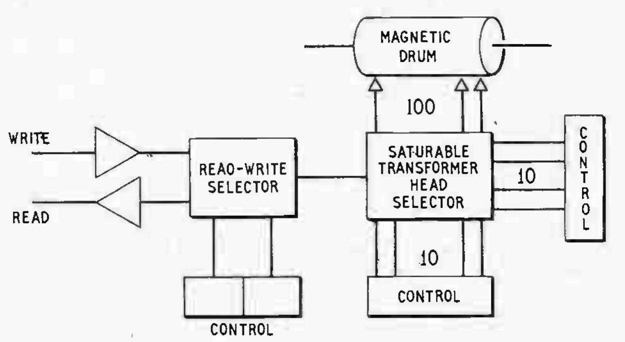
Error checking is a valuable feature of an information processing system and often results in overall economies, because less reliable apparatus and signals of inferior definition may be used, provided that errors can be detected when they occur, and that the detection apparatus itself is unlikely to introduce errors. The saturable transformer is inherently reliable and can be used as the error-detection device. In systems dealing with numbers in the radix of ten, a 'two markings out of five' code identifies each one of ten by unique code combinations. If an error-detecting saturable transformer has five control windings, one on each control lead, and the common output lead from a one-out-of-ten selector is taken via this transformer, it can be arranged that only when current is present in two windings is the switch closed. Current in more or less than 2 coils maintains the cores in one or other of the two possible saturated states and inhibits transmission. The function of checking the X- and Y-ordinate control signals may be included by design into the action of the group selector switch. Only 10 control circuits are necessary for this arrangement.

A model translator for telephone exchange codes has been constructed and exhibited using saturable-transformer switches and the techniques described[9]. It has been functioning satisfactorily for a number of years and the selection system is quite reliable.

Core Materials

The digit rate of typical magnetic-drum systems is 50 to 100 kc/s, and it was thought that this consideration would largely govern the choice of core material. Ferrite cores were readily available, having high

Fig. 4. *Application of magnetic switches to magnetic-drum system. One drum head is selected for connection to a common read or write amplifier*



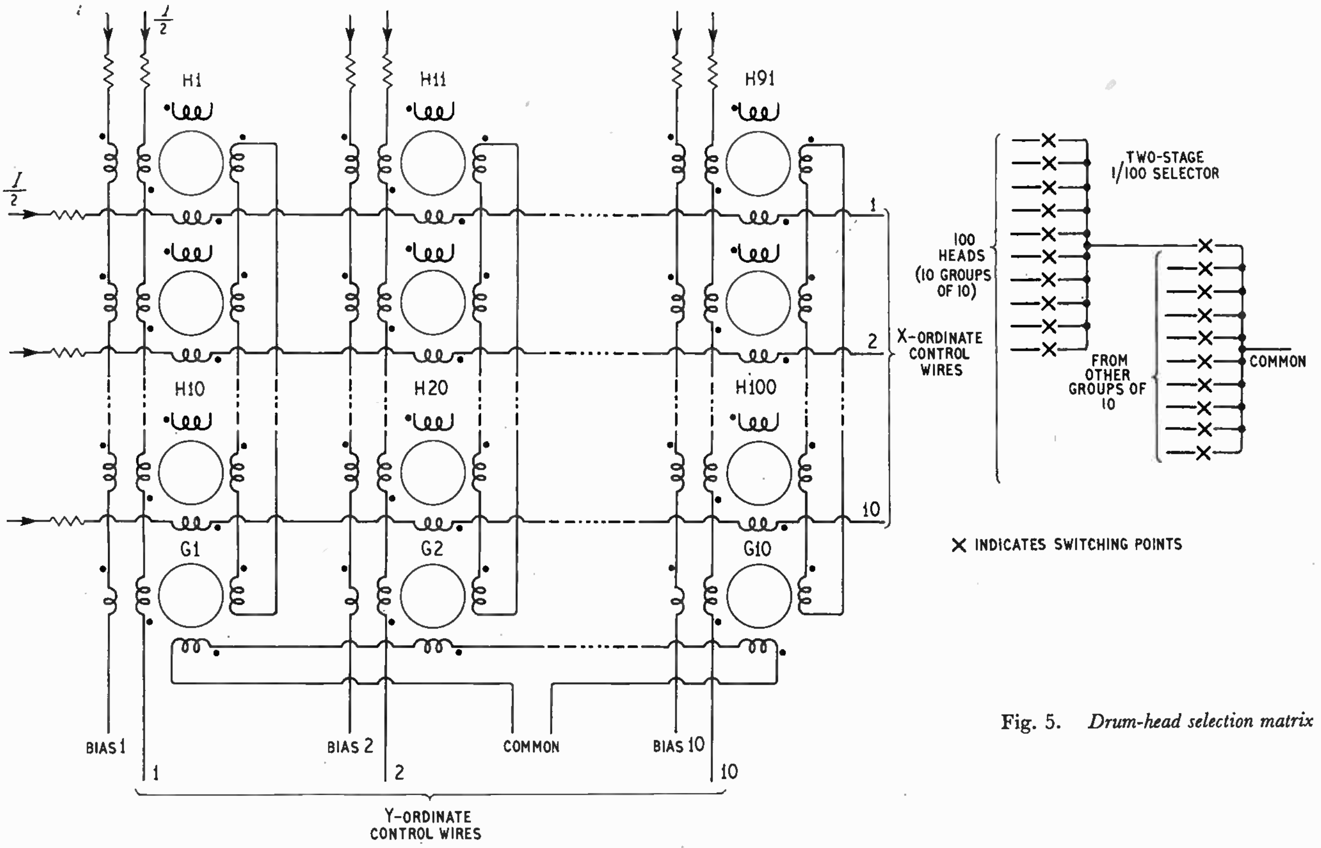


Fig. 5. *Drum-head selection matrix*

resistivity and giving useful permeabilities at these frequencies. However, it was found difficult to saturate them adequately, and their advantages were very much off-set by the problem of providing the very large magnetizing force needed to reduce the permeability of the material and give desired switching ratios.

Typical ferrite-core material has an initial permeability of 1,200, which falls to 200 at a magnetizing force of 2 ampere-turns/cm and is reduced only to about 120 for a magnetizing force of 10 ampere-turns/cm. A further reduction in permeability moreover, necessitates a much greater increase in ampere-turns and the requirements are somewhat excessive. Their use has been described, however, in a switch system, which imposes rather severe restrictions on the number of switches which can be grouped together[7].

An alternative is to use nickel-iron alloys which are available in the form of thin tape[8]. This tape has very high permeability and can be wound into toroids. The effective resistance of the material, although not as high as for ferrite, can be improved by coating the tape with a powder which insulates each turn from the next without reducing the permeability significantly. Cores wound with tape of 0·001-in. thickness or less exhibit properties of high permeability, low eddy-current loss, and a tendency to saturate at low magnetizing fields, due to the orientation of grain along the strip. They are thus superior to ferrite materials or laminated cores. A typical nickel-iron alloy tape core has an initial permeability of 30,000, which falls to 300 for a magnetizing force of only 2 ampere-turns/cm. A switching ratio of 100, therefore, places a relatively modest demand for current on the control bearing in mind that, in order to minimize the switching time of a core, the control turns must be kept as small as possible. The permeabilities quoted are for direct-current magnetization only. When an appreciable alternating current is superimposed, the ratio of permeabilities is reduced as the frequency increases.

Design Considerations

Experimental saturable transformers have been designed and constructed using two types of tape-wound toroidal cores and these switches formed the basis of investigation work on a drum-head selector. Full design and constructional information of these two switches is to be found in the Appendix. The drum heads are of low impedance and transistor read and write amplifiers also have relatively low input and output impedances. All wiring from the heads to common amplifiers through the switching network is therefore made with unscreened conductors, and the saturable transformers are designed to match these impedances.

In order that the same switch may be used for both read and write operations, it must be capable of handling a wide range of excitation voltages. A typical output from the low-impedance head is an almost sinusoidal waveform with an amplitude of 1 mV peak and a fundamental component of 50 kc/s, this being the digit rate. The writing signal, however, is squarer in shape and contains appreciable components up to 500 kc/s. A write signal of some 2 V peak must be applied to the head in order to establish sufficient current to record on the drum surface. In practice, about 10 ampere-turns gives a good writing flux in a head, of the type employed. The saturable transformer must therefore be

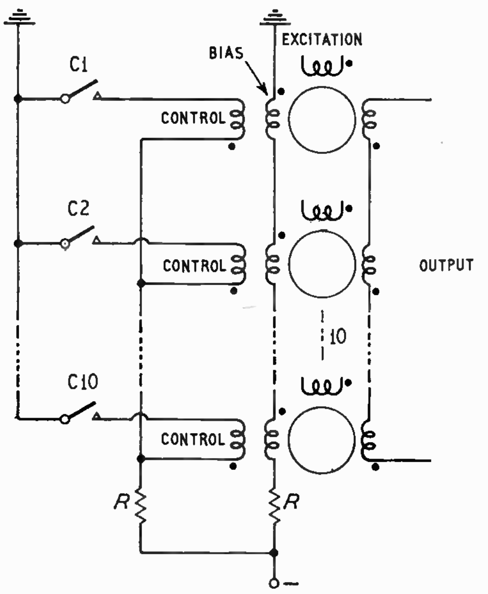


Fig. 6. *One-out-of-ten selector*

capable of handling a wide range of excitation voltages. A satisfactory design covering this range is possible with core materials which have high permeability extending throughout the useful *B-H* characteristic up to saturation.

A number of saturable transformers may be connected together as shown in Fig. 6 to form an N : 1 switch, N being 10 in this example. By this means, any one of ten heads can be connected to a common output lead. The arrangement is reversible so that the switch can act as a 1 : N device. A drum head is connected to each excitation winding on the ten switches and the output windings of all switches are wired in series. A group of ten switches was considered to be of about optimum size. Larger switching groups can be formed by connecting basic groups of ten switches in series-parallel combinations.

Contacts C1–C10 switch current through associated control windings of magnitude defined by a close-tolerance resistor R. Each switch has another winding called a bias winding. These windings are connected in series with a close-tolerance resistor R to the common supply. Bias and control windings are conveniently wound with equal numbers of turns so that R has the same resistance in both circuits. When all contacts C1–C10 are open, every magnetic switch is maintained open by current in the bias windings holding the transformer cores saturated. Any contact which closes switches current through a control winding and the ampere-turns cancel exactly those produced by the bias current. The saturable transformer is switched to its closed state, and connects the drum head to the common path. Without a bias winding, current would have to flow in every control winding to maintain the switches in their open state. This involves using more current and may be less reliable in operation. It is also possible to switch cores more rapidly when using the bias-current technique as explained later.

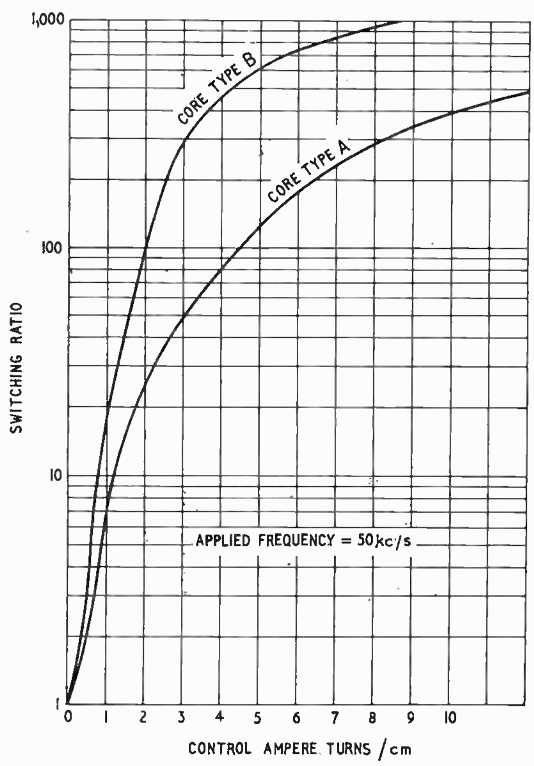
The degree to which cores must be saturated is determined by the proportion of output signal from the closed switch which may be dropped across the combined impedance of the nine other windings in series with the load and the degree of attenuation required of disturbing voltages from other heads at the common output. Provided an acceptable signal-to-noise ratio can be achieved, any loss in signal level can be restored by the amplifier. A switching ratio of 100 has been found more than adequate for this discrimination and gives about 3-dB attenuation of the wanted signal through the switch network. Fig. 7 shows the variation of switching ratio with control ampere-turns for two types of core. From these curves the required amount of current was determined to provide the desired switching ratio.

Measurements

A circuit was devised for obtaining the optimum ampere-turns for the control of cores of two types of material (A and B). Using this value of control ampere-turns, circuits were then investigated in order to achieve fast switching times using junction transistors in the control. A group of ten switches was constructed and measurements made of the insertion loss of the switch network and the attenuation of unwanted signals.

The cores were obtained from two suppliers and consisted of nickel-alloy tape, wound like a clock spring. Sample A, which was constructed from 1 mil tape, has a mean path length of 5·23 cm and a cross-sectional area of 0·226 square cm. Sample B, which was constructed from 2 mil tape, has a mean path length of 4·5 cm and a

Fig. 7. *Variation of switching ratio with control ampere-turns/cm*



cross-sectional area of 0·04 square cm. Both types of core were protected by polystyrene cases.

A 50-kc/s sinusoidal waveform, supplied from a constant-current source, was applied to the excitation winding of a saturable transformer. A 2-ohm resistor was placed in shunt across this winding. This generator network closely simulates a drum head, having an impedance of 2 ohms at 50 kc/s. The control ampere-turns were adjusted until the voltage from the correctly-terminated output winding was a maximum; this indicates complete cancellation of the opposing fluxes. The peak output voltage was set at 10 mV. Control current was then reduced in increments and the amplitude of the output voltage plotted against the out-of-balance current. The ratio of output voltage to 10 mV is referred to as the switching ratio. For core types A and B, a switching ratio of 100 requires 4·4 and 2·0 ampere-turns/cm of core path length respectively and ratios of 1,000 are possible with ampere-turns/cm of the order of 10.

These preliminary tests on the saturable transformer using core type B indicated that, although it was very suitable for switching read signals, it was unsuitable for handling the higher level write signals. The core cross-sectional area is insufficient to switch write signals without introducing excessive losses. Because a large number of turns are needed to keep the flux density in the core from reaching saturation during writing pulses, a leakage reactance is produced which gives the transformer a low-pass filter characteristic, with a cut-off frequency which severely attenuates a write pulse. Only core type A was used, therefore, in the read–write saturable-transformer switch in the final circuits.

The next problem to be solved is how to reduce the switching time of a saturable transformer so that a minimum of time is lost between read and write operations on a group of heads sharing amplifiers. Current must be established in the control winding against the back e.m.f. of its inductance. Initially, the inductance is very low and becomes greater as the core is taken out of saturation. The time taken to switch the core from its saturated to unsaturated state is a function of the voltage which causes the flux excursion between these two states. The relationship $V = N\,d\phi/dt$ can be integrated between the limits of flux excursion to give an expression for the switching time. If ϕ_s is the flux change in lines, V the applied voltage, and N the number of turns on the core, then

$$\int_0^T V\,dt = N\phi_s/10^8 \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

Therefore $T = \dfrac{N\phi_s}{V \times 10^8}$ seconds.

Ideally, therefore, the most rapid switching is produced when N is very small and V large. There is a lower limit to N governed by the available current in the control circuit. Likewise, there is an upper limit to V in a practical circuit, especially critical when transistors are used.

An additional requirement is introduced, because the current in the control winding must be well defined, once the core has fully switched, so that it cancels exactly that flux produced by the bias current. An ideal drive is a combination of constant-voltage and constant-current sources. A constant-voltage source should persist until the core has switched, after which it should become a constant-current source. An indication of when a core has fully switched is given by the increase of the inductance of its windings. However, it is not easy to use this as a means of switching the supply-source impedance from a low to high value without introducing more expense in circuit components. A simplified approach to this problem has resulted in a practical compromise.

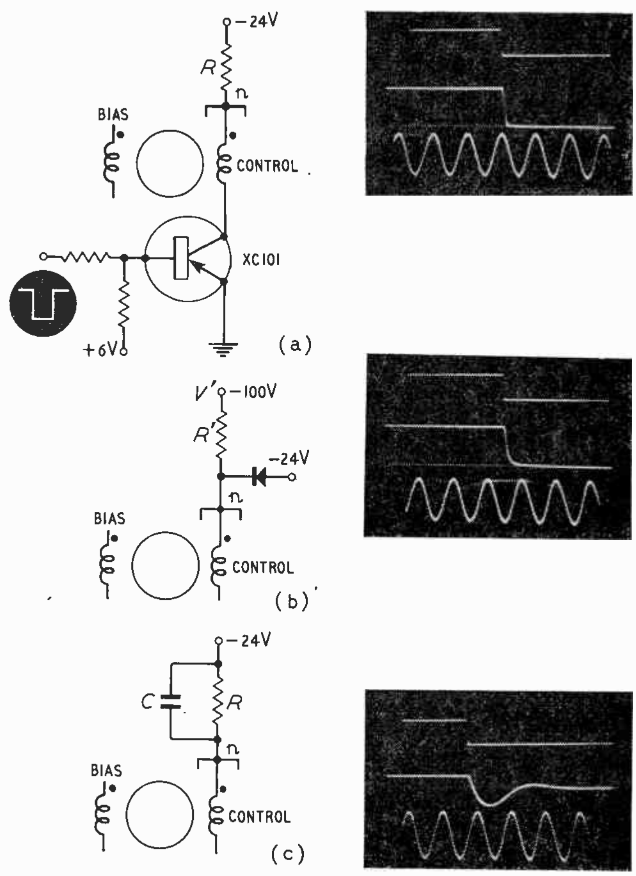


Fig. 8. *Different forms of circuit and the control-current waveforms. These are the middle waveforms in all oscillograms, the upper being the switching and the lower the 40-kc/s timing waveforms. In (a) and (b) I_{max} is 160 mA; in (c) it is 350 mA and the current scale in this oscillogram is one-third of that in the others*

A circuit which comprises a resistor and capacitor in shunt gives an approximation to the desired sequence of voltage and current generators. The charging of the capacitor is determined by the time constant formed by the control winding impedance and the capacitor value. Optimum values of capacitance have been found empirically and no great accuracy has been found necessary. Time constants greater or smaller than nominal increase the time taken in switching the core to a steady state. Any slight overshoot or undershoot causes only a small degree of attenuation during this extended switching time, as the core is substantially in a region of high permeability.

Three types of circuit shown in Fig. 8 were constructed to measure and compare the switching time of cores in the saturable transformers. All circuits use

XC101 transistors to switch current through the control windings, and the value of current was set to provide a switching ratio of 100. Waveforms shown are for saturable transformers using cores type A. A square wave is applied to the base of the transistor, which is connected in common-emitter configuration. The transistor is switched alternately in and out of conduction and connects the battery voltage across the control circuit. The simplest circuit is that of Fig. 8(a). When the transistor conducts, current rises to a value V/R and its rise time is shown in the accompanying oscillogram.

The circuit of Fig. 8(b) derives its control current from a much higher voltage and R' is correspondingly higher to limit the current to 160 mA. A diode clamps the voltage at the control common to — 24 V, when the transistor is switched off. When the transistor conducts, the voltage across the coil remains constant until $I = \dfrac{V' - V}{R}$, after which the potential falls until the applied voltage produces a current of V'/R. In this way, a constant voltage is maintained across the coil until the current has reached about 75% of its final value. Because the current rapidly reaches this value in a winding on the saturated core, little improvement in switching time is gained, as may be seen from the oscillogram. It should be noted, however, that if the core were initially unsaturated and control current were established to bias it into saturation, a much improved switching time would result because, due to the high initial impedance of the winding, a constant voltage would appear across the coil for a longer period, reaching 75% of its final value relatively late in the current switching period.

TABLE I

Saturable Transformer Switching Times

Core Type	Excitation Volts	Switch-on Time (μsec)	Switch-off Time (μsec)	Circuit Figure	Resistor Value (Ω)	Capacitor Value (μF)
A	0·1	250	50	8 (a)	150	—
	0·1	200	45	8 (b)	625	—
	0·1	100	50	8 (c)	150	0·3
	2·0	200	65	8 (a)	150	—
	2·0	150	60	8 (b)	625	—
	2·0	60	65	8 (c)	150	0·3
B	0·1	250	50	8 (a)	400	—
	0·1	200	45	8 (b)	1670	—
	0·1	25	50	8 (c)	400	0·1
	2·0	150	60	8 (a)	400	—
	2·0	130	55	8 (b)	1670	—
	2·0	15	60	8 (c)	400	0·1

Fig. 8(c) shows the circuit adopted. When the transistor is switched, current builds up to greater than the final value as defined by V/R. Until the capacitor has become substantially charged, a large proportion of the voltage appears across the coil and produces this increased current. A higher mean value of voltage is maintained across the winding than in the two previous circuits, hence a more rapid switching time is achieved. An optimum value of capacitance has been determined for both types of saturable transformer, by observing the rise times of a 50-kc/s sine wave at the output, when the transformers are switched. They are compared with the rise times obtained from a circuit without a capacitor across the resistor, and these waveforms, which also show the fall times as the transistors are cut-off, are given in Fig. 9. Tabulated results of the switching times of all three circuits, using both high and low level excitation voltages, are given in Table 1. These times represent the period which elapses in switching the output between 10% and 90% of its final amplitude.

The times taken in restoring cores to their saturated states do not vary greatly from one circuit to another. The rise of flux in each core is controlled by the applied voltage and current-limiting resistor of the bias circuit, which is the same in each case. No appreciable reverse voltage is produced at the collector when the control transistor is switched off, because the bias circuit always loads the transformer adequately and no precautions were found necessary to protect the transistor.

The oscillograms in Fig. 9 clearly illustrate that, whereas the current in all three circuits rises to 160 mA in less than 10 microseconds, the complete switching of the core takes a considerably longer time. Current rise-time was measured as the voltage produced across a 1-ohm non-inductive resistor in series with the control winding. Inserting values into the expression $T = N\phi_s/V$, the theoretical time taken to switch using cores type A under voltage-drive conditions is

$$T = \frac{150 \times 8{,}000 \times 2 \times 0{\cdot}226}{24 \times 10^8} \text{ seconds}$$

$= 226$ microseconds, and for cores type B

$$T = \frac{150 \times 8{,}000 \times 2 \times 0{\cdot}0403}{24 \times 10^8} \text{ seconds} = 40 \text{ micro-}$$

seconds, where $V = 24$ volts, ϕ_s is 8,000 lines/cm² multiplied by the core cross-sectional area. The measured period for switching type A under current-drive conditions is only 200 microseconds. We must conclude therefore that (a) the flux excursion is probably less than 8,000 lines/cm² due to remanence and (b) that the final 10% rise in output voltage takes a long time to be established. The oscillogram Fig. 9D confirms this to be the case. Another point of interest is that when handling a small excitation voltage, a transformer need not be fully switched before a full level output voltage is produced.

When ten saturable transformers are connected as shown in Fig. 6, and only one is taken out of saturation, the optimum load termination for maximum power transfer increases from 25 to 42 ohms. The impedance of the 9 output windings in series with the selected transformer output winding, reduces the voltage into the load by about 3 dB. Switching times remain substantially the same as those measured when only one isolated transformer is switched. No discernable break-through of noise from unswitched saturable transformers connected to the excitation voltage source appears at the common output.

Conclusions

Considerable economies in amplifier and switching control equipment can be achieved by using head-selector circuits of the types described. These possess most of the advantages of a simple relay-contact tree but with greatly increased speed and the avoidance of mechanical contacts. The matching problems encountered when using diode-gate selectors are avoided, because excitation and output windings of a switch can be designed to match correctly the impedances into which they work. The higher speeds of working made possible by tape-wound cores have opened up new fields for the investigation and application of saturable - transformer switches. Core circuits are inherently more robust and stable than many other electronic devices and can be considered, therefore, as more reliable. Although initial cost may be higher, the long and trouble-free life of this component offers considerable economies in subsequent maintenance and replacement. However, in many cases, a head-matching transformer is necessary anyway, so that the cost of extra windings is small. Selectors can be made, not only for working with magnetic-drum heads, but for any application which requires an n to 1 or n to m switch and values of n and m can be made large, without introducing excessive attenuation of a switched signal, by adopting the grouping technique. It is expected that as circuit designers become more aware of the potentialities of this useful component, it will be used increasingly for the solution of switching problems.

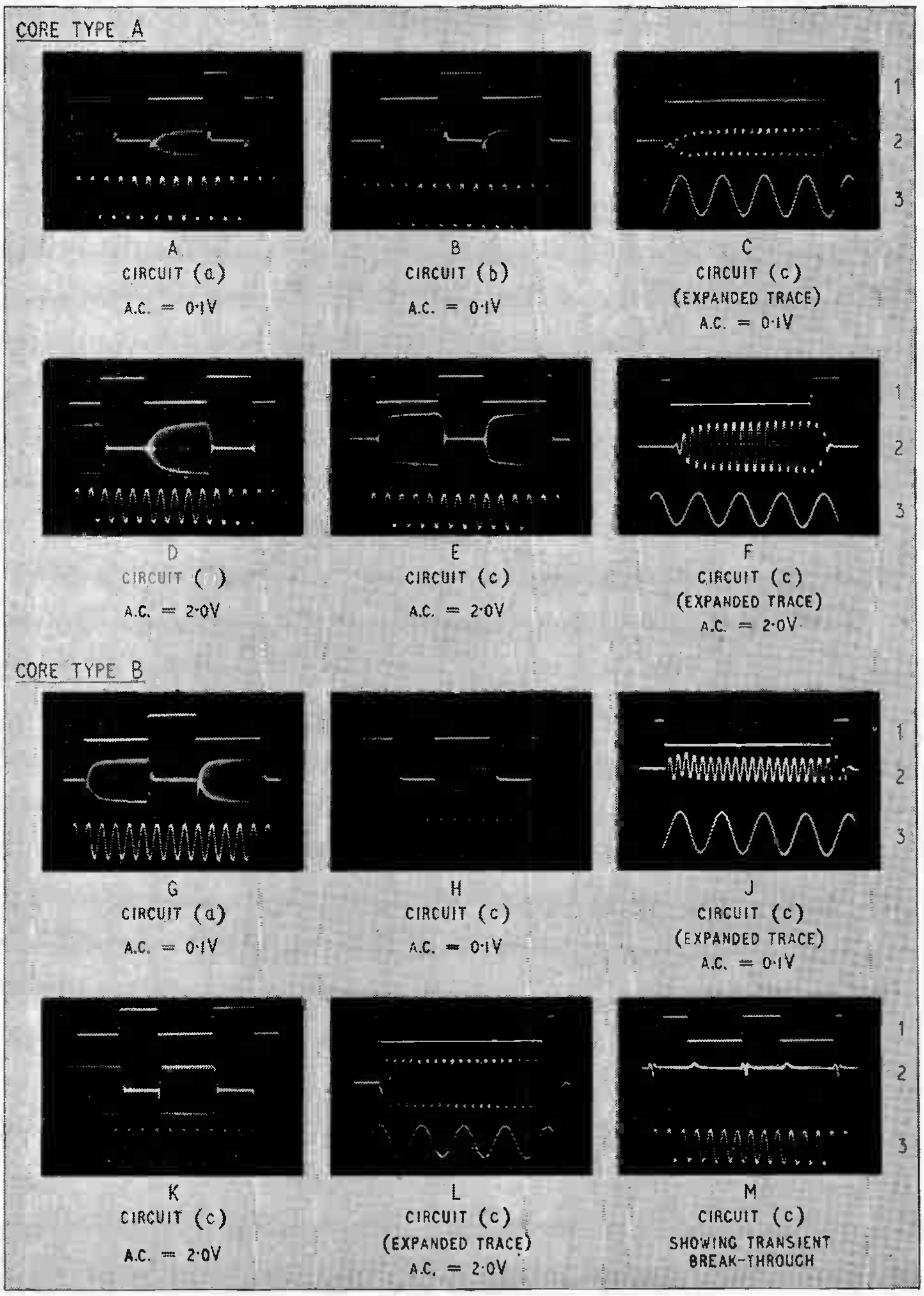


Fig. 9. *Oscillograms of saturable transformer switching times for the circuits of Fig. 8.*
Notes: 1. *Switching waveform;* 2. *Switched waveform* (50 *kc/s*); 3. *Timing waveform* (10 *kc/s*)

Acknowledgements

Acknowledgement is made to the Director of the Siemens Edison Swan Research Laboratory for permission to publish the paper. Thanks are due to Dr. J. E. Flood for his help in the preparation and to W. B. Deller and K. G. Warren for assistance with the experimental work. The cores used were supplied by G. L. Scott and Co. Ltd., and The Telegraph, Construction and Maintenance Co. Ltd.

APPENDIX

Design

The design of saturable transformer was based on the following specification:

Excitation winding to match drum-head impedance, (which is 2 ohms) at 50 kc/s. Output winding to match transistor read amplifier input impedance of 25 ohms.

Transmission loss through a closed switch when correctly terminated to be not greater than 1 dB.

Attenuation of signal by an open switch to be not less than 40 dB.

The shunt inductance of the excitation winding must produce a reactance of 20 ohms at 50 kc/s to give a transmission loss of 1 dB.

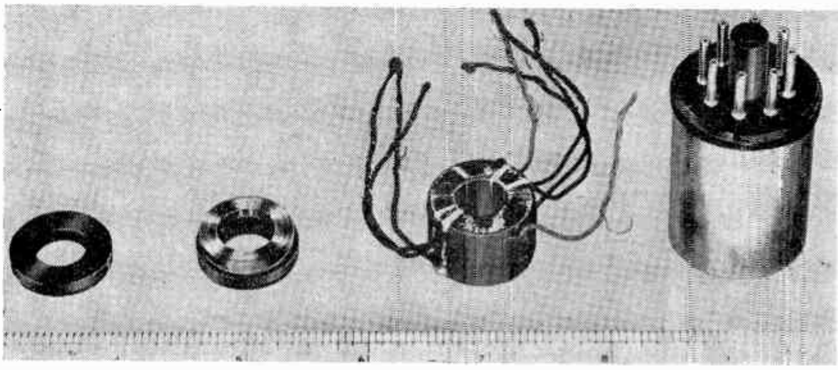


Fig. 10. *Construction of a saturable-transformer switch*

$$\text{Inductance } L = \frac{20}{2\pi f} = 63 \cdot 6 \text{ microhenrys}$$

$$L = \frac{4\pi N^2 \mu A}{l \times 10^9} \text{ henrys} \quad \ldots \quad (2)$$

where l is the mean magnetic path length in cm, N is the number of turns wound on the core, A is the cross-sectional area of the core in square cm and μ is the initial permeability of the core at the frequency of operation.

$$\text{Therefore } N = \left(\frac{L \times l \times 10^9}{4\pi\mu A}\right)^{\frac{1}{2}}$$

Details of core type A. $A = 0 \cdot 226$ cm², $l = 5 \cdot 23$ cm, $\mu = 20{,}000$. Therefore $N = 2 \cdot 5$ turns approximately.

The transformer comprises two cores, each one having no mutual coupling with its partner. Each winding requires an inductance of $L/2$ or $N^{\frac{1}{2}}$ turns, which is about $1 \cdot 75$. The method of joining together the windings is to have a whole number of turns wound on each core and then to connect the two windings in opposition; i.e., the inner of one to the inner of the other. It is necessary, therefore, to choose the next highest whole number, with is 2, and there will be a total of 4 turns on the excitation winding. Transformer ratio $= \left(\frac{25}{2}\right)^{\frac{1}{2}} = 3 \cdot 5$ approximately. Output winding turns per core = 7, giving a total of 14. The number of ampere turns required to give the degree of attenuation to the transformer in its saturated state is now determined by experiment as described in the section on Measurements. It was found that 150 turns on the control and bias windings did not call for currents of greater magnitude than could be handled by an XC101 transistor. With this relatively small number of turns, the inductance of the control winding was $0 \cdot 5$ H, measured at 3 kc/s, with $0 \cdot 1$ V a.c. applied.

In a similar manner, the turns required for core type B were 7 and 25 respectively for excitation and output windings. Control and bias winding turns were 150 each as before. Mean path length was $4 \cdot 5$ cm and the cross-sectional area was $0 \cdot 0403$ cm².

Construction

Each toroidal core was enclosed in a polystyrene box, constructed of two halves fitted together, for the protection of the tape. Excitation and output windings consisted of single-layer coils of 34 s.w.g. enamelled conductor, wound on opposite sides of the cores to minimize mutual coupling when the cores are saturated. Two cores were then placed together and wound with 150 turns of 38 s.w.g. enamelled conductor, which comprised the control winding. Another 150 turns was then wound opposite and became the bias winding. Coils were located, therefore, in the four quadrants of the core as shown in Fig. 10. Any inductive and capacitive coupling between windings is minimized by adopting this arrangement. The completed transformer was then taped and connected to the sockets of an octal valve base, to which was added an aluminium cover for complete protection. The final assembly could be conveniently in the form of a packaged module unit containing a number of switches.

REFERENCES

[1] F. C. Williams, T. Kilburn and G. E. Thomas, "Universal High Speed Digital Computors; A Magnetic Store", *Proc. Instn. elect. Engrs*, October 1951, Vol. 99, Part 2, p. 94.
[2] R. Bird, "The HEC4E Computor", *Proc. Instn elect. Engrs*, Part B, March 1956, Vol. 103, Supplement No. 2.
[3] D. W. Willis, "A Ferrite Core Switch Matrix for Magnetic-Recording—Head Selection", *Proc. Instn elect. Engrs*, Part B, October 1956, Vol. 104, Supplement No. 7, p. 464.
[4] G. M. Ettinger, "Magnetic Amplifiers", (Methuen Monograph).
[5] J. W. Brean, "Magnetic Matrix Switch reads Binary Output", *Electronics*, May 1954, Vol. 27, pp. 157–9.
[6] D. Katz, "A Magnetic Amplifier Switching Matrix", *Trans. Amer. Instn elect. Engrs*, May 1956, pp. 236–41.
[7] B. Moffat, "Saturable Transformers as Gates", *Electronics*, September 1954, Vol. 27, pp. 174–8.
[8] C. E. Richards, E. V. Walker and A. C. Lynch, "An Experimental Study of High Permeability Nickel-Iron Alloys", *Proc. Instn elect. Engrs*, July 1957, Vol. 104, Part B, p. 343.
[9] "Magnetic Drum Translator", Physical Society Handbook of Scientific Instruments and Apparatus, 1957, pp. 83–4.

SEMICONDUCTOR STRAIN GAUGE

Recent research[1] on the properties of semiconductors has shown that certain materials (such as germanium and silicon) possess a number of remarkable properties when used in gauges for the measurement of such quantities as tension, compression, acceleration, pressure, shear force and torque. Results have shown that the piezoresistive effect in these materials is so pronounced that strain gauges can be constructed which are about two orders of magnitude more sensitive than existing types. It was also noted that semiconductors permit the measurement of static and low-frequency stresses, as well as higher-frequency stresses up to the resonant frequency of the material.

The sensitivity of a strain gauge is denoted by its gauge factor (dR/RS) where R is the resistance of the unstrained piezoresistive element and dR is the change of resistance with strain S. If, for instance, the piezoresistive material in the unstrained state has a resistance of 50 Ω but increases in resistance by $0 \cdot 01$ Ω at a strain of $0 \cdot 0001$ unit/unit length, then the gauge factor would be $0 \cdot 01/(50 \times 0 \cdot 0001) = 2$. Now most metal-wire strain gauges have gauge factors between 2 and 4, although values up to 20 are noted for nickel. Using semiconductors, however, gauge factors up to 150 with germanium and 175 with silicon are obtained.

A torsional transducer has been constructed at the Bell Laboratories which, besides having a high sensitivity, responds to both steady-state and alternating torques, and has the added advantage that the angular displacement is negligible. This device employs a germanium cylinder ($0 \cdot 13$ in. diameter $\times$ $0 \cdot 5$ in. long).

As crystals can be prepared and cut so that their unstrained resistance is the same as that of a common wire-type element (i.e., 55 Ω), the same strain-gauge bridge circuit could be used by merely substituting the new semiconductor gauges.

[1] W. P. Mason, "Semiconductors in Strain Gauges", *Bell Laboratories Record*, January 1959.

MAGNETIC REFRIGERATION

"Never-never" had its heyday as a catchword some three generations before its rapturous adoption as a watchword, and its canonization as a device for gracious living in ungracious times. It was, you will remember, the signature tune and gimmick of the Captain of the "Pinafore". What he was never-never going to do, amongst other things, was to utter a mildly naughty word; a situation of such imperishable, ageless, gormless, and generally irresistible dramatic appeal that the reconditioned Mark II model ("My Pretty Pinnace"—a few 10-guinea seats for 1970 still available) will be floated before long. Meanwhile, humbly and in sackcloth and ashes, I crave the use of the expression for myself. I will never, never again inflict on you the Quantum quip, the ponderous wisecrack, the bogus literary allusion. It just isn't safe. For while I am dotting the *i*'s and crossing the *t*'s of some frivolity I fail to mind my *p*'s and *q*'s, and the scientific content slips. So it is never-never from now onwards—or, hardly ever. I shall keep silence, yea, even from good words: but it will be pain and grief to me. I will not be entreated even to finish that limerick that you started; still less will I be bribed; my muse is not for sale. In any case, it would be wicked to clutter up a topic such as magnetic cooling; for this is very simple and straightforward in its general principles and main ideas, and you are safe so long as you don't lose sight of them. I think it could be difficult if you let yourself be distracted by matters of detail, whether relevant or irrelevant; so I shall try for once to avoid both kinds.

The last article dealt with the principles of the gas refrigerator. This month I want to examine briefly the way in which these principles are taken over bodily and used in magnetic-cooling devices. I shall not attempt an exhaustive survey. The article "Magnetic Cooling" by E. Ambler and R. P. Hudson, in Vol. XVIII of *Reports on Progress in Physics* (1956), and Chapter 16 of the 1957 edition of Zemansky, between them cover most of the important experimental work. The theory of the magnetocaloric effect, and the specific heat anomalies associated with it, is given, for example, in L. F. Bates's "Modern Magnetism" (Cambridge University Press).

What is, I feel, more interesting than the details is the general way in which the whole topic ties up with electron paramagnetism and nuclear paramagnetism, and the properties of spin energy levels. The electron-spin levels, operative in microwave spectroscopy and optical fine-structure, give just about the right sort of energy differences to cover the range 1 °K down to 10^{-3} °K. The nuclear-spin levels, associated with radio-frequency spectroscopy and hyperfine structure, cover the range from 10^{-3} to 10^{-6} °K. In each case the temperature range, regarded as a ratio, is large—much greater than the range that ordinary forms of matter usually have to survive. Now, in the appropriate range, the energy-effect of either kind of spin can be isolated; the substance containing the spins can be disregarded, and the system considered solely as an assembly of spins, a sort of 'magnetic plasma' which is virtually a new form of matter. And by the time we have gone so far, we are left with something that is in many ways more like an ideal gas than any actual gaseous substance. Indeed, it will be necessary to distinguish between a 'magnetic' gas and a 'mechanical' gas.

Paramagnetics

Paramagnetic materials in bulk show no external magnetic effect in the absence of an external magnetic field. When placed in such a field they are magnetized in the direction of the field, the induced magnetic-moment-per-unit-mass (which we will here call the specific magnetization σ) being directly proportional to the magnetizing field H, and inversely proportional to the absolute temperature T °K. The constant ratio σ/H is called the susceptibility χ of the material. The statement $\chi = C/T$, where C is a property of the material, is known as Curie's law.

The first notable point about Curie's law is that any substance obeying it can be used for magnetic refrigeration. The second point is, that provided we assume that the paramagnetic contains particles with permanent magnetic moments of their own which are lined up by the field, and provided that the magnetic fields of these particles do not interfere with one another, then Curie's law can be deduced as simply and readily as the p, V, T relation for an ideal mechanical gas. The analogy is very close; for all we assume about the particles of a mechanical gas is that they possess inertia and don't interfere mechanically with one another. Now, a substance does not have to be a mechanical gas in order to behave like an ideal gas; many of the materials to which the ideal-gas equation is applied are in fact dilute solutions. All this business of vapour-pressure and osmotic pressure that you meet with in physical chemistry treats the solute as an ideal gas—and can do so because the particles are so far apart that they do not interact mechanically with one another.

Many of the calculations involving electrolytes, such as finding the e.m.f. of a concentration cell, treat the ions as an electrical ideal gas, in which the ions are so widely spaced that the electrical forces between them are negligible. You can probably see where all this is leading. A substance need not do very much in order to be an ideal gas. We only require of it that it shall have a large number of elementary particles, which can exchange energy with the surroundings by some method —mechanical, electrical, or magnetic—and which do not interact with one another by this method. And so, whether the gross physical state of a material is gaseous, liquid, or solid, it will act as a magnetic ideal gas provided that it contains elementary magnets, and that these are so well spaced that they do not interfere with one another magnetically. It must be added that this status only persists over the range of temperatures at which the elementary magnets are an important vehicle for exchanging energy with the surroundings. The third very useful point about Curie's law is that, when we do get the right sort of paramagnetic, it is obeyed very closely indeed down to extremely low temperatures.

Taking the crudest view, and considering a paramagnetic gas which is also a *mechanical* ideal gas of little magnets, in which the marshalling of the magnetic moments μ of the N particles per gram is in competition with random thermal agitation, the Langevin expression for the susceptibility at T °K is

$$\chi = \frac{N\mu^2}{3kT},$$

where k is Boltzmann's constant. A weak field is assumed, since it is certain that a great enough H could lead to magnetic interaction for any given spacing. The simplest general consideration shows that with a material that followed the above equation, and had no snags or anomalies, it would be just as easy to get very close to absolute zero by magnetic means as it is to approach absolute zero using an ideal mechanical gas in an expansion engine; that is, on paper.

Now consider a material with N paramagnetic ions per gram, for each of which the angular momentum vector is J. This means that there are $(2J+1)$ possible orientations of the magnetic axis of the ion with respect to an external magnetic field. In the 'term' expression for the state of an ion, this information is really given twice, both as superscript and suffix to the $S, P, D, F \ldots$ letter that indicates the value of the resultant orbital momentum, $(2J+1)$ appearing above, and J below, Thus, for the free ferric Fe^{+++} ion, the term is $^6S_{5/2}$, with $J=5/2$ and six possible orientations; for the chromic Cr^{+++} ion, when free, the term is $^4F_{3/2}$, with $J=3/2$ and four possible orientations. But all this doesn't really say what the value of J turns out to be when the ion is buried deep in a crystal lattice, surrounded by the *electric* field gradient due to all the other kinds of ion. The truth is that J can be determined by resonance methods, and all that matters is to know its *experimental* value. In any case, in this article we shall only want to discuss the simplest possible case, $J=\frac{1}{2}$, in order to explain the *kind* of thing that happens.

The 'molecule of magnetism' concerned in the resultant magnetic moment of a paramagnetic ion is the spin magnetic moment of one (or possibly more than one) electron that has escaped exclusion-principle cancellation, but the whole ionic set-up is involved in finding J. If there is no external field, the $(2J+1)$ states are not separated, but form a set of degenerate energy levels. In an external field they acquire their full status as separate energy levels, and are equally spaced from one another. The energy difference between adjacent levels is $g\mu_B H$ ergs, where g is the splitting factor which is very nearly 2, and μ_B is the Bohr magneton, about 9×10^{-21} erg per gauss.

It will be noted that the absolute c.g.s. system of electromagnetic units is being used and, in this particular work, there always seems to be some confusing identification of B and H, and of gauss and oersteds. In the usual derivation of the value of μ_B, the magnetic space constant turns out to be a part of μ_B; it looks, then, as if H above really does mean H, and as if gauss ought to be oersteds. They will be from now on; but I was just deferring to Kaye and Laby here over the units for μ_B. This more detailed picture does not alter the weak-field Langevin expression very much; it simply means that an average value for μ, namely $g\mu_B\sqrt{J(J+1)}$ is needed, giving

$$\chi = \frac{Ng^2\mu_B{}^2J(J+1)}{3kT}$$

Let us now start all over again, with N paramagnetic nuclei per gram, supposing that the only 'molecules of magnetism' are those due to nuclear spin. The effective nuclear magnetic moment is in many cases very close to that of a single nucleon, and its disposition in an external magnetic field is specified by the nuclear angular momentum vector I. The energy difference between adjacent nuclear-spin levels is $g_n\mu_n H$, where g_n is the nuclear splitting factor, and μ_n the nuclear magneton, value $1{\cdot}4 \times 10^{-24}$ erg per gauss, the ratio μ_n/μ_B being the same as the electron/proton mass ratio. The energy difference is about a thousandth of that for electron paramagnetism. Once again, the only difference appearing in the expression for χ is that μ is replaced by an average value $g_n\mu_n\sqrt{\frac{I+1}{I}}$, so that

$$\chi = \frac{Ng_n{}^2\mu_n{}^2(I+1)}{3kTI}.$$

It is relatively easy to obtain magnetic-ideal-gas conditions, using heavily hydrated crystals like ferric ammonium alum $Fe_2(SO_4)_3(NH_4)_2SO_4, 24H_2O$ and chromic methylammonium alum $Cr_2(SO_4)_3(CH_3NH_3)_2SO_4, 24H_2O$, in which the paramagnetic ions are outnumbered by more than 50 to 1 and are three or four times as far away from one another as you would expect ions to be in a simple rock-salt crystal type. Indeed, there is the prospect of being a bit too ideal for the task in hand; in the end, you have to have a real substance to do a real job.

A mechanical gas can interchange internal energy and external work by direct action; you can almost see the molecules battering on the receding piston and rebounding with reduced kinetic energy. A mechanical *ideal* gas couldn't do this, of course, because the piston would have zero collision cross-section for infinitely small projectiles, and they would just shoot through it unimpeded; we must allow an ideal gas to fall from grace sufficiently to make contact with its surroundings. But the magnetic gas starts at a disadvantage anyhow; the ions cannot move about, and the only way in which energy from the spin-levels can be exchanged with the surroundings is by magnetic interaction with the spins of neighbouring ions (spin-spin interaction), or by interaction with the thermal vibrations of the crystal lattice (spin-lattice interaction). The latter is the important stage, corresponding to bombardment of a piston. Weak spin-lattice interaction means that the substance will obey the Curie law closely; it also means that the spin-lattice relaxation time will be long, and that any thermal cycle may have to be a leisurely business. Strong spin-lattice and spin-spin interactions modify the Curie law to the Curie-Weiss law $\chi = C/(T-\theta_c)$, where θ_c is a characteristic temperature, the Curie point; this is, if you like, the analogue of the real-gas Van der Waals' equation—but we don't want anything quite so real just yet.

Thermodynamics and the Ideal Magnetic Gas

For an ideal mechanical gas, the equation involving p, V, and T, for unit mass is

$$\frac{pV}{T} = \frac{R}{M},$$

where M is the molecular weight.

For an ideal magnetic gas, the corresponding equation is

$$\frac{H}{\sigma T} = \frac{1}{C}$$

Before going further, there is a point about temperature to be taken up. We know that the 'T' in the ideal-gas (mechanical) equation is the temperature on the Kelvin scale; and if by any chance we have a gas departing slightly from ideal behaviour we know how to correct for this. If Curie's law is accurately obeyed, then 'T' in the magnetic ideal-gas equation is also the Kelvin-scale temperature; but in the absence of any independent check, this is a matter of assumption. What is usually done, recognizing that there will be slight departures even with the most favourable materials from Curie's law, is to say that T represents a temperature on the Curie scale, or on the magnetic scale; and the difference between this and the Kelvin scale is not enough to matter for the purpose of the present article.

Comparing the two equations above, the corresponding pairs of symbols are p and H, and V and σ. It would look neater if we were able to say that $1/V$ goes with σ; but this would not fit in with the external work calculation.

The external work done by unit mass of gas increasing its volume by dV c.c. at pressure p dynes/sq.cm. is pdV ergs; that done by unit mass of paramagnetic when σ increases by $d\sigma$ units at constant H is $-Hd\sigma$ ergs. All the argument that follows is stock 'ideal gas' algebra worked in magnetic symbols!

In the first-law equation, with the usual symbols for heat dQ supplied, increase in internal energy dU, and external work dW,

$$dQ = dU + dW.$$

Writing TdS (where dS is the entropy change) for dQ, and $-Hd\sigma$ for the external work dW,

$$TdS = dU - Hd\sigma,$$

the whole being expressed in energy units.

So,

$$TdS = dU + \sigma dH - \sigma dH - Hd\sigma = d(U - H\sigma) + \sigma dH,$$

and

$$d(U - H\sigma) = TdS - \sigma dH.$$

The left-hand side of this equation is the change in enthalpy of the system; this must be an exact differential, so that $\partial T/\partial H$ at constant entropy equals $\partial\sigma/\partial S$ at constant H, and $\left(\frac{\partial T}{\partial H}\right)_S = \left(\frac{\partial \sigma}{\partial S}\right)_H$. Writing C_H for the specific heat at constant field-strength, which is $(\partial Q/\partial T)_H$ and therefore $T(\partial S/\partial T)_H$, and cancelling, we get

$$\left(\frac{\partial T}{\partial H}\right)_S = -\frac{T}{C_H}\left(\frac{\partial \sigma}{\partial T}\right)_H.$$

Now, for a substance obeying Curie's law $\partial\sigma/\partial T$ is proportional to $-1/T^2$ and is always negative; hence $\partial T/\partial H$ at constant entropy is always positive. That is, *adiabatic magnetization is always accompanied by a rise in temperature; adiabatic demagnetization leads always to a fall in temperature.*

The general application of this to cooling, by magnetizing a paramagnetic, removing the heat generated by magnetization, isolating the material thermally, and then removing the magnetizing field, has been a standard procedure for the past quarter-century or so. Also, it looks very promising in that the lower the value of T, the greater is the value of $\partial\sigma/\partial T$ and the smaller the value of C_H. With just one great difficulty; every paramagnetic has a certain characteristic temperature at which the value of C_H becomes enormous, and it can only be used successfully for cooling in the range above this temperature, which we will call θ_m.

The Entropy of a Spin System

High-multiplicity values of J for an ion, or of I for a nucleus, affect the actual value of χ, but from the general point of view can simply be thought of as the $J=\frac{1}{2}$ or $I=\frac{1}{2}$ case several times over. In this case, we have a doublet (Fig. 1) with energy separation in a field H units which is H multiplied by the appropriate $g\mu$. One has to be a little careful here (not that it matters much in the end) about the way signs are used, bearing in mind first that the electron is a spinning *negative* charge, and second that the potential energy of a magnetic dipole lined up with a magnetic field is usually written as $-MH$, where M is the moment of the dipole. Whatever the rights and wrongs of this, in the figure the spin quantum-number $m=+\frac{1}{2}$ denotes the higher of the two levels, and $m=-\frac{1}{2}$ the lower. This *is* correct.

Next, how have we changed the entropy and the magnetic potential energy of the system by the application and subsequent increasing of H? The entropy

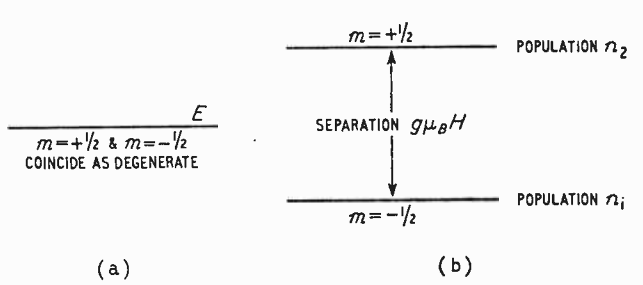


Fig. 1. *Electron-spin levels for the $J = \frac{1}{2}$ doublet, where the only possible values of the spin quantum number m are $+\frac{1}{2}$ and $-\frac{1}{2}$. (a) In the absence of an external field they are degenerate, and equally populated; (b) In an external field H, the levels, separated by $g\mu_B H$, are respectively $\frac{1}{2}g\mu_B H$ above and $\frac{1}{2}g\mu_B H$ below the original common energy E. But as the number n_1 in the lower state exceeds the number n_2 in the higher state, the total energy of the system has been lowered. An exactly similar figure, with the substitution of $g_n\mu_n H$ for $g\mu_B H$, represents the $I = \frac{1}{2}$ nuclear-spin doublet*

has been reduced; so has the magnetic energy. For entropy is the Representation of Chaos and, in aligning spins, one is introducing some ordering into thermal chaos; and the magnetic potential energy of a dipole is of course a maximum when its axis lies at right angles to H; the $-MH$ above really takes this maximum position as an arbitrary zero.

But we can, assuming the usual Boltzmann distribution, see that the energy has in fact been reduced; for, calling n_2 the number of ions in the higher state, and n_1 the number in the lower state, then

$$n_2/n_1 = \exp(-g\mu_B H/kT).$$

More spins populate the lower state than the higher one, so the total energy is lower than it was without the field.

The entropy of a system with given J in a given field H can be calculated, and all this is explained in the review article by Ambler and Hudson, and in the other references. The expression is

$$S = R \log_e (2J+1) + \int_0^H \left(\frac{\partial \sigma}{\partial T}\right)_H . \, dH.$$

As we have seen, $(\partial\sigma/\partial T)_H$ is negative, and this expression fits in, as it must do, with the elementary reasoning that shows that increasing H for a paramagnetic at constant temperature gives a *decrease* in the entropy S. Fig. 2 shows the entropy-temperature-field curves for ferric ammonium alum, with a possible continuous cycle outlined thereon.

The Cyclic Refrigerator

I hope I have got the idea of adiabatic demagnetization on its own in satisfactorily. The next point is the use of this in a magnetic Carnot-cycle as a continuously operated cyclical heat-pump. The difficulty with all these things is the realization of isothermal conditions when they are needed, and of adiabatic conditions in their turn when they are wanted. Fig. 3 shows the scheme of the magnetic refrigerator of Heer, Barnes, and Daunt (1954) for producing and maintaining temperatures below 1 °K. Here, A is a 15-gram specimen of ferric ammonium alum; at the lower end is the 'hot reservoir', liquid helium at 1·1 °K; above the working substance is the 'cold reservoir' R, of chromic potassium alum. Magnet 2, which gives up to 7,000 oersteds, is the main magnetizer. Magnets 1 and 3 serve to open and close communications to the reservoirs by means of the strips of lead V_1 and V_3. Lead at these temperatures is a superconductor in the absence of a magnetic field, which means that its thermal conductivity is very high indeed. Switching on the fields 1 and 3 at the appropriate stages in the cycle destroys the superconductivity and effectively insulates the reservoirs for the adiabatic strokes of the cycle. Referring back to Fig. 2, let us trace out the cycle ABCD. Fig. 2 is a 'tephigram', that is, a graph of entropy against temperature; the usual shapely Carnot-cycle kind of thing appears on this as a rectangle. It must do, for every isothermal step is a vertical line, and every adiabatic step a horizontal one. The pattern of equal-H lines is there just to help us to follow what is happening, and also to emphasize that the entropy is reduced by increasing H, which I am sure none of you really believes.

In the figure, AB and CD are the isothermals, and BC and DA the adiabatics. Along AB the salt is magnetized isothermally at about 1 °K, in contact with the liquid helium bath, to which it can reject heat. It is then isolated by manipulating switches V_1 and V_3, and demagnetized to the point C. Heat is then taken in from the cold reservoir at 0·2 °K along CD, the magnetic field being reduced during this process so as to keep the temperature constant. Finally, the working substance is magnetized adiabatically along DA until the temperature reaches 1 °K, when thermal contact with the helium bath is opened, and the magnetization is completed isothermally. The shaded area, CDEF, represents the heat extracted from the cold reservoir during one cycle. This arrangement does not produce any spectacularly low temperature; its interest lies in the fact that it is a heat-pump that reproduces completely in the range below 1 °K the physical behaviour of the gas-cycle refrigerator, and does so with a fidelity to the idealized Carnot-cycle that no mechanical gas could ever achieve!

A fascinating extension of the method has been used for anisotropic paramagnetic salts, which have different susceptibilities in different directions. Magnetized at 1 °K along an axis of high susceptibility, the specimen is demagnetized adiabatically simply by rotating it until an axis of low susceptibility lies parallel to H.

Orders of Magnitude

The entropy-reduction process of magnetization can be regarded as a tug-of-war between the organization of magnetic potential energy '$g\mu H$' and the chaos of random thermal energy represented by kT. As k is $1{\cdot}38 \times 10^{-16}$ erg/degree, μ_B is 9×10^{-21}, and μ_n is $1{\cdot}4 \times 10^{-24}$ unit, we are in a position to compare the two contestants for any given H, say 10^4 oersteds. In such a case, since g is about 2, $g\mu_B H$ is about 2×10^{-16} erg; this is the value of kT at about 1·5 °K, and far outweighs kT at temperatures in the 1 degree to 10^{-3} degree range. The only way of making any significant changes of internal energy or entropy is by the opera-

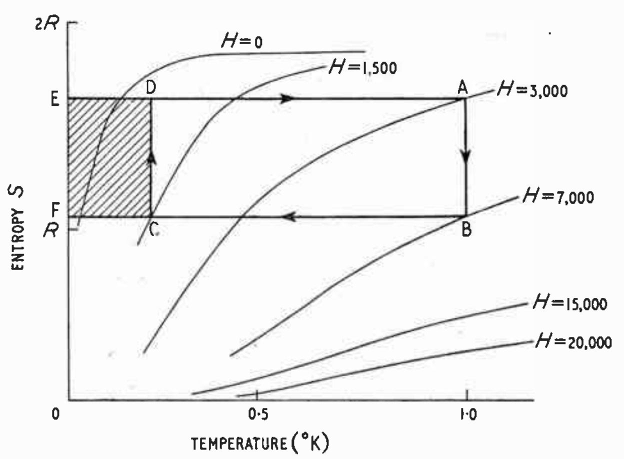


Fig. 2. *Temperature-entropy curves for a paramagnetic plotted for different values of the external magnetic field H oersted. Entropy is plotted in units of R, and temperature in degrees K. The rectangle ABCD, comprising two isothermal stages, AB at 1·0 °K and CD at 0·2 °K, and two adiabatics, BC demagnetizing and DA magnetizing, is the magnetic counterpart of the Carnot cycle. It is achieved in the magnetic heat-pump shown in Fig. 3. The working substance is iron ammonium alum*

tion of a change in H on the electron spins. It is not just an elegant and rather fancy technique; no other method is available, or even conceivable, for getting at 'magnetic matter'.

Two other points should be dealt with here. The first is the tie-up with spectroscopy. Putting in the values of k, Planck's constant h, and the velocity of light c into the expression $hc\lambda^{-1} = kT = g\mu_B H$, you can see that a temperature of 1·5 °K is equivalent to a wave-number of about 1·5 cm^{-1}, which is the level-separation in a field of 10^4 oersteds. This represents a microwave-frequency quantum; your elders who still imagine that spectroscopy means looking at spectra may perhaps prefer to think of it as an optical fine-structure.

The second is this business of the characteristic temperature θ_m and the specific heat anomaly. This was worked out many years ago by Schottky, and all the books can quote you the formulae for the variation of C_H with temperature. J. K. Roberts and L. F. Bates between them supply the simple *physical* picture that we really want. Roberts, citing the distribution function that takes the place of the Maxwell distribution in this sort of calculation, says that we get the anomalously large specific heat because the heat energy has to be distributed among $(2J+1)$ degrees of freedom; Bates shows that the energy of interaction between the ions, U_i, is related to the characteristic temperature θ_m by the expression $U_i = k\theta_m$. It seems, then, that the specific-heat maximum must be a sort of resonance effect, occurring at the wave-number corresponding to θ_m; a matter of microwave spectroscopy, but with the crystal lattice itself furnishing the incident radiation that is usually piped in through the plumbing. As I mentioned earlier, this is a very large effect indeed; the authorities differ by the odd ton more or less, but all agree that, at its θ_m, one gram of paramagnetic has a thermal capacity exceeding that of several tons of copper or lead at the same temperature.

So far these have been electron-spin calculations. Putting in the value $g_n = 1$, and μ_n for μ, the $g_n\mu_n H$ figure for nuclear-spin level-splitting is about 10^{-3} of that for electron-spin splitting in the same field, and the corresponding T about 10^{-3} °K. Quite clearly, with any electron-paramagnetic atom, the nuclear-spin effect will be swamped by the electron-spin effect at temperatures like 1 °K. But once you get down to the appropriate region below 10^{-3} °K, nuclear-spin energy changes become large compared with kT, and are also the only kind of energy change that can be effected. Nuclear-spin-lattice interactions are so weak, and the corresponding relaxation times so long, that temperature reduction by nuclear demagnetization takes some time to observe. Just as θ_m sets a lower limit to cooling by an electron paramagnetic, interaction between nuclear spins gives a comparable anomaly at a temperature θ_n, of the order of 2×10^{-6} °K. It would seem that this is then somewhere near the lowest temperature attainable. An account of experiments using a copper specimen, initially magnetized in a field of $2{\cdot}8 \times 10^4$ oersteds at 0·01 °K, which cooled to about 2×10^{-5} °K on adiabatic demagnetization, is given in an article by N. Kurti, F. N. H. Robinson, Sir Francis Simon, and D. A. Spohr in *Nature* for 1st September 1956.

This seems the appropriate point at which to stop.

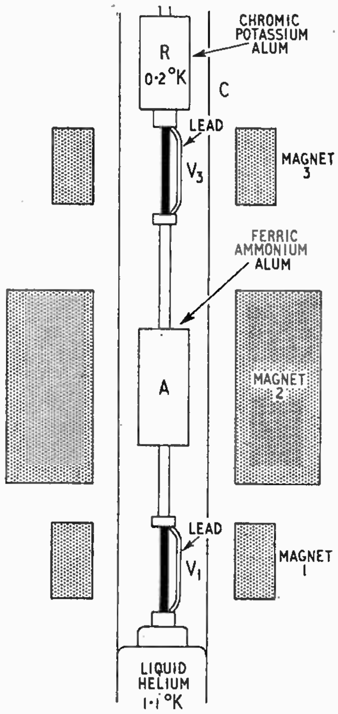


Fig. 3. *Scheme of the magnetic refrigerator, or heat-pump, of Heer, Barnes, and Daunt; R is the cold heat source, at 0·2 °K; the hot heat sink is liquid helium at 1·1 °K. The magnet 2 operates on the working substance A. Switching off magnet 3 puts A into contact with R by allowing V_3 to be superconducting; similarly, switching off magnet 1 puts A into communication with the liquid helium through V_1. When both magnet 1 and magnet 3 are on, then A is thermally isolated from both reservoirs and so can be magnetized or demagnetized adiabatically. The container C is a German silver tube, and the whole of this is surrounded by liquid helium. Appropriate Dewar vessels occupy the space between C and the magnets, but are not shown. Both Fig. 2 and Fig. 3 are adapted from diagrams in the article by Ambler and Hudson referred to in the text*

What we have been doing is to look at a new kind of matter, an ideal gas composed not of material molecules but of spins, a world of magnetic interactions which replace crude mechanical bombardment as a means of energy transferences. At least two interesting fields open for later discussion, namely the properties of materials at very low temperatures, and the place of low-temperature experiments in fundamental research; and now that we have seen how they get there, we may be able to explore this region before long.

RADIO INDUSTRY COUNCIL

The Radio Industry Council has announced that Air Marshal Sir Raymund George Hart, K.B.E., C.B., M.C., has been appointed a director in succession to Vice-Admiral J. W. S. Dorling, who retired in October 1958.

It has also been announced that the Electronic Engineering Association (E.E.A.) will take full responsibility for the capital goods side of the industry. This means that E.E.A. will no longer be represented on the R.I.C., leaving B.R.E.M.A., B.R.V.C.M.A. and R.E.C.M.F. to devote their attention to the affairs of the domestic broadcast industry.

CORRECTIONS

In the article "Hall Effect in Semiconductor Compounds" in the January issue, the following corrections should be made:—

In the author's degrees, "Lond." should be "Hon.".

In Equ. (11), k_n should be k_h.

The last sentence in column 1, p. 6, should read "The reduction of K is due to several causes, . . . "

On p. 7, column 1, in line 11, substitute, "In this case, . . ." for "As a result of this . . .".

In the first sentence on p. 9, column 2, "dB" should be inserted after "−8·5".

In the caption to Fig. 17, the frequency should be 10 kc/s, not 10 Mc/s.

In the article "Lens Aerial Design", on p. 73 of the February issue, the name of one of the authors was unfortunately given incorrectly. It should be P. Foldes.

D

High-Power Transistor D.C. Converters

DESIGNS FOR SILICON AND GERMANIUM TRANSISTORS

By T. R. Pye, B.Sc.*

SUMMARY. *Transistor circuits for converting from d.c. to d.c. (or a.c.) are reviewed. The transformer-coupled push-pull circuit is examined in some detail, and examples are given of designs using both silicon and germanium power transistors.*

Radio and electronic equipment in aircraft, missiles and vehicles must usually operate from low voltage d.c. supplies. To provide the higher voltages required by thermionic valves, rotary converters or vibrator power units have generally been used.

Rotary converters and vibrators have the disadvantage of depending upon moving parts. The vibrator reed may fracture, the contacts may fuse together, and converter bearings and brushes become worn. In addition, sparking at brushes and contacts will occur, particularly at high altitudes, and may cause serious interference.

The transistor converter can replace the rotary converter and vibrator unit in most applications, and will generally show a worthwhile improvement of efficiency. As no moving parts are involved, reliability will be greatly enhanced, and the service life should extend to many tens of thousands of hours.

With transistors at present available, converters can give output powers between about one milliwatt and one hundred watts, at efficiencies of 70–90%. Input voltages may range from about 1½ to 50 volts. Series or parallel connection of the transistors permits still higher output powers and input voltages.

Possible Converter Circuits

To achieve high efficiency in a transistor converter it is necessary to ensure that the transistors are either bottomed (i.e., passing a high current with a low voltage drop) or cut-off (passing only a small leakage current): in each case power dissipation is low. During switching between these states, hole storage may allow conduction with a high applied voltage, and consequent high dissipation; for low average dissipation it is, therefore, necessary for the switching time to be short compared with the conduction time.

These considerations exclude a sinusoidal oscillator, but suggest the generation of a pulse or square waveform; this can be transformed to any desired voltage and rectified if necessary. The ringing-choke and transformer-coupled circuits both use this broad principle, and they will now be compared.

The Ringing-Choke Converter

In this circuit (Fig. 1) a current, I_c, flowing through the bottomed transistor rises linearly in the transformer primary P, storing inductive energy in the core. A constant voltage is induced across the feedback winding F, and a constant current I_b flows into the base through resistor R. When I_c reaches βI_b (where β is the common-emitter current gain) the transistor comes out of bottoming and the rise of I_c can no longer be maintained, the base drive falls, and the transistor is cut-off. The resonant circuit formed by the inductance with its self and other capacitances now 'rings' and a voltage, of a polarity allowing the diode to conduct, is induced in the transformer secondary. While this voltage exceeds the potential to which the capacitor C is charged, current will flow to the capacitor; when it falls below this the circuit continues its ringing and the reverse half-cycle switches the transistor on again. Some of the energy stored in the inductance is thus delivered to the load.

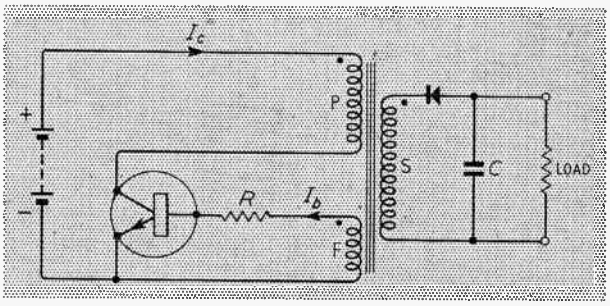


Fig. 1. *Basic circuit of ringing-choke converter*

The power delivered to the load depends on the base drive, and is largely independent of the load. The voltage regulation of the basic converter is therefore very poor, although stabilizing circuits can effect considerable improvement. If the transistor conduction period is short compared with the 'ringing' period (and this is desirable for low ripple) the transistors and diode will experience high inverse voltages, especially under light or no load conditions.

The circuit has however proved useful for low powers (from a few milliwatts to a few watts) particularly with low supply voltages; further details have been published[1].

The Transformer-Coupled Converter

In this circuit, transistors are used to switch the polarity of the supply across the primary of a transformer: a square-wave voltage is induced in the secondary winding and full-wave rectification then gives d.c. with little ripple. The circuit can be self-oscillating;

* *Texas Instruments Limited, Bedford.*

alternatively, the switching transistors may be driven by a square-wave oscillator, perhaps similar to that described below, but supplying little power. In this driven converter the output transformer must handle the full output power without saturation. This arrangement, though complicated, may score for the highest powers (as long as the large transformer is acceptable). Moreover, a single oscillator may drive more than one converter: this scheme may be attractive if wide load variations are expected, as converters may then be switched in or out according to load variations. In the self-oscillating converter, windings on the output transformer drive the switching transistors, and switching occurs when the core saturates. Although inefficient when underloaded, this circuit gives good efficiency at optimum load; regulation is also good. The self-oscillating converter is of most general application for medium and high powers, and to this type the remainder of this article is largely devoted.

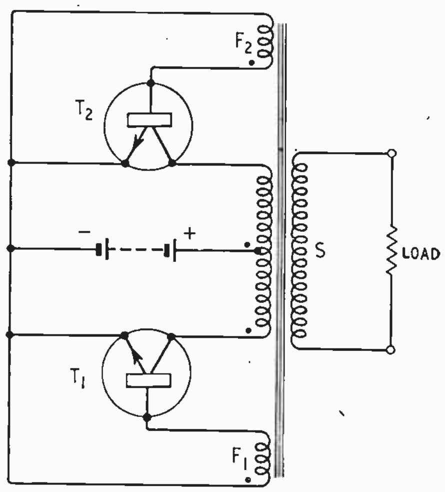
Transformer-Coupled Self-Oscillating Converter

Fig. 2 illustrates the basic circuit, devoid of starting and rectification components. The transistors are seen to be in the common-emitter configuration; although the n-p-n type is shown, the operation of the circuit using p-n-p types is exactly the same, except that all polarities are reversed.

Suppose that transistor T_1 is conducting, and that no load is connected to the transformer secondary. The primary inductance will at first be practically constant, and the primary current will rise linearly according to $dI/dt = V/L$, as shown at point A in Fig. 3. A constant voltage will be induced in F_1 and F_2; the polarity of the voltage at F_1 will be such as to maintain conduction of T_1, and that at F_2, will ensure that T_2 is completely cut-off.

When the core material begins to saturate, the primary inductance will fall and the rise of collector current will accelerate, until it reaches a value of βI_b (where I_b is the constant base current provided by F_1). Then dI/dt will become zero (point B) and the base voltage will fall to zero. The collector current will also begin to fall, although it will be prolonged as a result of hole storage. This reduction of current will reverse the polarity of F_1 and F_2 and initiate the conduction of T_2. The process will continue as an oscillation; Fig. 3 shows waveforms over two complete cycles.

Fig. 2. *Push-pull self-oscillating converter*



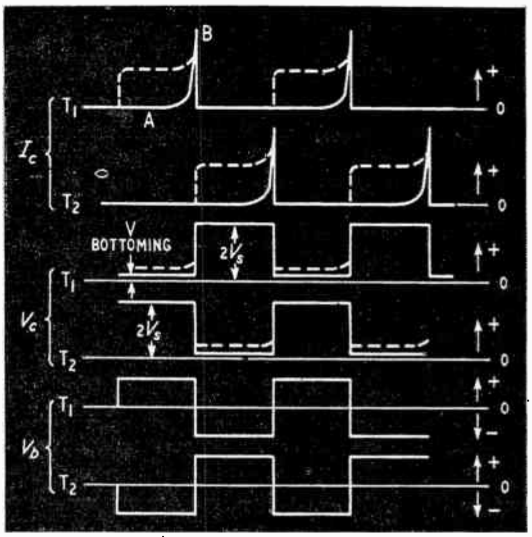


Fig. 3. *Voltage and current waveforms for the two transistors of Fig. 2*

If the secondary winding S has a fixed resistive load, a constant load current will be imposed on the rising inductive current, as shown by the dotted line in Fig. 3. The load imposed on F_1 and F_2 by the transistor base circuits will also result in a small constant collector current.

An expression will now be derived for the oscillation frequency.

If a voltage V is applied to an inductance having n turns, the core flux rise will be given by

$$n\,d\phi = V\,dt \qquad (1)$$

If the saturation flux ϕ_{max} undergoes complete reversal in a time t, we can integrate (1) as follows:

$$\int_0^t \left(\frac{V}{n}\right) dt = \int_{-\phi_{max}}^{+\phi_{max}} d\phi$$

$$\text{Hence } t = \frac{2\,\phi_{max}\,n}{V}$$

If we apply this result to the converter and assume that the two primary windings have equal turns n (this need not necessarily be so—see the section describing the asymmetrical push-pull circuit) and that the transistor voltage drop is neglible, for a supply voltage of V_s the oscillation frequency will be given by:

$$f = \frac{V_s}{4\,\phi_{max}\,n} \qquad (2)$$

Starting Circuits

If the supply voltage is connected to the basic converter of Fig. 2, oscillations are unlikely unless the flow of transistor leakage current is sufficient to give a base current which will ensure a loop gain greater than unity.

If a capacitor (of a few microfarads) is connected momentarily between the collector and emitter of one transistor, oscillations will probably commence. Although this procedure is useful for experimental work, it is hardly suitable for normal use.

A practical alternative is shown in Fig. 4; on connecting the supply a voltage pulse will be applied to the transistor base. However, the converter may not start if the supply voltage is increased gradually from zero instead of being applied suddenly; also, resistor R is in series with the base drive and will result in a loss of efficiency.

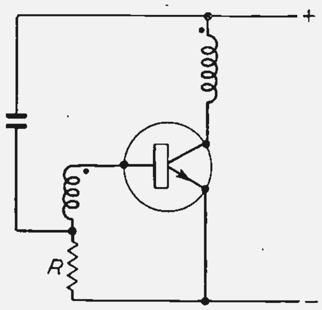


Fig. 4. *One transistor of Fig. 2 with a starting resistor added*

A better method is shown in Fig. 5. The diode D will not conduct initially, but a current will flow via resistor R into the bases of both transistors, causing the transistor with higher β to conduct first.

It will be seen that when the converter has started, base current for the conducting transistor can flow through D; a small voltage drop will of course result.

The transistor current necessary to initiate oscillations may be considerably less than the full load current. The simplified treatment following, due to Stephenson[2], allows a rough assessment of this current (and hence of R in Fig. 5), although other considerations (notably the nature of the load, the transformer leakage inductance and the low-current β of the transistor) make experiment advisable for a final choice.

If r_b and r_e in Fig. 6 represent the transistor internal base and emitter resistances (this of course is only an approximation) then

$$\delta V_b = r_b \delta I_b + r_e \delta I_e$$

$$= (r_b/\beta)\,\delta I_c + r_e\,(\delta I_c + \delta I_b)$$

$$\frac{\delta V_b}{\delta I_c} = \frac{r_b}{\beta} + r_e\left(1 + \frac{1}{\beta}\right) \approx \frac{r_b}{\beta} + r_e \quad \ldots \quad \ldots \quad (3)$$

Now, if the secondary load reflected across the primary is R_L and g_m denotes the change of collector current for a small change of base voltage, then the transistor gain G is given by:

$$G = \frac{\delta V_c}{\delta V_b} = \frac{\delta V_c}{\delta I_c}\frac{\delta I_c}{\delta V_b} = g_m R_L \ldots \quad \ldots \quad \ldots \quad (4)$$

Taking N as the turns ratio of the primary and feedback windings (i.e., $N = \delta V_L/\delta V_b$), the loop gain of the stage will exceed unity (and oscillation will occur) if $g_m R_L/N > 1$ or, from (3), if $R_L/N > r_e + r_b/\beta$ The approximate value of r_e is given by $25/I_e$ where I_e is the emitter current in milliamperes.

Hence, for oscillation $I_e > \dfrac{25}{\dfrac{R_L}{N} - \dfrac{r_b}{\beta}}$ mA .. (5)

This enables the minimum I_e and (from β) I_b per transistor to be calculated; R of Fig. 5 can then be chosen.

Rectification and Smoothing

The converter secondary voltage has a square waveform, and can be used as it is in some applications, such as driving small synchronous or induction motors; rectification will generally be needed, however.

For maximum overall efficiency a bridge rectifier is preferable. Each rectifier must withstand a peak-inverse voltage equal to the d.c. output voltage.

If economy is preferable to efficiency, a conventional full-wave circuit with centre-tapped transformer secondary may be used. Each rectifier must withstand twice the d.c. output voltage; for low output voltages, rectification with only two rectifiers may now be possible. Efficiency will be lower since twice as many secondary turns as in the bridge circuit will be required.

In either arrangement the peak-inverse rating of the diodes should provide an adequate safety margin to cope with any transient voltage peaks which may be present in the secondary waveform.

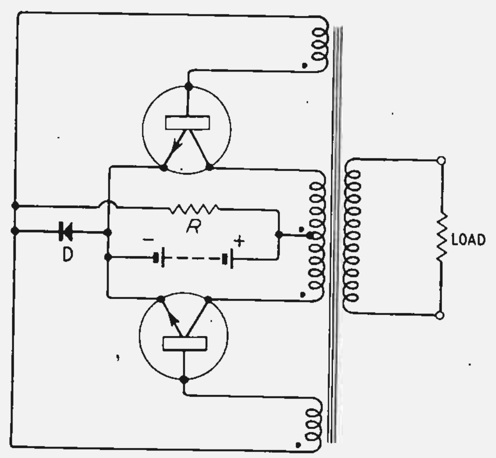


Fig. 5. *Circuit of Fig. 2 with starting resistor R and diode D*

Silicon diodes are compact and efficient, and are available in ratings to cover all normal requirements. Fig. 7 illustrates typical forward characteristics of the Texas Instruments 1S001–5 (750 mA) and 1S111–5 (400 mA) types, both of which are available for peak-inverse voltages up to 600 V and can operate in ambient temperatures up to 150 °C. It will be seen that the forward voltage drop at 25 °C is approximately 0·8 V over the normal working current range.

The rectified, but unsmoothed, output waveform will contain deep 'troughs', representing the rise- and fall-times of the transformer waveform. As the width of the voltage 'troughs' depends on many factors (such as the core saturation characteristics, the transformer leakage inductance and the transistor cut-off frequency), smoothing

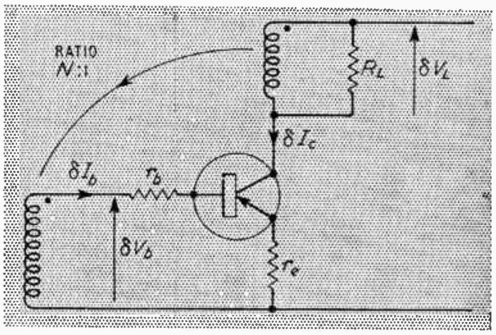


Fig. 6. *Circuit used to describe Equ.* (5)

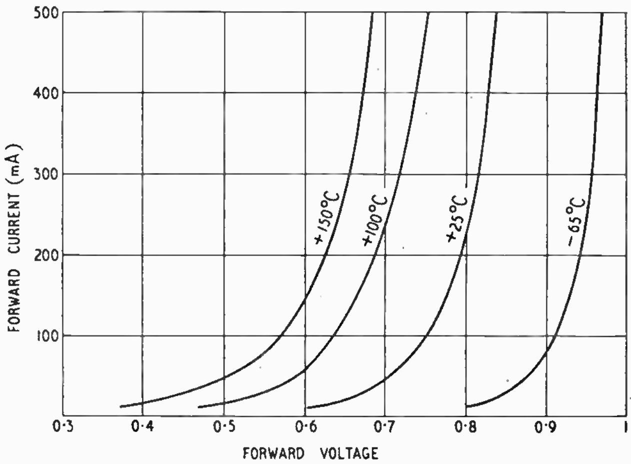


Fig. 7. *Typical static forward characteristics of* 1S001–5 *and* 1S111–5 *silicon diodes*

components are best chosen by experiment. As a very rough guide, a converter operating at 200–400 c/s and delivering a few hundred milliamps might require 1 or 2 μF to give ripple of a few per cent. A smoothing choke will often be unnecessary.

Converter Power Capabilities

For a fixed input voltage, the maximum output of a converter will depend on the permissible peak collector current, the allowable dissipation at the prevailing ambient temperature, and the minimum tolerable efficiency.

Power losses in the circuit, which together will decide the overall efficiency, may be split up as follows:

Transformer Core Losses. Since the core material must traverse the hysteresis loop every cycle, hysteresis losses are independent of load and are decided by the volume of core material, the core hysteresis characteristics, and the operating frequency. Eddy-current losses will be relatively small.

Transformer Copper Losses. These will increase as the square of the current and hence also (approximately) of the output power.

Rectification Losses. Since the rectifier voltage drop is practically constant (see last section) the losses will be proportional to power output.

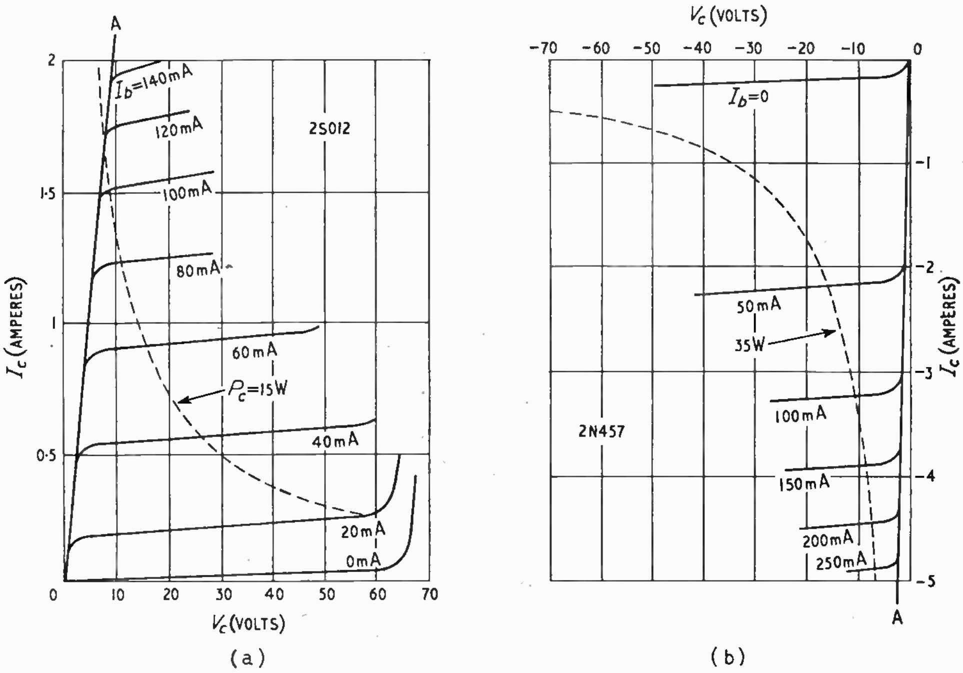
Starting Circuit Losses. R of Fig. 5 will cause a constant loss, so that it should have the highest resistance consistent with reliable starting. D of Fig. 5 will also cause a constant loss, because the base current is independent of load.

Transistor losses. These will include leakage current, transient, forward and input losses.

The open-circuit emitter leakage current, I_{c0}, for a transistor (whether silicon or germanium) doubles for roughly 10 °C rise. At high ambient temperatures the converter transistor which is cut-off may pass appreciable current and, in this condition, will experience a voltage equal to twice the supply; the dissipation may become appreciable. The effect is likely to be small with silicon transistors, as they have a comparatively low initial value of I_{c0}.

During switching, hole storage may allow current to

Fig. 8. *Collector-voltage—collector-current characteristics (a) of* 2S012 *silicon and (b)* 2N457 *germanium transistors*



flow when there is appreciable applied voltage; peak dissipation may then be high. The resulting mean dissipation can be assessed by observing current and voltage waveforms or subtracting other known losses from the total losses.

During conduction, the transistor will 'bottom' at a voltage almost proportional to the collector current, as if it contained a resistor in series with the collector. This resistance, the collector saturation resistance R_{cs}, is represented by the slope of the line A in the typical output characteristics for silicon and germanium power transistors shown in Fig. 8. Forward loss is therefore approximately proportional to collector current (or power output) squared.

The transformer drive winding will provide a constant base current sufficient for the maximum collector current: this will result in a constant input loss.

It will be seen that core losses and input loss cause efficiency to rise with output; the other losses cause it to fall. Efficiency will in fact rise from zero to an optimum value, and then fall away: the maximum efficiency reached and the rate of fall after the optimum

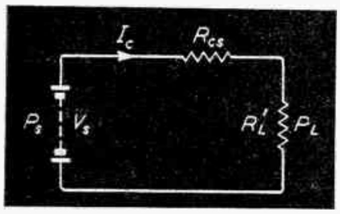


Fig. 9. *Equivalent circuit of transistor in the bottomed condition*

point will largely depend on R_{cs}. This resistance is of particular significance with silicon transistors, and the following analysis considers the converter output and efficiency when other sources of loss are neglected (see Fig. 9). The supply, voltage V_s, passes a current I_c through R_{cs} to the load R_L' (the secondary load R_L reflected across the primary)

Power Supplied $P_s = V_s I_c$

$$\text{Load Power } P_L = P_s - I_c^2 R_{cs} = I_c(V_s - I_c R_{cs}) \quad (6)$$

$$\frac{dP_L}{dI_c} = V_s - 2I_c R_{cs} = 0 \text{ for maximum } P_L$$

Hence for maximum output

$$I_c = V_s/2R_{cs} \quad \ldots \quad (7)$$

Efficiency

$$\eta = \frac{P_L}{P_s} = \frac{P_s - I_c^2 R_{cs}}{P_s} = \left(1 - \frac{I_c R_{cs}}{V_s}\right) 100\% \quad (8)$$

This shows that, for maximum output, $\eta = 50\%$; this result could anyway be anticipated since, for maximum power transfer, the impedance of a generator must equal that of the load.

From (6) and (7) this maximum output

$$P_{Lmax} = \frac{V_s}{2R_{cs}}\left(V_s - \frac{V_s R_{cs}}{2R_{cs}}\right) = \frac{V_s^2}{4R_{cs}} \quad \ldots \quad (9)$$

The current for any output can be found from (6), since $R_{cs} I_c^2 - V_s I_c + P_L = 0$

$$\text{Hence } I_c = \frac{V_s - \sqrt{V_s^2 - 4P_L R_{cs}}}{2R_{cs}} \quad \ldots \quad (10)$$

Only the negative sign for the square root is admissible.

The way in which these equations and considerations are used will be clarified when typical designs are discussed later in this article.

Transistor Requirements

The maximum collector current of a transistor is decided on considerations of instantaneous dissipation and of linearity (β will fall with high collector currents).

Typical voltage breakdown characteristics for silicon and germanium transistors are shown in Fig. 10. It will be seen that current multiplication causes the 'turnover voltage' to be higher with the emitter open-circuit (or with base and emitter connected) than with the base open-circuit.

During the half cycle when a converter transistor is cut-off, it will experience a collector voltage equal to twice that of the supply, and a small base voltage tending to cut it off still further. From this it would appear that a safe supply voltage would be half the open-circuit emitter breakdown voltage.

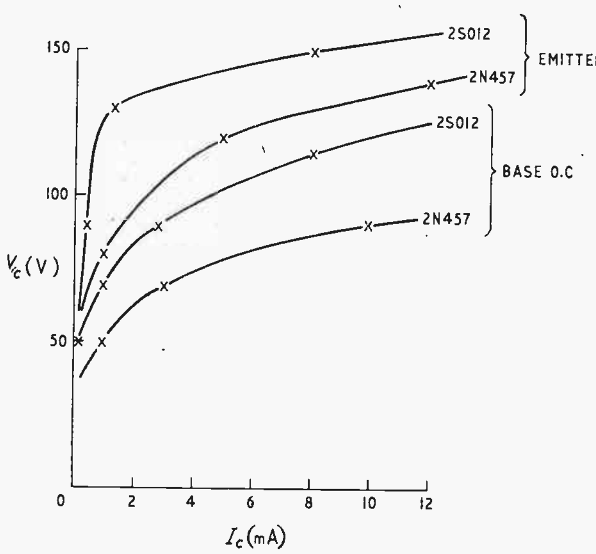
Unfortunately, at the end of the conduction cycle hole storage may allow current to continue when the transistor voltage has risen to the supply voltage or even higher. Under these conditions the transistor open-circuit base turnover voltage is relevant, and should preferably also equal twice the supply voltage if avalanche breakdown is to be avoided.

Transistor dissipation is likely to limit power output only at high ambient temperatures; this is best illustrated by examples:

The 2S012 silicon transistor has a maximum rated dissipation of 15 W at 100 °C. With the maximum value of R_{cs} of 5 ohms and a total supply current of 1 amp, the mean dissipation will be only $2\frac{1}{2}$ W in each transistor; this is well within the rating.

The corresponding figures for the 2N457 germanium transistor are 50 W at 25 °C and 30 W at 60 °C. For a typical R_{cs} of 0·05 ohm, 5 amps will give a mean dissipation of less than one watt.

Fig. 10. *Voltage breakdown characteristics of* 2S012 *and* 2N457 *transistors*



These figures ignore leakage current, transient and input dissipations, as these will normally be small. They show that heat-sink requirements are usually modest. General information on heat sinks has been given[3], although the converter designs which follow suggest suitable sizes.

It is important that the peak collector currents of the two transistors should be practically the same, otherwise the lower current will limit the output prematurely. The fall of β at high currents tends to equalize these peaks; however, transistors with widely differing β should not be used.

Since the feedback winding provides an almost constant voltage, variations of transistor input impedance will affect base (and hence collector) currents. Input characteristics for a selection of silicon and germanium transistors are shown in Fig. 11; the variations, particularly with germanium, are considerable. The addition of a small base series resistor, of the same order as the input impedance, will tend to give constant-current drive. The minimum resistance possible must be used otherwise drive power will be lost and hole-extraction efficiency on switch-off impaired (although this may be restored by a bypassing capacitor). The base resistor can be common to the transistors and in series with the diode.

The precise operation of any converter design should be carefully investigated, particularly with respect to transistor treatment, as the transistor and transformer parameters involved make exact behaviour hard to predict. Transistor voltage and current waveforms can be observed on an oscilloscope, and the application of these waveforms to the oscilloscope X and Y plates gives a useful display of collector dissipation throughout the complete cycle.

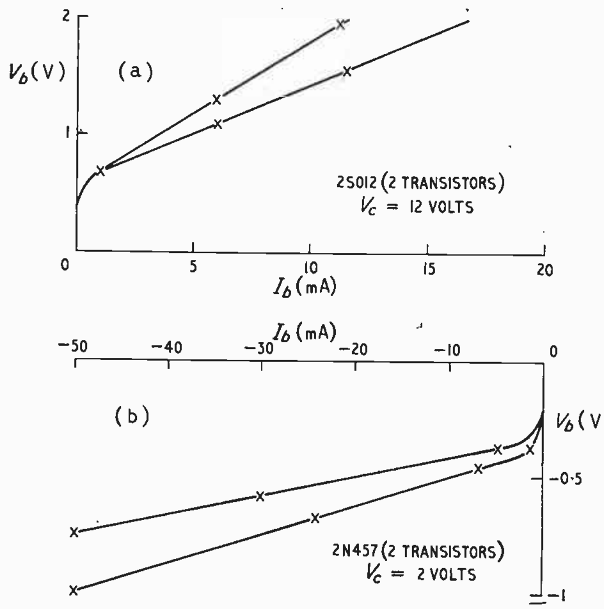


Fig. 11. *Input characteristics of* 2S012 (*a*) *and* 2N457 (*b*) *transistors*

Transformer Design

The transformer core material must have low hysteresis loss when taken to complete saturation and, preferably, a high saturation flux density. A high permeability should be maintained until saturation, in order that the inductive current shall be small compared with the load current. All this suggests the use of a material with a square hysteresis loop.

Ferrites with this characteristic are available for use in magnetic shifting registers and memory matrices. Unfortunately, they show a comparatively high hysteresis loss when taken to complete saturation and, in common with most ferrites, have a low saturating flux density (about 3,000 gauss).

Certain nickel-iron materials, such as H.C.R. alloy (made by Magnetic and Electrical Alloys Ltd.) also show this square loop characteristic, and have a high saturating flux density (15,000 gauss) with a low saturation hysteresis loss (650 ergs per cycle per c.c.).

The choice of core material and of operating frequency (unless this is decided by other considerations, such as ripple frequency) will depend on the power output required. In accordance with normal transformer-design practice, the most efficient transformer will result when core and copper-wire resistive losses are made roughly equal.

At power levels of less than a watt, the maximum nickel-iron core volume for reasonable efficiency will become very small. A ferrite core may then be preferable, and high-frequency operation (at several kilocycles) will counteract the low saturation flux density and give reasonable copper losses.

For higher powers a nickel-iron core is essential, and a comparatively low operating frequency will be necessary. Core hysteresis losses are directly proportional to frequency; eddy-current and residual losses increase according to a square law. With H.C.R. this causes the losses at 5 kc/s to be about fifty times greater than at 400 c/s: the optimum frequency will generally be from 200–500 c/s.

Low transformer leakage inductance is necessary for rapid switching, and this can best be achieved with a toroidal winding on a strip core. This is an expensive form of construction; by careful design, satisfactory results can, however, be achieved with square-loop laminations of a type often used in transductors. A range of such laminations is obtainable in H.C.R. material; the choice of lamination pattern and stack thickness is best decided by a 'cut and try' method, along the following lines:—

A possible lamination stack is chosen for the power to be handled (suggested sizes are given later). An operating frequency is chosen, and with expression (2) this gives the number of primary turns. By allocating roughly half the winding space to the primary, the primary resistance, and that of the secondary reflected into the primary may be estimated: the copper loss can then be compared with the core loss, obtained from the core volume and the operating frequency. A high proportion of copper loss will suggest raising the frequency or increasing the core volume (and vice versa): the effect on the total loss of these alternative measures can then be compared.

The feedback turns are chosen to give the full load collector current, allowing for the starting diode and added base resistance voltage drop.

The secondary winding is chosen to give the required output voltage, bearing in mind that the primary voltage is less than the supply as a result of R_{cs}. Slight readjustment of the secondary turns may be necessary to cover the resistive drop in the winding.

Typical Designs

Silicon Transistor Converter

This converter was designed to give 15 watts output from a 24-volt supply; it uses two Texas 2S012 Silicon Power Transistors and can operate in ambient temperatures up to about 130 °C. The circuit is shown in Fig. 12.

The 2S012 has a maximum collector-to-emitter voltage of 60 V. The collector-to-base breakdown voltage will exceed this, so that operation will be safe with supply voltages up to 30 V.

The maximum R_{cs} for transistors currently available is 5 ohms. Taking the worst possible case, if other losses are ignored, expression (9) gives the maximum output as $V_s^2/4R_{cs} = 28$ watts, for 50% efficiency. For 70% actual efficiency, which might be the minimum tolerable, the maximum output will be at least 15 watts.

As a rough guide, lamination patterns may be chosen from Table 1. For 15 watts, pattern 226 is indicated. A square stack of these laminations will have a volume of 19·8 c.c.

TABLE 1

Power Output, Watts	Pattern
1 — 3	224
3 — 10	225
10 — 30	226
30 — 100	227

The hysteresis loss of H.C.R. to the saturation flux density of 15,000 gauss is stated by the makers as 650 ergs/c.c./cycle.

For a typical operating frequency of 400 c/s, loss = 650 × 400 × 19·8 ergs/second = 0·515 joule/second or 515 mW.

For an output of 15 watts this corresponds to a loss of 3½%, and is therefore acceptable.

Expression (10) gives the collector current for P_L = 15 watts as 0·76 A. If all losses apart from R_{cs} are assessed at 3 W, the actual I_c will be 0·87 A, and the voltage appearing across the primary will be 24 — (0·87 × 5) = 19·6 volts.

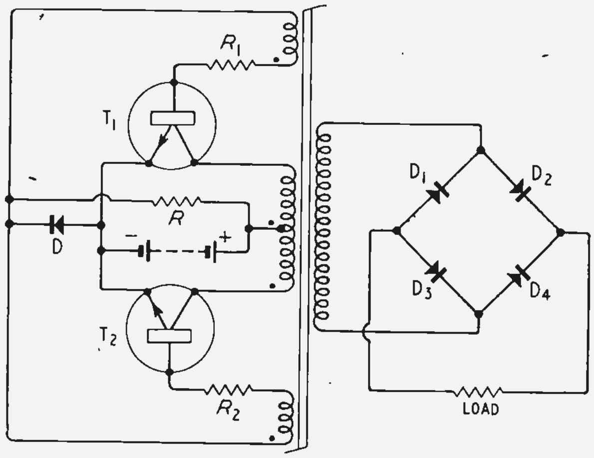
In expression (2) ϕ_{max} must be measured in webers. Since 1 weber = 10^8 lines and 1 gauss = 1 line per sq. cm., the expression can be rearranged as

$$n_p = \frac{V_s}{4B_{max}\,af} \times 10^8$$

where B_{max} is in gauss, a is in sq. cm. For a square stack of 226 laminations, a = 1·61 cm².

$$\text{Hence } n_p = \frac{19{\cdot}6 \times 10^8}{4 \times 15{,}000 \times 1{\cdot}61 \times 400} = 50{\cdot}5 \text{ turns.}$$

Taking into account the thickness of the bobbin, the area of the winding space is approximately 0·4 sq. in. Allowing 0·1 sq. in. for each primary and a suitable space factor, 20 s.w.g. wire is indicated. Two layers



	SILICON CONVERTER	GERMANIUM CONVERTER
T1, T2	2S012	2N457
D	1S111	1S001
R	3·3kΩ	8·2kΩ
R1, R2	10Ω	3·3Ω
D1–4	1S113	1S003

IN THE GERMANIUM CONVERTER, D AND THE SUPPLY VOLTAGE ARE REVERSED. THE TRANSISTORS ARE p-n-p, NOT n-p-n

Fig. 12. *Practical converter circuit with full-wave rectifier. The core symbol is suggestive of a rectangular hysteresis loop and is used to indicate a saturating core*

will give 45 turns, involving a change of frequency to 450 c/s.

For a full load output voltage of 200 (rising to about 240 V on no load), allowing for secondary and rectifier voltage drop, the secondary = 204/19·7 × 45 = 460 turns. This should be of 28 s.w.g. in order to fill the remainder of the bobbin. The copper losses in primary and secondary are roughly the same as the core hysteresis loss and both are small compared with R_{cs} loss: the lamination pattern and operating frequency are therefore suitable.

The feedback turns can be calculated from the V_{be}/I_c characteristic shown in the 2S012 data, and reproduced in Fig. 13: 2·8 volts will be required for an I_c of 0·9 amp. Allowing 1 V drop across the starting diode and 0·7 V across the base resistor, 4·5 V will be required, giving 11 turns for each base winding; 28 s.w.g. is suitable.

To give low leakage inductance it is preferable to sandwich the feedback windings between the primaries, which in turn should be placed in the middle of the secondary, wound in two equal sections.

The input resistance of the 2S012 at high currents is about 30 ohms, and varies little between transistors. Series base resistors of 10 ohms will drop just less than the 0·7 V allowed and give sufficient current equalization throughout the cycle.

The emitter current required for starting may now be calculated from expression (5). Here, R_L = 24/0·9 = 26·6 ohms and N = 45/11 = 4·1. For low currents, r_b will be higher than usual—say 40—and the effective value will be increased by the added base resistors, giving a total of 50 ohms; β will also be lower—say 10. Hence

$$I_e > \frac{25}{\frac{26 \cdot 6}{4 \cdot 1} - \frac{50}{10}} = 19 \text{ mA}$$

The total I_b required will be 3·8 mA, so that $R = 6{\cdot}2$ kΩ. In fact, 3·3 kΩ were found to be necessary: this difference can be explained by the rather arbitrary choice of r_b and β, both of which affect the result considerably, and the fact that the transistors oppose each other until one establishes a higher collector current.

Fig. 14 (a) and (b) show, respectively, the variation of efficiency with load, and the output-voltage regulation curve.

At full load the dissipation is less than 4 watts per transistor. At temperatures up to 100 °C, a copper or aluminium heat sink of about 3 inches square will be sufficient; at higher temperatures a larger one (preferably of copper) may be necessary.

Germanium Transistor Converter

The circuit of this is exactly the same as that of the silicon version; the changed component values are also in Fig. 12. Two Texas 2N457 germanium power transistors are used, and over 100 watts may be obtained from a 24-volt supply at about 90% efficiency.

The R_{cs} of the 2N457 is typically 0·05 ohm, with a maximum of 0·2 ohm. The maximum current of 5 amps will give a maximum bottoming voltage of 1V. In this case, I_{cmax} and not R_{cs} limits the output; the arguments above concerning R_{cs} may be ignored, although an allowance must be made for the bottoming voltage when considering the actual transformer primary voltage.

Table 1 suggests using type 227 laminations and, to reduce hysteresis losses, a rather lower frequency, say 300 c/s, is preferable.

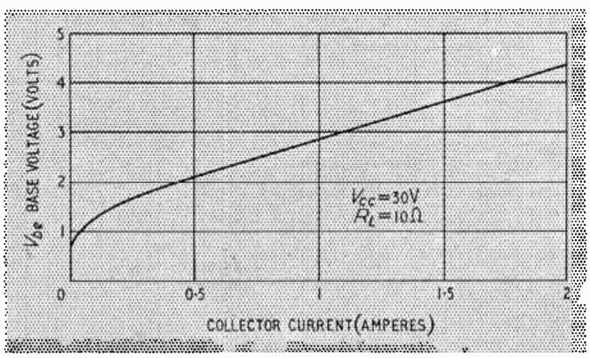
By a similar process to that used for the silicon converter, the windings are calculated as follows:—

Primaries:	Each 54 turns 18 s.w.g.
Secondary:	588 turns 26 s.w.g. (for 250 volts)
Feedback windings:	10 turns each, 26 s.w.g.

The input resistance of the 2N457 is about 10 ohms, and base resistors of 3·3 ohms give adequate current sharing.

The total effective r_b is now 13·3 ohms, $N = 5{\cdot}4$, $R_L = 4{\cdot}8$ ohms and β at least 30. This requires I_e of 55·5 mA, and $R = 6{\cdot}5$ kΩ. The experimental converter actually started with 8·2 kΩ, showing that the β of one transistor was higher than supposed.

Fig. 13. *Base-voltage—collector-current characteristics of* 2S012 *transistor*



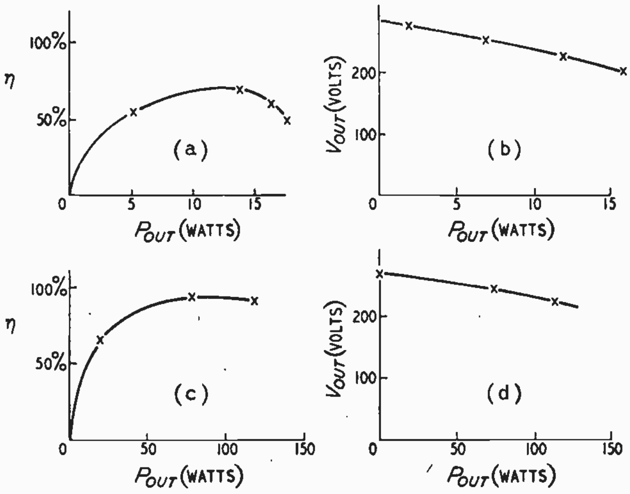


Fig. 14. *The converter efficiency and regulation are shown at* (*a*) *and* (*b*) *for a silicon transistor and at* (*c*) *and* (*d*) *for a germanium type*

The mean R_{cs} dissipation at full load will be less than 2½ watts per transistor, and a small heat sink (say 4 or 5 inches square) will allow operation in ambient temperatures up to about 65 °C (assuming a maximum junction temperature of 80 °C): this is half the temperature permissible with the silicon version.

In neither design is smoothing shown, as ripple specifications may vary widely. As a rough guide, however, a single capacitor of 2 μF gave 4 V peak ripple (2%) on full load with the silicon converter, and necessitated changing the starting resistor to 3·3 kΩ in order to start on full load. For the same percentage ripple, the germanium converter required 4 μF.

Fig. 14 (c) and (d) show efficiency and regulation characteristics for the germanium converter. The first graph shows that considerably higher power and efficiency can be obtained with germanium transistors than with the silicon ones at present available (but, of course, at much lower ambient temperatures).

It is anticipated that silicon transistors will soon be available with R_{cs} as low as 2 ohms; this improvement should be accompanied by an increase of the maximum collector voltage. The silicon transistor will then rival the germanium type for use in converters, particularly with high supply voltages, and even at lower temperatures where germanium could be used. For example, a 2-amp silicon transistor with 100-volt collector rating and R_{cs} of 2 ohms could give at least 80 watts from a 50-volt supply.

Using selected 2S012 transistors having an R_{cs} of 2½ ohms, the circuit of Fig. 5 has given 30 watts at 75% efficiency when operated from a 24-volt supply.

Alternative Circuits

Load in Transistor Emitters

The case of a power transistor is usually connected electrically to the collector. When it is desirable to have the converter transistor cases connected (when,

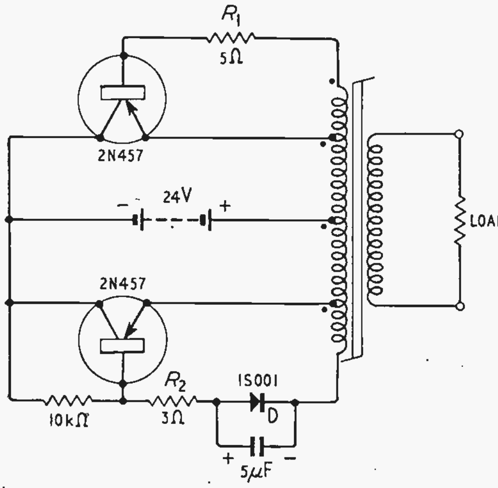


Fig. 15. *Circuit permitting the collectors to be joined together and to a heat sink*

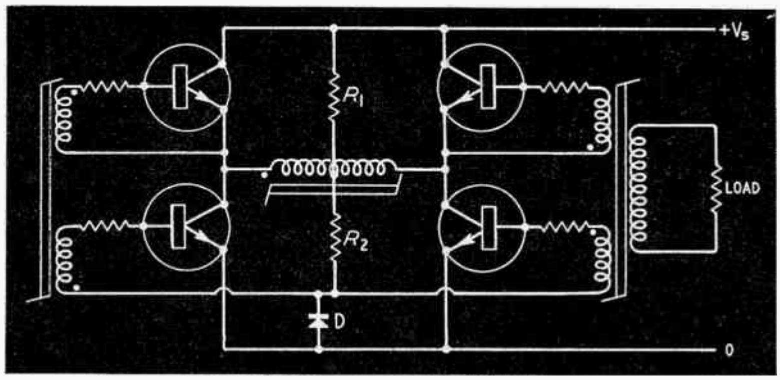


Fig. 16. *Series-connected transistors for high supply voltages; diagonal transistors conduct together. All windings are on the same core*

for example, they are on the same heat sink) the transformer primaries may be placed in the emitter circuits, as shown in Fig. 15. R_1 is larger than R_2, to compensate for the diode voltage drop. C is intended to bypass the high initial forward resistance of D.

Parallel Connection of Transistors

In order to increase the power output of a converter, transistors may be connected in parallel. This will reduce R_{cs} (relevant for silicon transistors) and increase the allowable primary current (relevant for germanium).

Precautions will be necessary to ensure equal current sharing; for example, by providing separate base series resistors, or very low resistances in each emitter. It is desirable also that the values of R_{cs} should be similar.

Series Connection of Transistors

When the supply voltage is too high for the conventional circuit Fig. 16 may be adopted[2]. The senses of the feedback windings cause diagonal transistors to conduct together; the transistors, when cut off, experience only the supply voltage. R_1 serves to bypass the top transistors when starting: R_1 and R_2 together need only provide a collector voltage for the two lower transistors well above their bottoming voltage.

The two lower feedback windings must provide a higher voltage than those above, to offset the diode voltage drop.

A converter using four 2S012 or 2N457 transistors could operate from a 50-volt supply and deliver, respectively, about 30 or 200 watts.

Unsaturated Output Transformer Circuit

Where wide load variations are likely, higher efficiency and kinder treatment of transistors may result from using an output transformer which does not saturate, with a saturating drive transformer. The basic form of this circuit is shown in Fig. 17. The significant effect of this arrangement is that the collector currents do not rise to a peak value, determined by the available drive and regardless of load (as at point B in Fig. 3), but to only a little more than the load currents. This occurs because switching does not happen as a result of the increase of load caused by output-transformer core saturation (as in the normal saturating converter), but because the increase of magnetizing current of T_{r2} at saturation causes an increasing voltage drop across R, until the transistor drive is insufficient for the particular load on the output circuit.

This circuit is described elsewhere[4], together with variations which permit close frequency control with varying input voltage and load.

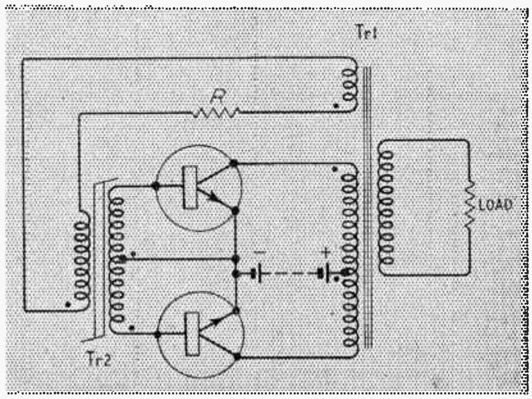
Asymmetrical Push-Pull Circuit

If the numbers of turns on the two primaries in the basic circuit of Fig. 2 are made unequal, one transistor can be made to handle most of the power; the other transistor can be a small-signal type as it need only reset the core.

The rectification will be half-wave, for only the power transistor is loaded, and therefore the ratio of the power-transistor conduction time to that of the reset transistor should be high; the higher this is, however, the greater will be the voltage across the power transistor during the reset period.

The performance of this circuit can approach that of the symmetrical version when supply voltages much less

Fig. 17. *For high efficiency an unsaturating output transformer Tr_1 may be used with a saturating drive transformer Tr_2*



than the power transistor breakdown voltage are used, and it will be more economical.

This arrangement has been analysed elsewhere[5]; the symmetrical circuit is of course a particular version of this general case.

Voltage Multiplication

The square output waveform of the symmetrical converter is particularly adaptable to voltage multiplication.

For economy and good regulation the transformer secondary winding should give the highest possible output voltage, requiring the minimum of voltage multiplication.

The maximum output voltage will depend on the leakage reactance and self-capacitance of the transformer, as these increase the rise and fall times of the output waveform. Insulation problems may also limit the output voltage, in view of the comparatively small size of the transformer.

When this is within the control of the designer, the highest possible supply voltage should be chosen, to reduce the step-up ratio and hence the leakage inductance.

Standard voltage-multiplication circuits may be used; several have been described in a report on silicon-rectifier circuits[6].

A useful technique, to ensure reliable starting of a converter feeding into the capacitative load of a voltage multiplier, is to place a double-anode power zener diode between the transformer and the multiplier; this may have a zener voltage of 20–90 V and, until oscillations provide that voltage, the converter will effectively be unloaded.

A voltage-doubling converter to give about 2 kV has been constructed; the primary side is identical with Fig. 5 and the secondary side feeds a single-ended voltage doubler-circuit. Performance figures are:—

Input voltage: 24 V

No load output voltage: 2·6 kV

Full load output voltage: 1·9 kV for 12·4 watts, with an efficiency of 65%.

It will be seen that power output and efficiency are little inferior to the earlier 250-volt version.

It appears that, with this size of transformer, insulation and wire gauge will limit secondary voltage, perhaps to about 4 kV (peak).

REFERENCES

1 L. H. Light and P. M. Hooker, "Transistor D.C. Convertors", *Proc. Instn elect. Engrs*, November 1955, Vol. 102, Part B, p. 775.

2 W. L. Stephenson, "A Four-Transistor D.C. Converter Circuit", Mullard Technical Communications, December 1958, Vol. 4, No. 36.

3 D. M. Goodway and J. S. Walker, "The Use of Heat Sinks with Silicon Power Transistors", Texas Instruments Limited Applications Report, October 1958, Vol. 1, No. 3.

4 J. L. Jensen, "An Improved Square Wave Oscillator Circuit", *Trans. Inst. Radio Engrs*, C.T-4, September 1957, No. 3, p. 276.

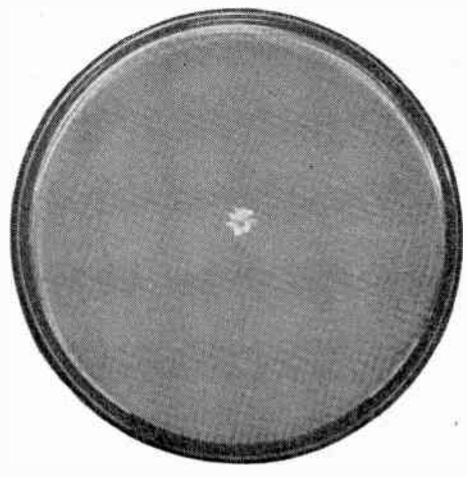
5 R. F. Shea, "Transistor Circuit Engineering", p. 422 (John Wiley, 1957).

6 B. A. Unvala, "D.C. Power Supply Circuits using Silicon Rectifiers", Texas Instruments Limited Applications Report, February 1959, Vol. 1, No. 4.

SCAN MAGNIFICATION

A new system of scan magnification, developed by the Mullard Research Laboratories, was demonstrated recently at the Physical Society Exhibition. Two radar tubes were used, each supplied with identical scanning currents and final anode voltages. The tube in the right photograph was scanned conventionally while the tube in the left photograph employed scan magnification by means of a static magnetic field between the deflection yoke and the screen. The magnification obtained was about twelve times linear.

The outstanding feature of this system, however, is power economy. In a recent article (*J. Telev. Soc.*, November 1958) on transistorized television receivers, it was stated that a power saving of 100 : 1 for the line time-base and approximately 4 : 1 for the field time-base could be achieved by using such a technique.



The fundamental basis of the system is to pass the normally-deflected electron beam through a magnetic field, the intensity of which increases away from the centre of the tube. This field produces an additional deflection which increases as the beam moves off centre, because the more the beam is initially deflected the stronger the additional deflecting field into which it moves. Considerable difficulties arise in obtaining equal magnification in the two directions and also in obtaining good focusing.

The practical arrangements are thus a good deal more complicated than this simple explanation would suggest.

Downcoming Radio Waves

MEASUREMENT OF CHARACTERISTICS

By James R. Wait, M.A.Sc., Ph.D.*

It is the purpose of this article to outline a scheme for measuring the angle of arrival, azimuth, and polarization of a downcoming radio wave. Despite the existence of extensive literature[1,2,3,4,5] on the subject of polarization errors in direction finders and related subjects, there does not seem to have been a method available for analysing the characteristics of an arbitrary downcoming plane wave from measurements made on the ground.

Essentially, there are four quantities needed to characterize the incident wave completely: azimuth α, elevation angle θ, and electric field components E'' and E' parallel and perpendicular to the plane of incidence, respectively. Two of these quantities, E'' and E', are complex. In most cases, however, a knowledge of the magnitudes and relative phase between E'' and E' is all that is required. In view of these facts, it is believed that the simplest scheme to accomplish the task consists of a crossed-loop direction finder operating in combination with a four-element-type Adcock aerial.

Choosing a cartesian co-ordinate system, with the z axis directed vertically, the surface of the ground is taken to be the xy plane. A pair of crossed loops is located at the origin. The intersection of their planes coincides with the z axis. The loop axes make an angle α with the x and y axis, respectively. Without loss of generality, the normal of the downcoming wave makes an angle θ with the z axis, and is contained in the yz plane. The E'' component of the field is contained in the yz plane, whereas the E' component is perpendicular to this plane. The situation is illustrated in Fig. 1. The four vertical aerial elements of the 'Adcock' are now located at a distance $s/2$ from the origin; the pair 'a' and 'b' are on the axis of loop No. 1, and the pair 'c' and 'd' are on the axis of loop No. 2.

Since the vertical pairs of the 'Adcock' respond, in the ideal sense, only to the E'' component of the downcoming wave, it is desirable to establish the azimuth angle α from the voltages V_a, V_b, V_c and V_d induced in the four elements. These are given by

$$\left.\begin{aligned} V_a &= A \exp[-ik \sin\alpha \sin\theta\,(s/2)] \\ V_b &= A \exp[+ik \sin\alpha \sin\theta\,(s/2)] \\ V_c &= A \exp[-ik \cos\alpha \sin\theta\,(s/2)] \\ V_d &= A \exp[ik \cos\alpha \sin\theta\,(s/2)] \end{aligned}\right\} \quad \ldots \quad \ldots \quad (1)$$

where $k = 2\pi/\text{wavelength}$,

and $A = E'' \sin\theta[1 + R''(\theta)]\,h$

with h the effective height of the vertical elements, which are assumed identical. $R''(\theta)$ is a reflection

* *National Bureau of Standards, Boulder Laboratories, U.S.A.*

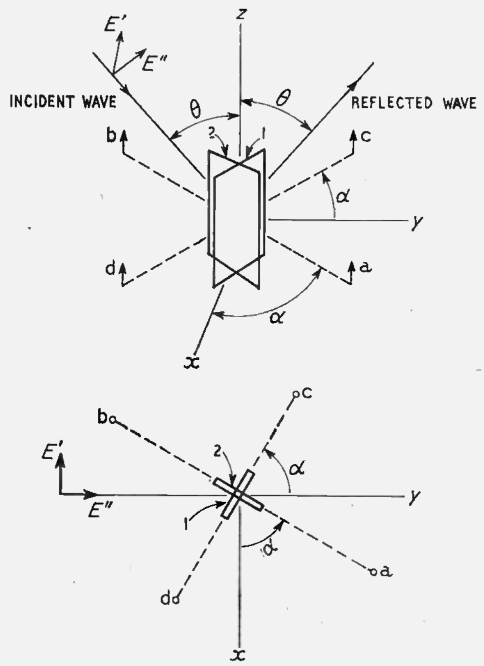


Fig. 1. *Combined 4-element Adcock and crossed-loop direction finder.*

coefficient for plane waves with parallel incidence, being given by

$$R''(\theta) = \frac{N^2 \cos\theta - (N^2 - \sin^2\theta)^{\frac{1}{2}}}{N^2 \cos\theta + (N^2 - \sin^2\theta)^{\frac{1}{2}}} \quad \ldots \quad (2)$$

where $N = \left[\dfrac{\sigma + i\omega\epsilon}{i\omega\epsilon_0}\right]^{\frac{1}{2}}$ is the complex refractive index of the ground expressed in terms of the conductivity σ, the dielectric constant ϵ of the ground and the dielectric constant ϵ_0 of the air.

In accordance with the well-known principle of the 'Adcock' the voltage differences $(V_b - V_a)$ and $(V_d - V_c)$ are formed and applied to the field coils of a goniometer device. The bearing angle ϕ, indicated by the search coil, is then given by

$$\tan\phi = \frac{V_b - V_a}{V_d - V_c} = \frac{\sin[ks \sin\alpha \sin\theta]}{\sin[ks \cos\alpha \sin\theta]} \quad \ldots \quad (3a)$$

When $ks \sin\theta \ll 1$, it is seen that

$$\tan\phi \simeq \tan\alpha \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3b)$$

or $\alpha = \phi$, or $\phi + \pi$.

In most applications to low-frequency direction finding, s is small compared to the wavelength, and the 'spacing error', resulting when Equ. (3b) is used in place of Equ. (3a), is negligible. Therefore, the azimuth is obtained with an ambiguity of π, which can be removed by a special sense device.

The voltages V_1 and V_2 are induced in the loops Nos. 1 and 2, respectively, by the magnetic field components in the direction of the loop axes. The components of the tangential magnetic field at the surface of the ground are given by

$$\left.\begin{aligned} H_x &= [1 + R''(\theta)]\, E''/\eta \\ \text{and}\quad H_y &= [1 - R'(\theta)]\, E'/\eta \end{aligned}\right\} \qquad (4)$$

with $\eta = 120\,\pi$ ohms, the characteristic impedance of free space. $R''(\theta)$ is defined by Equ. (2), and $R'(\theta)$, (the corresponding reflection coefficient for a perpendicularly-polarized incident wave) is given by

$$R'(\theta) = \frac{\cos\theta - (N^2 - \sin^2\theta)^{\frac{1}{2}}}{\cos\theta + (N^2 - \sin^2\theta)^{\frac{1}{2}}} \qquad (5)$$

The voltages V_1 and V_2 induced in the loops Nos. 1 and 2, respectively, are

$$\left.\begin{aligned} V_1 &= \eta h_1\,(H_x \cos\alpha + H_y \sin\alpha) \\ \text{and}\quad V_2 &= \eta h_1\,(-H_x \sin\alpha + H_y \cos\alpha) \end{aligned}\right\} \qquad (6)$$

where h_1 is the 'effective height' of the loop aerials. For a small loop of area A and turns N, it is well known that

$$h_1 = i\,k\,N\,A$$

On combining Equs (4) and (6), it follows that

$$\left.\begin{aligned} V_1/h_1 &= E''(1+R'')\cos\alpha + E'(1-R')\sin\alpha\cos\theta \\ V_2/h_1 &= -E''(1+R'')\sin\alpha + E'(1-R')\cos\theta\cos\alpha \end{aligned}\right\} \qquad (7)$$

It appears that another independent equation is still required. A quantity chosen is the average voltage $\overline{V}$ induced in the elements of the Adcock; this is given by

$$\overline{V} = \frac{V_a + V_b}{2} = \frac{V_c + V_d}{2} = hE_0''(1+R'')\sin\theta \qquad (8)$$

The complete solution is now obtained by combining Equs (6), (7) and (8), and it is written conveniently as follows

$$\text{(i)}\quad \tan\alpha \simeq \tan\phi \qquad (9a)$$

$$\text{(ii)}\quad E_0''(1 + R'') = [V_1\cos\alpha - V_2\sin\alpha]/h_1 \qquad (9b)$$

$$\text{(iii)}\quad \sin\theta = \frac{\overline{V}h_1}{(V_1\cos\alpha - V_2\sin\alpha)h} \qquad (9c)$$

$$\text{(iv)}\quad E_0'(1 - R') = \frac{V_1\sin\alpha + V_2\cos\alpha}{h_1\cos\theta} \qquad (9d)$$

The above four relations are written in this way to illustrate the idea that α, $E_0''(1 + R'')$, $\sin\theta$ and $E_0'(1 - R')$ are to be obtained successively from the measured data (ϕ, V_1, V_2 and $\overline{V}$).

In most instances, at low radio frequencies, $|N^2| \gg 1$ for average or well-conducting ground. In this case $(1 + R'')$ and $(1 - R')$ can be replaced by 2. If the ground is not sufficiently well conducting, it is necessary to measure another independent quantity in order to calculate R'' and R'. A possible choice would be the voltage V_v induced by the downcoming wave in a loop with its axis oriented in the vertical direction. It is given by

$$V_v = h_2(1 + R')\,E'\cos\theta \qquad (10)$$

where h_2 is the effective height of the loop. Combining Equs (10) and (9b) it easily follows that

$$\frac{1 - R'}{1 + R'} = \left[\frac{V_1\sin\alpha + V_2\cos\alpha}{V_v}\right]\frac{h_2}{h_1} \qquad (11)$$

enabling R' to be solved for explicitly. To simplify the calculations, it is permissible to employ the approximation

$$R' \simeq \frac{\cos\theta - N}{\cos\theta + N} \qquad (12)$$

for $|N^2| \gg \sin^2\theta$, so that

$$N \simeq \cos\theta\left[\frac{V_1\sin\alpha + V_2\cos\alpha}{V_v}\right]\frac{h_2}{h_1} \qquad (13)$$

The reflection coefficient R'' is obtained conveniently from the approximate relation

$$R'' \simeq \frac{N\cos\theta - 1}{N\cos\theta + 1} \qquad (14)$$

also valid for $|N^2| \gg \sin^2\theta$.

Of course, there are other methods available for determining N. The one mentioned above has the merit that the downcoming wave itself is employed to calculate R' and R''.

As a matter of completeness, the complex polarization P of the downcoming wave can be expressed as

$$P = \frac{E'}{E''} = \frac{V_1\sin\alpha + V_2\cos\alpha\,(1+R'')}{V_1\cos\alpha - V_2\sin\alpha\,(1-R')}\cdot\frac{1}{\cos\theta} \qquad (15)$$

This can be conveniently rewritten

$$P = \tan(\alpha + \delta)\left(\frac{N + \cos\theta}{N\cos\theta + 1}\right) \qquad (16)$$

where the approximate values of the reflection coefficient have been used and

$$\delta = \tan^{-1} V_2/V_1$$

In the preceding analysis it is assumed that the surface of the ground is flat. The influence of curvature becomes important at near grazing incidence. In this region (i.e., θ near 90°) modified forms for $1 + R''$ and $1 - R'$ must be used[6].

This concludes the theoretical outline of this proposed method to measure the characteristics of a downcoming ionospherically-reflected wave.

REFERENCES

[1] R. L. Smith-Rose and R. H. Barfield, "An Investigation of Wireless Waves Arriving from the Upper Atmosphere", *Proc. roy Soc.*, 1926(A), No. 110, p. 580.

[2] R. L. Smith-Rose and R. H. Barfield, "Further Measurements of Wireless Waves Received from the Upper Atmosphere", *Proc roy. Soc.*, 1927(A), No. 116, p. 682.

[3] R. H. Barfield and W. Ross, "A Short Wave Adcock Direction Finder", *J. Instn elect. Engrs*, 1937, Vol. 81, p. 682.

[4] H. W. Breuninger, "Brauchbarkeits grenzen des n-Mast Adcock-Peilers", *Hochfrequenztech. u. Elekroakust.*, 1942, Vol 59, p. 50.

[5] F. Adcock and C. Clark, "The Location of Thunderstorms by D.F. Methods", *Proc. Instn elect. Engrs*, Part III, 1947, Vol. 94.

[6] J. R. Wait and A. M. Conda "Pattern of an Antenna on a Curved Lossy Surface", *Trans. Inst. Radio Engrs*, 1958, Vol. AP-6, p. 348.

NUCLEAR REACTOR COURSES

The United Atomic Energy Authority has announced that a second course on the control and instrumentation of reactors will be held at the Harwell Reactor School from 7th to 17th July 1959 inclusive, and will be open to British and overseas students.

A course for senior technical executives has also been arranged and will be held at the Reactor School from 15th to 25th June 1959 inclusive.

The fee for each course is 50 guineas (exclusive of accommodation) and those who are interested should apply for forms and details from The Principal, Reactor School, A.E.R.E., Harwell, Didcot, Berks.

Transistor Equivalent Circuit

ALLOYED-JUNCTION TYPES AT HIGH FREQUENCIES

By D. A. Green, B.Sc.(Eng.), A.M.I.E.E.*

SUMMARY. *The purpose of this article is to present a new equivalent circuit for a transistor, which is valid at all frequencies where the device gives useful gain. The arrangement is based on the well-known low-frequency one, but contains capacitive elements to represent the reactive effects observed. From this circuit the performance of a particular transistor can be calculated and a discussion of the three available connections shows that the common emitter is the most useful for high-frequency amplifiers. Formulae are given for the hybrid parameters, power gain and the conditions necessary for optimum gain to be obtained in the common-emitter connection. Calculations based on these formulae are compared with practical measurements for a sample of ten typical transistors and the agreement is shown to be within the limits of experimental error.*

In the application of complex devices, such as transistors, it is often convenient to represent physical processes by means of well-known circuit elements. Such equivalent circuits should accurately represent the device over its whole useful frequency range.

The T equivalent circuit (consisting of three resistances and a current generator) is well known, but it only accurately represents a transistor at low frequencies. For audio-frequency transistors, reactive effects (not predicted by the T circuit) occur at frequencies above a few kilocycles per second. While for currently available high-frequency transistors the range of usefulness may be extended by an order of magnitude or so, at the frequencies where such transistors are often used, this T circuit is of little use.

That transistors exhibit reactive effects has long been established. Shockley[1] has calculated the capacitance existing across a reverse-biased p-n junction, such as a collector-base diode, and the frequency dependence of the common-base current gain has long been known. Many equivalent circuits have been proposed (most of which suffer from some disadvantage) and it is not proposed to deal extensively with them. The object of the work reported here was to modify the accepted low-frequency circuit to account for experimentally observed discrepancies.

List of Symbols

c_c = collector capacitance
c_e = emitter capacitance
c_f = feed-through capacitance
C_n = neutralizing capacitance
D = minority carrier diffusion constant for the base material
F_m = the highest frequency at which a transistor will oscillate in a lossless circuit
g_m = mutual conductance
h = h_{11} for the intrinsic transistor
h_{11}, h_{12}, h_{21}, h_{22} = The four hybrid parameters of a quadripole defined by the equations

$$V_1 = I_1 h_{11} + V_2 h_{12}$$
$$I_2 = I_1 h_{21} + V_2 h_{22}$$

i_b = current flowing in the base resistance of a transistor
i_c = collector current
K = Boltzmann's constant
k = transformer turns ratio
G = power gain
q = electronic charge
r_b = intrinsic base resistance
r'_b = extrinsic base resistance
r_c = collector slope resistance measured in common emitter
r_e = emitter resistance
R_g = optimum generator resistance
R_L = optimum load resistance
R_n = neutralizing resistance
T = absolute temperature
V_{eb} = base to emitter voltage
W = base thickness
Y = admittance
Z = impedance
α = common-base current gain
β = common-emitter current gain
ω = angular frequency
ω_g = mutual conductance cut-off frequency
ω_c = collector cut-off frequency, (as defined in paragraph on 'Power Gain in Common Emitter')
ω_α = common-base current-gain cut-off frequency
ω_β = common-emitter current-gain cut-off frequency

The Equivalent Circuit

Departures from the purely resistive T network may be observed by measuring any of the commonly used sets of quadripole parameters over a frequency range. The hybrid parameters are the most convenient practical measure of a transistor, and the common-emitter connection is probably the most used arrangement.

Performance of a Typical Transistor

The four h parameters for a typical high-frequency transistor are shown in Figs. 1 to 4 for the common-emitter connection over the frequency range 1 kc/s–10 Mc/s. The quantity g_m in Fig. 3 is defined as $\delta i_c/\delta V_{eb}$. A factor of great interest for small-signal

* *Standard Telephone & Cables Ltd. Now with Submarine Cables Ltd.*

amplifiers is the neutralized, conjugate matched, power gain which is plotted in Fig. 5 for the same transistor.

Both $|h_{21}|$ and $|g_m|$ fall at a rate of 6 dB/octave from their low-frequency value while the variation of $|h_{11}|$ represents a step between purely resistive high- and low-frequency values. The graphs of $|h_{12}|$ and $|h_{22}|$ also indicate step functions between low- and high-frequency values but are modified by a further rise at the highest frequencies. The maximum rate of change of power gain with frequency (in Fig. 5) is 9 dB/octave.

Derivation of an Equivalent Circuit

A convenient starting point is the common-emitter resistive T circuit of Fig. 6. Consider first the graphs

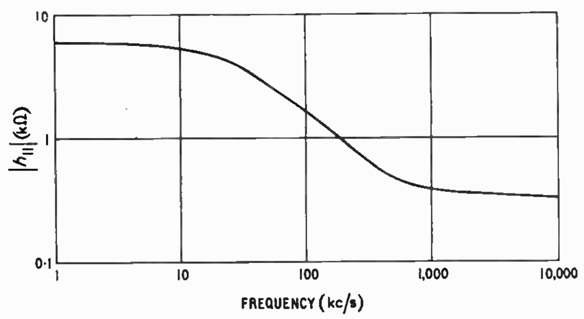


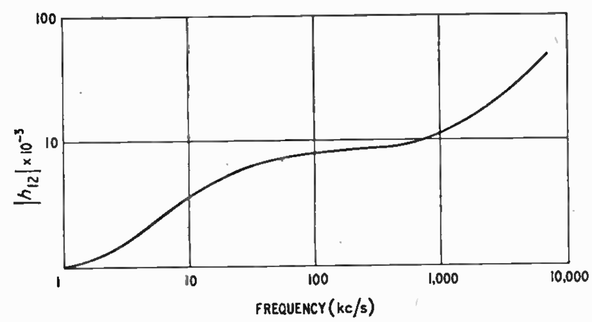
Fig. 1. *Frequency variation of input impedance for a typical transistor*

of $|h_{11}|$, $|h_{21}|$ and $|g_m|$. The shapes of these curves can be obtained by adding a low-pass resistance-capacitance filter to the input circuit. Such a network will give the 6 dB/octave slope observed for both $|h_{21}|$ and $|g_m|$ and will also provide a step function in input impedance. These components are r'_b and c_e as shown in Fig. 7.

If a capacitance c_c is added to represent the collector-junction capacitance (between terminals (b) and (c) in Fig. 6) a step function would be expected for $|h_{12}|$ between the values of $r_e/(r_e + r_c)$ at low frequencies and $c_c/(c_c + c_e)$ at high frequencies. The rise observed in $|h_{12}|$ at higher frequencies may be represented by c_f in Fig. 7, giving a direct coupling between input and output.

Fig. 7 gives a complete representation of the transistor and is shown below to hold at any frequency where the

Fig. 2. *Frequency variation of voltage feedback parameter for a typical transistor*



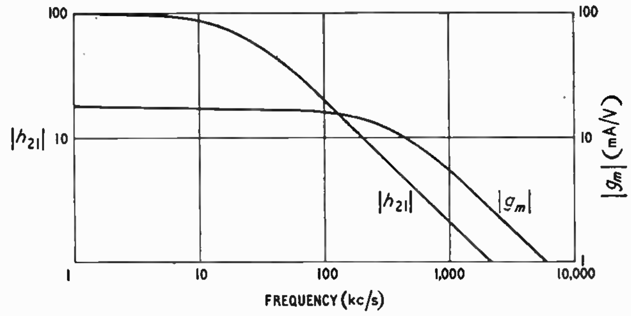


Fig. 3. *Frequency variation of forward transfer function for a typical transistor*

device will provide power gain. It may be considered as an intrinsic transistor consisting of r_e, r_b, r_c and β together with the extrinsic elements c_c, c_e, c_f, and r'_b.

Extrinsic capacitances of a similar order of magnitude to c_f also exist between base and emitter, and collector and emitter leads. They are unimportant in the common emitter since their effects are negligible compared with the reactive effects of c_c and c_e. However,

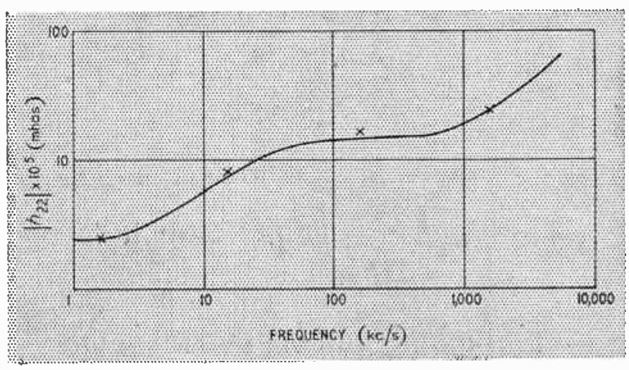
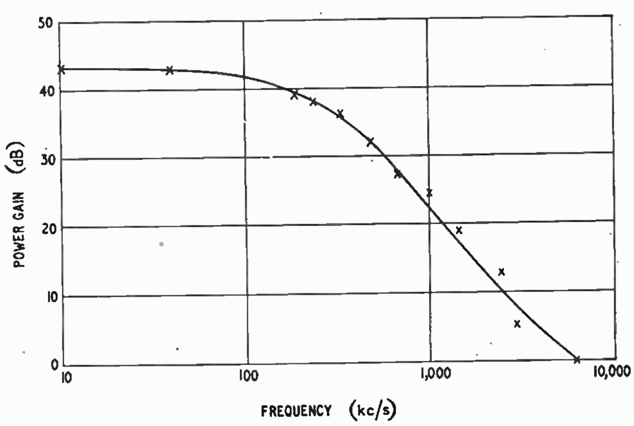


Fig. 4. *Frequency variation of output admittance for a typical transistor. Experimental results indicated by continuous line and calculated results by crosses*

Fig. 5. *Neutralized, conjugate matched, power gain versus frequency for a typical transistor. Experimental results indicated by continuous line and calculated results by crosses*



c_f is important since it is similar to the Miller capacitance of a valve.

Performance in Terms of the Equivalent Circuit

Hybrid Parameters

The circuit of Fig. 7 can be reduced by a series of T and π transformations to a network containing four frequency-dependent parameters. This is shown in Fig. 8, where

$$Y_1 = \frac{r_c\,(1 + j\omega/\omega_\beta)}{H} \quad \cdots \quad (1)$$

$$Y_2 = \frac{r_e + j\omega\,(c_c r_c h + c_f H)}{H} \quad \cdots \quad (2)$$

$$Y_3 = \frac{r'_b[1 - (\omega^2 c_c r_c)/\omega_\beta] + r_b + j\omega c_c r'_b r_c \beta}{H} \quad \cdots \quad (3)$$

$$Y_m = \beta r_c / H \quad \cdots \quad (4)$$

Where
$$H = r_c(h + r'_b)\,(1 + j\omega/\omega_g) \quad \cdots \quad (5)$$

$$h = r_b + \beta r_e \quad \cdots \quad (6)$$

$$\omega_\beta = \frac{1}{(c_e + c_c)h} \quad \cdots \quad (7)$$

$$\omega_g = \frac{r'_b + h}{(c_e + c_c)hr'_b} \quad \cdots \quad (8)$$

In deriving these expressions the following assumptions have been made:

$$\beta \gg 1$$
$$c_c r_c \beta \gg c_e(r'_b + r_e)$$
$$r_c \gg r_e$$

The two angular frequencies, defined as ω_β and ω_g, are where $|h_{21}|$ and $|g_m|$ have fallen to $1/\sqrt{2}$ of their low-frequency values, and the phase shift associated with them is $\pi/4$ radians. The quantity h is h_{11} for the intrinsic transistor in the common-emitter connection.

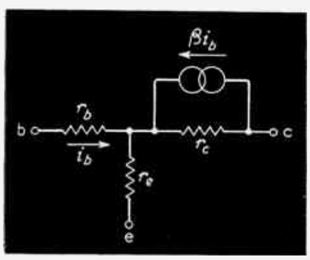


Fig. 6. *The low-frequency T equivalent circuit for a transistor*

From Equs. (1) to (5) the h parameters may easily be derived. If c_f is neglected the results are:

$$h_{11} = \frac{(r'_b + h)\,(1 + j\omega/\omega_g)}{(1 + j\omega/\omega_\beta)} \quad \cdots \quad (9)$$

$$h_{12} = \frac{r_e[1 + (j\omega c_c r_c)/(c_e \omega_\beta r_e)]}{r_c(1 + j\omega/\omega_\beta)} \quad \cdots \quad (10)$$

$$h_{21} = \frac{\beta}{(1 + j\omega/\omega_\beta)} \quad \cdots \quad (11)$$

$$h_{22} = \frac{r'_b + r_b + r_e - \omega^2 c_c r_b r_c + j\omega c_c r_c(h + \beta r'_b)}{r_c(h + r'_b)\,(1 + j\omega/\omega_g)} \quad (12)$$

The only parameter where c_f is important is h_{12}. Including this term the result is:

$$h_{12} = \frac{r_e\left[1 + j\omega\left(\dfrac{c_c r_c}{\omega_\beta c_e r_e} + \dfrac{c_f h}{r_e}\right)\right]}{r_c\left[1 + j\omega\left(\dfrac{1}{\omega_\beta} + \dfrac{c_f h}{r_e}\right)\right]} \quad \cdots \quad (13)$$

Unilateralization and Neutralization

The parameter of prime importance in small signal amplifiers is power gain. For a transistor, this is usually of interest for conjugate matching at input and output terminals. Since instability is often experienced between such loads, neutralization is usually required. The gain so obtained is dependent on the electrode chosen as the common terminal between input and output so that it is first necessary to decide which arrangement gives optimum gain.

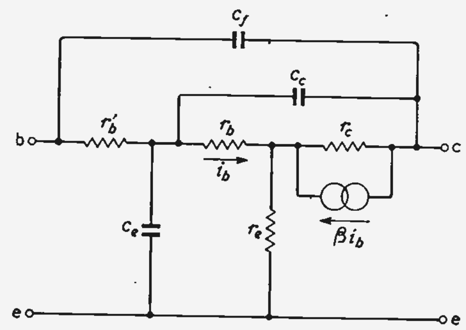


Fig. 7. *The new equivalent circuit*

Any network containing shunt feedback, such as the transistor represented in Fig. 7, is bidirectional; i.e., a voltage applied to its output terminals will produce a voltage at its input. By definition such a network is unilateralized if its feedback is reduced to zero. A special case of unilateralization is obtained if only the feedback produced by reactive elements is cancelled, and is known as neutralization. Either unilateralization or neutralization may be applied to a given network in many ways, in general, the resulting transfer function will depend on the method of cancelling feedback. If feedback is predominantly reactive, there will be little difference between unilateralization and neutralization. It has been shown by Mason[2] that, if unilateralization is achieved with lossless linear bidirectional components the resulting power gain is independent of the terminal chosen as common to input and output.

In general, it is impossible to satisfy this criterion in practice so that some variation in gain is to be expected when the common terminal is altered. Under certain conditions it is possible to achieve a greater gain than would be expected from the Mason U function. As an example some figures given by Giacoletto[3] for transfer performance using the same unilateralization circuit for the three connections may be of interest.

Calculated power gain in Mason circuit	25 dB
In unilateralized common-emitter circuit	29 dB
In unilateralized common-base circuit	17·5 dB
In unilateralized common-collector circuit	0·7 dB

In Fig. 9 is shown the most commonly used method of unilateralization. To show the dependence of gain on the method of removing feedback it must be discussed in detail.

Fig. 9(a) shows a bidirectional four-pole active network, whose input admittance is given by

$$Y_{in} = \frac{1}{Z_i} + \frac{(1+\mu)Z_L}{Z_f(Z_0 + Z_L)}$$

Let Z_L be chosen so that the fraction of the total output power (fed back via the unilateralizing network) is small. Then, for a constant input voltage an output power independent of the power fed back is obtained. Input power is then only a function of input admittance, so that the power gain is directly proportional to the apparent admittance obtained by allowing for the effects of a neutralizing network.

In Fig. 9(b), the same network is shown with shunt unilateralization, achieved by a phase-reversing transformer of turns-ratio $1 : k$ and the impedance Z_n. The condition for unilateralization is $kZ_f = Z_n$. The apparent input admittance of the whole network is then

$$Y'_{in} = \frac{1}{Z_i} + \frac{1}{Z_f} + \frac{1}{kZ_f}$$

Unilateralization could also be achieved by adding an impedance of $-Z_f$ in parallel with Z_f, when the input admittance becomes $1/Z_i$, or by using a series feedback arrangement. Thus, since gain is proportional to Y_{in} for this network, the result is dependent on the type of unilateralization employed.

However, if k is large and if the inherent feedback is small (i.e., $Z_f \gg Z_i$), $Y'_{in} \approx 1/Z_i$. Then, approximate expressions can be obtained by taking unilateralization to imply $Z_f = 0$.

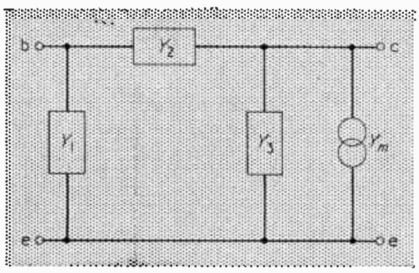


Fig. 8. *π form of the new circuit*

For the experimental work associated with this article the shunt unilateralization system shown in Fig. 9(b) has been adopted. The maximum gain from this arrangement occurs when $k^2 = R_e(Y_{in})/R_e(Y_{out})$, where R_e denotes the real part.

The value of k used was 0·1, and the input admittance has been taken as Y'_{in} above.

Choice of Common Terminal for Maximum Power Gain

Assuming the Mason criterion of lossless bidirectional feedback elements, the choice of a common terminal is unimportant from the point of view of power gain. In practice, these conditions can never be achieved, and the nearest approximation will be obtained from the connection giving the highest feedback impedance, Z_f in Fig. 9. From Fig. 8 this can be calculated for the three configurations, the results being:

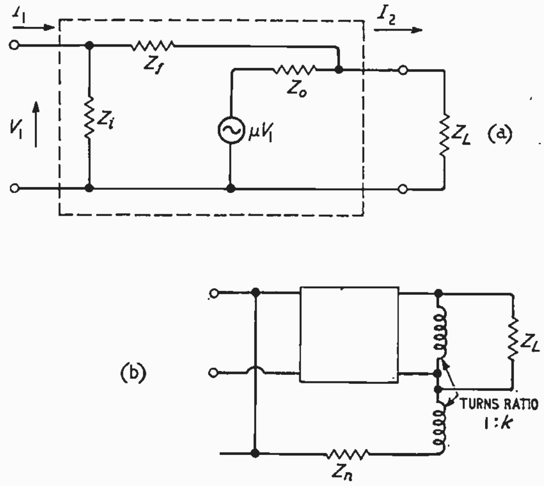


Fig. 9. (a) *Four-pole network with internal feedback.* (b) *Shunt neutralization*

(a) *Common emitter.* $(Z_f)_e = 1/Y_2$

(b) *Common base.* By earthing the point (b) in Fig. 8 the common-base form is obtained, but Y_m then appears between input and output. To an approximation, $v_{be} = 0$ for an applied voltage between collector and emitter, so that

$$(Z_f)_b = 1/Y_3$$

(c) *Common collector.* When point (c) is earthed the current generator appears across the output and $(Z_f)_c = 1/Y_1$. Thus neglecting c_f

$$(Z_f)_e = \frac{H}{(r_e + j\omega c_c r_c h)}$$

$$(Z_f)_b = \frac{H}{r'_b\,[1 - (\omega^2 c_c r_c)/\omega\beta] + r_b + j\omega c_c r'_b r_c \beta}$$

$$(Z_f)_c = \frac{H}{r_c + (j\omega r_c)/\omega\beta}$$

Now, in general, r_c is much greater than any other resistive parameter and $c_e \gg c_c$. Thus it is obvious that $(Z_f)_c < (Z_f)_e$. Considering $(Z_f)_b$: since β is large and r'_b is comparable with h, then for reasonable values of emitter current the imaginary term in the denominator of $(Z_f)_b$ is greater than the corresponding one in $(Z_f)_e$. In general, $r'_b > r_e$, and for frequencies of interest $(1 - \omega^2 c_c r_c/\omega\beta)$ is positive. Thus the denominator of $(Z_f)_b$ is greater than that of $(Z_f)_e$ and since $(Z_f)_c < (Z_f)_e$ it can be inferred that $(Z_f)_e$ is the largest of the three terms.

Thus the common-emitter connection will give the greatest gain at any frequency in a neutralized circuit. When this feedback impedance is highest, the least amount of neutralization is required so that the common emitter would be expected to yield both highest gain and the most stable arrangement. This is supported by the work of Giacolleto, referred to above. For this reason, only the common emitter has been considered in detail in this article.

Power Gain in Common Emitter

Consider a neutralized amplifier. If the neutralization network is treated as part of the amplifier, then it may be considered to have an h_{11}, an h_{21}, an h_{22} and an h_{12} equal to zero. It is well known that maximum power gain is obtained when input and output are conjugately matched. If I_1 is the input current and I_2 the output current under these terminations, the input power will be $|I_1|^2 \, \mathrm{R_e}(h_{11})$ and the output power will be $|I_2|^2/\mathrm{R_e}(h_{22})$ (where $\mathrm{R_e}$ denotes the real part). Thus the power gain will be:

$$G = \left|\frac{I_2}{I_1}\right|^2 \cdot \frac{1}{\mathrm{R_e}(h_{11}) \,.\, \mathrm{R_e}(h_{22})}$$

Now, under conditions of conjugate matching at the output only one half of the short-circuit output current will flow in the load. Thus $\left|\frac{I_2}{I_1}\right| = \left|\frac{h_{21}}{2}\right|$

$$\text{and } G = \frac{|h_{21}|^2}{4\,\mathrm{R_e}(h)_{11} \cdot \mathrm{R_e}(h_{22})} \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

Substituting from Equs. (1) to (5) gives

$$G = \frac{r_c\beta^2}{4(r'_b+r_b+r_e)} \cdot \frac{(1+\omega^2/\omega^2{}_g)}{(1+\omega^2/\omega_g\omega_c)(1+\omega^2/\omega_\beta\omega_g)} \quad (15)$$

$$\text{where } \omega_c = \frac{r'_b + r_b + r_e}{c_c r'_b r_c \beta}$$

If $\omega \gg \omega_\beta$

$\omega \gg \omega_g$

$\omega \gg \omega_c$

then Equ. (15) reduces to

$$G \simeq \frac{\beta\omega_\beta}{4\,c_c r'_b \omega^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

and, if the frequency at which power gain has fallen to unity is defined as F_m, then

$$F_m{}^2 \simeq \frac{\beta f_\beta}{8\,\pi\, c_c r'_b} \simeq \frac{\alpha f_a}{8\,\pi\, c_c r'_b} \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Since $\alpha f_a \simeq \beta f_\beta$, as shown by Shea[4].

F_m is identical with the figure of merit (G) often quoted by American workers.

Optimum Generator and Load Impedances

For a neutralized amplifier the optimum load and generator resistance are given, approximately, by

$1/R_g = \mathrm{R_e}\left(\frac{1}{h_{11}}\right)$ and $R_L = 1/\mathrm{R_e}(h_{22})$, Where $\mathrm{R_e}$ is the real part. From Equs. (9) and (13) these are

$$R_g = (r'_b + h) \cdot \frac{(1 + \omega^2/\omega_g{}^2)}{(1 + \omega^2/\omega_\beta\omega_g)} \quad \ldots \quad \ldots \quad (18)$$

and

$$R_L = \frac{r_c(r'_b + h)}{r'_b + r_b + r_e} \cdot \frac{(1 + \omega^2/\omega_g{}^2)}{(1 + \omega^2/\omega_g\omega_c)} \quad \ldots \quad (19)$$

Neutralization Component Values

If $k = 1$ in Fig. 9(b), the condition for neutralization is that $Y_h = Y_2$. Thus from Equ. (2), if R_n and C_n are the required parallel combination of components, then

$$R_n = \frac{(h + r'_b)\,(1 + \omega^2/\omega_g{}^2)}{(r_e/r_c) + (\omega^2 c_c h/\omega_g)}$$

$$\text{and } C_n = \frac{c_c h\omega_g - (r_e/r_c)}{\omega_g\,(h + r'_b)\,(1 + \omega^2/\omega_g{}^2)} + c_f$$

If the operating frequency is well below ω_g both C_n and R_n are independent of frequency.

TABLE 1

No.	r'_b Ω	r_b Ω	r_c kΩ	r_e Ω	β	c_c pF	c_e pF	c_f pF
1	35	480	73·5	22	23	61·5	1,760	0·46
2	60·8	230	76	32·8	100	40·2	1,750	0·3
3	181	852	77	29·3	46	15·4	1,540	2·2
4	152	2,810	22·7	15	100	17·6	382	4·6
5	182	770	73·5	33·8	60	12·8	948	3·0
6	162	792	71·5	23·6	44	15·6	1,405	2·9
7	382	2,728	23·8	11·6	82	13·9	496	2·8
8	565	2,150	38	43·6	111	19·9	1,660	1·2
9	220	3,720	50	39·5	73	24·8	2,160	2·9
10	194	1,110	44·6	22·7	49	25	835	5·6

Experimental Verification

To check the validity of Fig. 7 a sample of ten transistors was taken, so as to form a representative selection of currently available devices. The results obtained are shown in Table 1. From measurements at a low frequency, values can be assigned to the intrinsic transistor components. A high-frequency measurement of h_{11} gave r'_b, and then from either ω_β or ω_g, c_e was calculated. From a plot of h_{12}, the values of c_c and c_f were obtained.

Since h_{22} is not needed to find the component values, a convenient check is to compare measured and calculated values of it. These are shown in Fig. 4 for the transistor used. Considering the accuracy of measurement for the h parameters the result is considered satisfactory.

From the values given in Table 1, the power gain and maximum frequency of oscillation have been calculated. They are compared with experimental results in Table 2. The average error for power gain is 0·24 dB and for F_m is 0·26 Mc/s. Although individual errors are much larger they are not all in the same direction, and can be attributed to experiment rather than theory. A comparison between theory and experiment over the

TABLE 2

No.	Power Gain		F_m	
	Measured dB	Calculated dB	Measured Mc/s	Calculated Mc/s
1	30·5	33·8	8·2	6·25
2	27·5	33·3	6·7	6·54
3	27	23·8	4·8	5·57
4	32·5	31	11·2	12
5	29·5	33	7·4	7·9
6	27·2	26·2	7·2	6·6
7	29·5	28·6	8·1	7·24
8	18·2	15·7	2·2	2·5
9	20	19·6	3·4	3·3
10	28·6	28	7·2	5·92

frequency range 10 kc/s–10 Mc/s is given in Fig. 5. As a further check, the quantity F_m has been measured for thirty transistors and the results are shown plotted against matched unilateralized power gain in Fig. 10.

When measuring F_m the power gain of the device is tending towards unity and the difference between neutralized and unneutralized results is negligible. It is necessary to match both input and output impedances and to use a lossless frequency-determining circuit. Experimental results show that F_m is not a function of the common terminal, which is in agreement with Mason's results.

Variations with Operating Conditions

It is inherent in any semiconductor device that variations of d.c. supply conditions and temperature produce changes in the equivalent circuit parameters.

The dependence on operating conditions of the intrinsic transistor components are well known, and will not be discussed here.

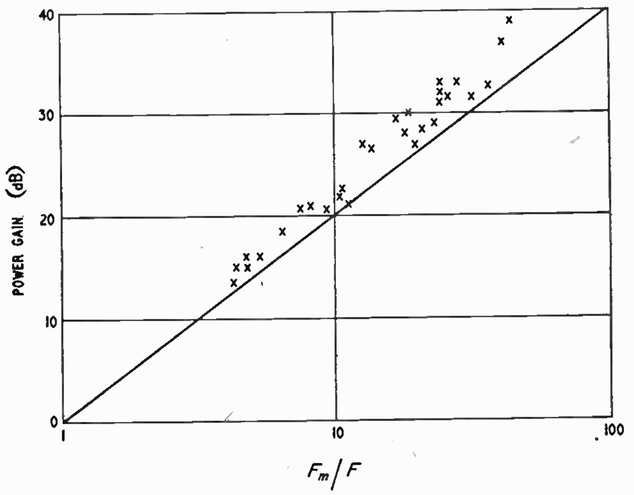


Fig. 10. *Correlation between power gain and maximum frequency of oscillation. The line represents a slope of 6 dB/octave*

Extrinsic Base Resistance

This is largely situated outside the area where transistor action occurs, so that little variation occurs with d.c. condition. A transistor with low r'_b will show a greater dependence on operating conditions because a larger percentage of the semiconductor is influenced by the emitter and collector junctions. Experiment indicates that this resistance is frequency independent.

Emitter Capacitance

It can be shown that this capacitance is given by

$$c_e = \frac{q}{KT} \cdot \frac{W^2}{2D}$$

Thus a linear increase with emitter current is to be expected. However, at high current densities the diffusion constant D is halved in value, halving the slope of c_e versus I_e. A typical result is given in Fig. 11. Due to the Early effect[5], a reduction of c_e with increasing collector voltage is to be expected.

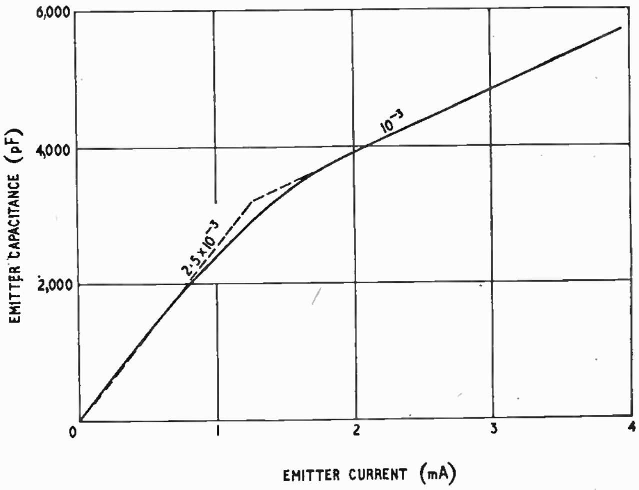


Fig. 11. *Variation of emitter capacitance with emitter current. The slopes indicated are in farads/ampere*

Collector Capacitance

The collector depletion layer width is a function of the reverse bias applied to the junction, thus c_c is a function of collector voltage. The variation is of the form $c_c \propto 1/V_c^n$ where n is between $\frac{1}{2}$ and $\frac{1}{3}$, depending on the physical properties of the junction.

Feed-through Capacitance

Measurements show that this is independent of d.c. conditions and temperature, indicating that it is purely an interelectrode capacitance.

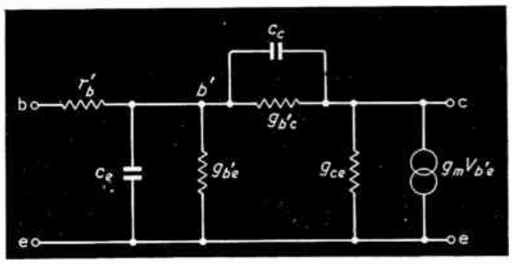
Typical transistor mounts have interlead capacitances ranging from 0·5 to 2 pF, accounting for the larger part of the measured values.

Discussion

From the foregoing results it may be concluded that the circuit of Fig. 7 is an accurate representation of a transistor at any frequency where power gain may be obtained.

At this point it is of interest to briefly discuss one other

Fig. 12. *Equivalent circuit proposed by Giacoletto*



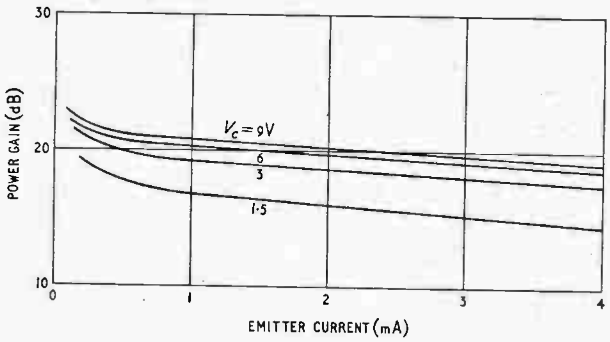


Fig. 13. *Variation of gain with d.c. operating conditions*

proposal for an equivalent circuit. The one generally used is the form proposed by Giacoletto[6], shown in Fig. 12. If the feed-through capacitance is added to this arrangement it can be reduced to the form of Fig. 7.

Despite the apparent redundancy of the new circuit, even when Giacoletto's circuit could be modified to be exact, it is the author's opinion that it contains advantages that justify its introduction. First, Fig. 7 contains the accepted low-frequency equivalent, whereas Giacoletto's does not. Secondly, every component in Fig. 7 can be associated with a physical process in the transistor, so that the effects of changes in design or operating condition can be easily assessed. In the π form, the three conductances are interrelated so that a change in one component alters them all. Finally, as transistors are improved, it is to be hoped that the reactive elements may be disregarded when the low-frequency solution (always contained in any deduction from Fig. 7) is easily obtained.

The results of most interest in the design of small signal amplifiers are the expressions for power gain at any frequency and the maximum frequency of oscillation given in Equs. (15) and (17). The agreement between theory and practice is good enough for all practical applications. It has been stated that power gain falls with a slope of 6 dB/octave, so that F_m may be taken as an indication of high-frequency gain. Equ. (15) shows that the maximum possible slope of gain versus frequency is 12 dB/octave, and in Fig. 5 the maximum slope obtained is 9 dB/octave. Thus, as shown in Fig. 10, an extrapolation down in frequency from F_m at 6 dB/octave will yield a pessimistic figure for gain. The error may be of the order of 10 dB.

Another frequently-quoted parameter is the common-base current-gain cut-off frequency. This is only one of many factors involved and, by itself, is of little use in specifying transistor gain. It is the author's opinion that the only standard of comparison for high-frequency performance is matched and unilateralized power gain. Since it has been shown that maximum gain is obtained in the common emitter it is then logical to quote transistor parameters for this arrangement only and to treat it as the standard connection.

It is not easy to see the variation of power gain with operating conditions from Equ. (15). Practical results are shown in Fig. 13. The increase of gain with collector voltage results from a decrease in c_c and an increase in ω_β, while current variations appear to be of second-order significance only. If neutralization, input and output impedances are not adjusted with change of operating conditions, mismatch will occur. The result of this is to produce a falling gain characteristic with a decrease of either emitter current or collector voltage which is suitable for automatic gain control. However, both input and output impedances are dependent on d.c. conditions so that bandwidth variations will occur with the application of a.g.c.

The equivalent circuit can be modified to give a common-base arrangement if the intrinsic collector resistance is replaced by βr_c and the current generator by αi_e. It is then necessary to introduce a feed-through capacitance between emitter and collector terminals, and c_f may generally be neglected.

Expressions similar to the equations above can be deduced for this network.

From the analogy between common-base transistor and common-grid valve applications it might be expected that this mode would be of greater use at high frequency. However, from Mason's U function it can be inferred that F_m is the same for all connections, and Giacoletto has shown that the most stable configuration is the common emitter. He quotes one transistor in the three arrangements, and without neutralization finds:

Common base:	unstable between 20 kc/s and 3 Mc/s.
Common emitter:	unstable between 15 kc/s and 300 kc/s.
Common collector:	unstable between 10 kc/s and 9 Mc/s.

Thus the common emitter has not only maximum gain to recommend it, but is also the arrangement requiring neutralization over the smallest frequency band. If the device is stable without neutralization, in general the gain difference between neutralized and unneutralized operation is negligible.

So far, only alloy-junction germanium transistors have been measured. There would seem to be no fundamental reason why the equivalent circuit should not hold for silicon devices. No work has been performed on grown junction types where it has been reported that r'_b is frequency dependent. Some work would appear to be necessary before applying the circuit to such types.

Acknowledgements

Acknowledgment is due to Standard Telephones & Cables Ltd., for permission to publish this paper, and to Mr. R. Chapman of that Company for his assistance in discussing the properties of the equivalent circuit.

REFERENCES

[1] W. Shockley, M. Sparks and G. K. Teal, "The P-N Junction Transistor", *Phys. Rev.*, July 1951, Vol. 83, p. 151.
[2] S. J. Mason, "Power Gain in Feedback Amplifiers", *Trans. Inst. Radio Engrs*, June 1954, Vol. C.T.1, p. 20.
[3] L. J. Giacoletto, "Performance of a R.F. Alloy Junction Transistor in Different Circuits", Transistors 1, (R.C.A. Laboratories).
[4] R. F. Shea, "Principles of Transistor Circuits", Chapman & Hall Ltd., London.
[5] J. M. Early, "Effects of Space Charge Widening in Junction Transistors", *Proc. Inst. Radio Engrs*, November 1952, Vol. 40, p. 1401.
[6] L. J. Giacoletto, "Study of P.N.P. Alloy Junction Transistor from D.C. Through Medium Frequencies", *RCA Rev.*, December 1954, Vol. 15, p. 506.

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.

Ultimate Noise Factors

SIR,—The present-day interest in amplifiers, such as masers and parametric amplifiers,[1] which have very low noise temperatures, leads one to consider the ultimate, or best possible noise temperature of such devices.

Let us consider the case of a radar amplifier which is designed to amplify a pulse length of say τ. Then the lowest possible peak power p that this amplifier can receive will be when

$$\tau p = h\nu \qquad (1)$$

Where h = Planck's constant
ν = frequency of reception

This, of course, represents the smallest amount of energy that can exist at a frequency ν. It is, in fact, one photon or quantum of energy.

Now it will be remembered that the noise power, p_n, from a source of temperature T is

$$p_n = kTB$$

where k = Boltzmann's Constant
and B = the receiver bandwidth.

In the radar case we have considered

$$B \approx \frac{1}{\tau}$$

so that

$$p_n \approx \frac{kT}{\tau} \qquad (2)$$

Now if the ultimate noise temperature of an amplifier is called T_{min} when the noise power is equal, on the average, to one photon per pulse length, then we can write

$$p_n = p$$

and

$$T_{min} = T$$

Substituting in (1) and (2) we get the not altogether unexpected result that

$$T_{min} \approx \frac{h\nu}{k}$$

Numerically this is conveniently written as

$$T_{min} \approx 4{\cdot}8 \times 10^{-5} \text{ degree Kelvin per Mc/s.}$$

Thus for instance at 35,000 Mc/s $T_{min} \approx 1{\cdot}7$ °K.

A practical present-day amplifier of this frequency might have a noise temperature of say 17,000 °K and could, therefore, be said to have an 'ultimate noise factor' of 40 dB. No matter what is done, this amplifier cannot greatly benefit by having its noise factor improved by much more than 40 dB.

O. NOURSE.

Admiralty Signal and Radar Establishment,
Portsmouth, Hants.
29th January 1959.

REFERENCE

1 J. P. Wittke, "New Approaches to the Amplification of Microwaves", *R.C.A. Review*, Vol. XVIII, No. 4, p. 441

Transistors

SIR,—Your editorial of November 1958 on transistors calls for comment. It is suggested that a paper by Treharne[1] was a "milestone in the history of the transistor". This sweeping statement is, we feel, inaccurate and misleading. We propose to discuss firstly the limitations of the basic philosophy outlined in your editorial and then to point out a number of basic inadequacies and inaccuracies which surely must remove the above paper from the "milestone" class.

There are obvious topological similarities between certain valve and transistor circuits and the analogies between grid and base, anode and collector, cathode and emitter have been shown in almost every text-book on the subject. Equally obvious is the fact that many circuit techniques developed for valve electronics can be reapplied in transistor circuitry. The point which is not made clearly in the paper being discussed is that effective circuit design is impossible unless it is based on a knowledge of the electrical behaviour of the transistors themselves. Because of the comparative complexity of transistor operation, and the consequent variation of many parameters with operating point and temperature, it is our experience that this knowledge of electrical behaviour must come from a basic physical understanding of the device. This view is shared by many eminent educationalists[2, 3].

Nonetheless, there is a tendency among valve-trained engineers to cling to the use of analogy in transistor circuitry and to avoid making any attempt to understand the comparatively simple principles of semiconductor electronics. This approach often results in poorly designed circuits. Even more important is the fact that the transistor is only one member of a fast-growing family of solid-state devices. Whilst the transistor has some vague resemblance to the thermionic valve most other members of the family are quite different and the analogue concept becomes completely useless.

Surely too much stress cannot be laid on the necessity of the progressive engineer being able to read and intelligently interpret articles appearing in the technical press. To be able to understand and assess the value of advances in the semiconductor field requires a sound background rather than an ever-increasing set of rules.

Treharne's paper presents its transistor material in terms of a set of unproven and often misleading or inaccurate rules—it is an arbitrary "cook book" treatment of a type which can scarcely be recommended to thinking engineers. A careful study of this paper will show that the approach adopted gives rise to errors and omissions in a number of fundamental aspects of transistor application. For example, the concept of power flow is virtually neglected, the thermal properties of transistors are covered incompletely and the importance of high frequency properties is greatly underestimated.

We have proved in our own postgraduate courses that valve-trained engineers can assimilate the physical background material necessary for a thorough appreciation of device behaviour and, furthermore, we have found that this more logical presentation takes no more time than the limited analogue approach suggested in your editorial. Treharne's paper occupies 29 pages, we suggest that 25 pages of papers by Moll and Ebers[4,5] or the 26-page paper of Johnson[6] would be a much more useful introduction to the transistor field. It is, we feel, regrettable that you have drawn your readers' attention to this paper without adequately appraising the worth of its philosophy or content.

Department of Electrical Engineering,
University of Melbourne,
Australia.

D. E. HOOPER.
A. R. T. TURNBULL.
23rd January 1959.

REFERENCES

1 R. F. Treharne, "Analogous Transistor System Design and Nodal Methods of Construction with Applications to Research Equipment and Prototype Evaluation," *Proc. Instn Radio Engrs, Aust.*, July 1958, Vol. 19, p. 319.
2 F. E. Terman, "Electrical Engineers are Going Back to Science," *Proc Inst. Radio Engrs*, Vol. 156, No. 44, p. 738.
3 A. E. Ferguson, "Professional Training of Communication Engineers." *J. Instn Engrs*, 1958, Vol. 30, p. 267.
4 J. L. Moll, "Junction Transistor Electronics," *Proc. Inst. Radio Engrs*, 1955, Vol. 43, p. 1807.
5 J. J. Ebers and J. L. Moll, "Large Signal Behaviour of Junction Transistors," *Proc. Inst. Radio Engrs*, 1954, Vol. 42, p. 1761.
6 H. Johnson, "Basic Transistor Device Concepts," Transistors I (R.C.A., 1956).

Our critics appear to have misunderstood our reasons for saying that Treharne's paper is a "milestone in the history of the transistor". In our opinion, most of this paper could have been, and ought to have been, written in the very early days of the junction transistor. If it had been, we should not have regarded it as a milestone. It is precisely because it has at long last appeared and "brings the outlook on the junction transistor round to the point where, in our opinion, it should have started" that we so regard it.

We do not think that our Editorial should have led anyone to suppose that Treharne's paper contained everything about transistors. We said that it "should be of enormous help to anyone familiar with valves who is making his first tentative approach to the application of transistors". We said just what we meant and our opinion is unchanged. It should be obvious that a paper of only 29 pages which also, as its title states, covers constructional methods, could not possibly be more than an introduction to the subject.

Our critics say that "effective circuit design is impossible unless it is based on a knowledge of the electrical behaviour of the transistors themselves". Surely, this is so obvious that it does not need stating. They say also, however, that "this knowledge of electrical behaviour must come from a basic physical understanding of the device". In fact, the precise knowledge needed for design must come from measurements made on transistors themselves.

We consider the comments on Treharne's paper itself to be unjustified. Probably no aspect is treated completely; that would hardly be possible in a paper of wide coverage. Most aspects, if not all, are adequately treated for the beginner with transistors. The "rules", although they are so labelled, are really concise statements about transistors and their uses, and most of them have a good deal of additional explanatory matter.

In their final paragraph, our critics cite some papers which they consider preferable to Treharne's. We have naturally looked at them. They are good papers but are, in our opinion, unsuitable for the newcomer to the transistor. We still maintain that he will find Treharne's paper most helpful but, when he has gained some experience, let him read Moll's by all means. He will then be in a position to derive most benefit from it.

We discuss the more fundamental matter of learning about transistors in our Editorial on p. 81.—Ed.

Power Line Aerial

Sir,—In the process of a program of research on the propagation of radio waves at very low frequencies, a unique transmitting system has been designed. An eight-mile section of a medium-voltage, single-phase power transmission line has been isolated with parallel-resonant circuits which are tuned to the operating frequency of the antenna. There are two of these 'traps' located at each end of the antenna section (one in each line) and two located at the centre of the antenna. The end traps serve principally to prevent the radio-frequency energy from coupling past the antenna section into the remainder of the power-distribution system. The centre traps isolate the antenna section into two halves so that it can be fed as a balanced centre-fed dipole. These traps are similar in principle to the networks used by power companies employing carrier-current communications systems on their power lines. However, the radio-frequency traps have been designed to present a minimum impedance to the flow of 60-c/s power and a maximum impedance to the radio-frequency power. In this manner it has been possible to operate the transmitter at the centre of the antenna section and draw 50 kW of 60-c/s power from the line as it simultaneously is being employed to radiate very low frequency energy.

A few other protective circuits have been included in the system. Lightning arresters are necessary to protect the traps from lightning surges on the line, and large blocking capacitors are used to keep the high 60-c/s line voltage out of the transmitter. However, the expense of the equipment required to convert the power line into an antenna was a small fraction of the cost required to build a separate antenna. Since the antenna section is half-wave resonant, it was also not necessary to employ tuning devices.

In order to minimize the cancelling reflections of the image antenna which is below the surface of the earth, ground of extremely low conductivity has been selected. The antenna is located in the Sierra Nevada Mountains in California where the underlying substructure is primarily granite. The ground conductivity is 50 times smaller than the conductivity of the Mojave Desert. A paper concerning the problem of the low efficiency of horizontal antennas was presented at the 1958 VLF Symposium at Boulder, Colorado.

At the present writing, the antenna has been used for a small number of propagation experiments. Operating at a frequency of 8,400 c/s and approximately 10 kW into the antenna terminals, measurements of extremely strong ionospheric reflections have been made near vertical incidence. The data indicate rapid and extreme fading effects at certain hours of the day. Pulsing experiments have as yet been unable to reproduce detectable whistler mode propagation.

As mentioned above, the eight-mile section is nearly half-wave resonant at 8,400 c/s. The input impedance of the antenna is approximately $80-j50$ ohms (capacitive). Although a large part of the input resistance consists of the ground loss, selection of the optimum underlying ground material has improved the efficiency of the antenna. No measurement has been made as yet of the radiated power of the antenna. Satisfactory signals have been detected 250 kilometres distant. No measurements have been made at greater distances.

When a stable crystal oscillator has been constructed, it is intended to conduct a number of phase and frequency-stability experiments in addition to the ionospheric measurements. It is hoped that the signal strength will be great enough to conduct the stability measurements on a transcontinental basis.

California Institute of Technology, R. M. Golden.
Pasadena, R. V. Langmuir.
California, U.S.A. R. S. MacMillan.
21st January 1959. W. V. T. Rusch.

New Books

The Technical Writer

By J. W. Godfrey and G. Parr, M.I.E.E. Pp. 340. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 45s.

The term 'technical writer' is one which is more and more being used in a rather restricted sense to denote someone, usually employed by an industrial organization, whose main occupation is the writing of reports, instruction books, and publicity material on technical subjects. A secondary part of his work may be concerned with their production in small or large quantities.

In this book, the term is used somewhat more broadly than this, for aspects of book and article writing are touched upon. Nevertheless, the content of the book shows that it is mainly intended for the technical writer in the modern specialized meaning of the phrase.

The book should be of considerable use to him, for it covers in some detail most matters with which he is concerned. There are chapters on writing, style and presentation, illustrations, printing processes, preparation of copy, editorial procedure, typography, and so on. References to specialized literature on particular aspects of the general subject are included.

The writer of articles and books is likely to find the book less useful, for the considerable part dealing with the technicalities of printing and allied processes is of no direct concern to him. He will, nevertheless, benefit from reading it. The understanding of what happens in the process of turning his MS into printed material, which he will gain from it, will undoubtedly help him in his dealings with editors and publishers. W.T.C.

A Performance Index for Feedback Control Systems based on the Fourier Transform of the Control Deviation

By Jens G. Balchen. Pp. 17. Published by Norges Tekniske Vitenskapsakademi, Trondheim, Norway, and available from Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5, Sweden. Price Kr. 7 (Swedish).

This monograph, which is No. 1 in the Acta Polytechnica Scandinavica Mathematics and Computing Machinery Series, deals essentially with the derivation of a new performance index for determining, in conjunction with a graphical frequency-response analysis, the optimum adjustment of linear control systems.

Principles and Applications of Random Noise Theory

By Julius S. Bendat. Pp. 431 + xxi. Chapman & Hall Ltd. (for John Wiley), 37 Essex Street, London, W.C.2. Price 88s.

This book covers a wider field than its title suggests, for it provides a discussion of a mass of techniques associated with communication theory, probability, prediction, filtering, correlation, etc. The amount of information given is impressive while the presentation is thorough and detailed in the American manner. The thorough-going communications engineer is probably obliged to master this book, but one cannot help hoping that one day somebody will succeed in presenting the subject in a more readily understandable form.

The present treatment is highly mathematical and it is not too much to say that it cannot be understood without closely following the mathematical development.

Oszillatoren mit Schwingkristallen

By W. Herzog. Pp. 317 + xi. Springer-Verlag, Heidelberger Platz 3, West Berlin, Germany. Price DM 45.

A comprehensive treatment of the design of crystal oscillators employing valves and transistors. (In German.)

Metallic Rectifiers and Crystal Diodes

By Theodore Conti. Pp. 155 + ix. Price $2.95.

A well-illustrated non-mathematical book for the technician, servicing engineer and student. Contents include: History of rectifiers; rectifier construction; metallic rectifier characteristics and notation; crystal diode structures and characteristics; basic design data; applications; fault finding, repair and replacement.

Basic Pulses

By I. Gottlieb. Pp. 176. Price $3.50.

An elementary 'picture-book' course on pulse theory, which covers the measurement of pulses, composition of pulses, pulses in frequency-selective circuits, transient analysis of pulses, wave-shaping techniques and pulse generators.

Fundamentals of Transistors. 2nd Edition

By Leonard Krugman. Pp. 168 + xiii. Price $3.50.

The first edition of this book was reviewed in the May 1955 issue of *Wireless Engineer* when it was stated that the book "will be extremely useful to the budding engineer and the more serious technician, and many qualified engineers will find it very helpful. Few who are interested in transistors can afford to be without it".

This new edition contains additional sections which deal with the theory, construction and operation of the surface barrier, intrinsic, drift, avalanche and spacistor types of semiconductor devices. Illustrative circuits and design theory, covered in the sections dealing with amplifiers, oscillators and h.f. applications, have been modified. In addition, review question and lists of references have been added at the end of each chapter.

How to Troubleshoot a TV Receiver. 2nd Edition

By J. Richard Johnson. Pp. 150 + vi. Price $2.50.

Gas Tubes

Edited by Alexander Schure, Ph.D., Ed.D. Pp. 72 + vii. Series No. 166–24. Price $1.50.

An elementary book on gas-filled tubes which includes chapters on the ionization in gases, gaseous rectifiers, gas-tube voltage regulators, thyratrons and other types of gas tubes.

The above five books are available from John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A.

Standard-Frequency Transmissions

CHANGES IN THE MSF STANDARD

(Communication from the National Physical Laboratory)

Since June 1955 the frequencies of the MSF Standard-Frequency Transmissions have been monitored in terms of the caesium atomic frequency standard at the National Physical Laboratory and have been maintained within $\pm$ 5 parts in 10^9 of the nominal value. The corrections published in *Electronic & Radio Engineer* have been given to $\pm$ 1 part in 10^9, a higher accuracy being available on application to the N.P.L. From March 1959 it is proposed to give the corrections to $\pm$ 1 part in 10^{10}, and also to change the form in which the corrections are presented. To many users the corrections are of no interest as the uncorrected values are sufficiently accurate for their purpose. The transmissions are, however, intended to be of service in as many applications as possible and some of these require the highest accuracy that can be given.

Frequencies are, of course, expressed as the number of cycles per unit of time and, when the physicist or radio engineer is concerned with very precise measurements, he must consider the nature of this unit and the accuracy with which it is given. This is rendered all the more important by the developments which have been taking place in the determination and the definition of the astronomical unit of time.

Until recently the unit was based on the rotation of the earth on its axis, the resulting time being known as Universal Time (U.T.) and the unit as the mean solar second. It is known that this unit varies in a periodic manner, with an irregular variation superimposed. A correction is applied for the periodic variation, by observatories, giving a time scale known as U.T.2, which is thus based on the mean rate of rotation of the earth. The N.P.L. atomic standard indicated a nearly uniform change in the unit of time of U.T.2 of 5 parts in 10^9 per year from September 1955 to January 1958[1].

In consequence of the non-uniformity of U.T.2 a new unit, known as the second of Ephemeris Time (E.T.) or simply as the second, was adopted by the International Committee of Weights and Measures in 1956. This is defined as a certain fraction of the tropical year for 1900.0, and is most readily determined by observations of the position of the moon. It is more constant than the previous unit but, unfortunately, the astronomical observations required for its determination are less precise and the results must be averaged over a long interval to give the required accuracy. A joint programme of work between the U.S. Naval Observatory and the National Physical Laboratory has given the average value for the period June 1955–March 1958 in terms of the caesium frequency[2]. The value obtained is 9 192 631 770 $\pm$ 20 c/s. Although the accuracy of the result is considerably less than the precision of the atomic standard, a higher accuracy can be obtained only by averaging over still longer intervals of time. There seems to be a good reason therefore for introducing the value now for defining the frequencies of the MSF service, and for making the second (of E.T.) immediately available.

There is, however, a complication to be considered. By international agreement, observatory time signals and the seconds pulses associated with standard-frequency transmissions are based on U.T.2. At the end of 1958 the difference in the units of time of U.T.2 and E.T. is such that a clock which indicates U.T.2 has a frequency which is about 170 parts in 10^{10} less than one which indicates E.T.

The seconds pulses emitted on the MSF service are generated by the same clock which emits the standard frequencies. Hence, if the latter were based on E.T. the seconds pulses would depart from U.T.2 at the rate of 1·5 milliseconds per day. In order that the pulses may be closely related to U.T.2 it would be necessary to make step adjustments in the MSF pulses at frequent intervals.

In order to avoid frequent step adjustments it is proposed to operate MSF as follows:

1. The standard controlling the MSF transmissions will be set so that the deviation in frequency during 1959 will average 170 parts in 10^{10} low with respect to caesium. The actual frequency of operation will be maintained within $\pm$ 50 parts in 10^{10} of the average value.

2. Near the end of each year a value of the off-set frequency for use during the next year will be announced. Thus, changes in the average frequency of MSF will be made at yearly intervals, or not at all in some years.

3. The corrections will be published in *Electronic & Radio Engineer* as before, and will give the values of the transmitted frequencies in

terms of the caesium standard, based on an assumed frequency of 9 192 631 770 c/s.

4. The phase of the seconds pulses will be maintained in agreement with the time-scale U.T.2 as determined by the Royal Greenwich Observatory, by means of step adjustments of 20 milliseconds or a multiple of this amount made when necessary on the first day of the month. The corrections to the pulses in terms of U.T.2 will continue to be distributed by the Observatory.

Apart from these step adjustments, the time interval between n pulses is 10^5n cycles of the controlling standard and values of frequency obtained by using the pulses are the same as those obtained by using the carrier wave. The frequency corrections published in *Electronic & Radio Engineer* are therefore applicable, if allowance is made for any step adjustments.

The frequency corrections are considered to be accurate to $\pm$ 1 part in 10^{10}. So far as measurements in terms of the caesium standard are concerned, it is immaterial what value is ascribed to its frequency so long as the value is stated. When a value is finally adopted by international agreement, the results of measurements made in terms of the present value can easily be converted to correspond to the new value. The same corrections can be used to express the values in terms of E.T. but the values will have the wider limits of $\pm$ 22 parts in 10^{10}.

Example

Let F be the frequency of the transmission, F_0 its nominal value and ΔF the published deviation, which will be in the region of -170×10^{-10}. Then the frequency of the transmission in terms of the caesium standard with an assumed value of 9 192 631 770 is

$$F_{cs}\ (9\ 192\ 631\ 770) = F_0\ (1 + \Delta F).$$

The frequency in terms of the average value of U.T.2 for 1958 is

$$F_{UT2}\ (1958) = F_0\ (1 + \Delta F + 170 \times 10^{-10}).$$

For the sake of continuity, it may be desired to express the frequency in terms of the assumed value for caesium used previously, namely 9 192 631 830. This is given by

$$F_{cs}\ (9\ 192\ 631\ 830) = F_0\ (1 + \Delta F + 65 \times 10^{-10}).$$

REFERENCES

1 Essen, L., Parry, J. V. L., Markowitz, W., and Hall, R. G., *Nature, Lond.*, 1958, Vol. 181, p. 1054.

2 Markowitz, W., Hall, R. G., Essen, L., and Parry, J. V. L., *Phys. Rev. Letters*, 1958, Vol. 1, p. 105.

Deviations from nominal frequency for January* 1959

Date 1959 January	MSF 60 kc/s 1500 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	+ 2	+ 6
2	+ 2	+ 5
3	+ 2	N.M.
4	+ 1	N.M.
5	+ 1	− 1
6	N.M.	− 1
7	+ 1	− 1
8	+ 1	0
9	+ 2	0
10	N.M.	N.M.
11	+ 2	N.M.
12	+ 2	0
13	+ 2	0
14	+ 2	+ 1
15	+ 2	+ 1
16	+ 2	+ 1
17	+ 2	N.M.
18	+ 2	N.M.
19	+ 2	+ 2
20	+ 2	+ 2
21	+ 2	+ 2
22	+ 2	+ 2
23	+ 2	+ 2
24	+ 2	N.M.
25	+ 2	N.M.
26	+ 2	+ 3
27	+ 2	+ 3
28	+ 2	+ 3
29	+ 2	+ 3
30	+ 2	+ 3
31	+ 2	N.M.

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

MEETINGS

I.E.E.

6th March. "Potted Histories of Instruments Designed for Medical Purposes", by H. S. Wolff, and "Co-ordination of Research in Medical Electronics", by Dr. R. G. Willison, commencing at 6 p.m.

9th March. "Need British Standards be Modified to Obtain Trade Abroad", discussion to be opened by T. M. H. Stubbs at 5.30.

10th March. "The Laplace Transform—A Tool for the Electrical Engineer", discussion to be opened at 6 o'clock by A. C. Sim, B.Sc.

18th March. "New Amplifying Techniques", by C. W. Oatley, O.B.E., M.A., M.Sc., at 5.30.

19th–20th March. Convention on Stereophonic Sound Recording, Reproduction and Broadcasting.

23rd March. "High-Quality Microphones", by M. L. Gayford, B.Sc., at 5.30 in the Lecture Theatre and "Effects of Argon Content on the Characteristics of Neon-Argon Glow-Discharge Reference Tubes", by F. A. Benson, D.Eng., Ph.D. and F. M. Chalmers, B.Eng., at 5.30 in the Tea Room.

2nd April. "Women in Engineering", discussion to be opened at 5.30 by Sir Willis Jackson, D.Sc., D.Phil., Dr.Sc.Tech., F.R.S.

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

Brit. I.R.E.

19th March. "Instrumentation in Field Physiology", by Dr. H. Wolff, at 6.30 p.m.

25th March. "Radio Telemetry", symposium; sessions commence at 3 and 6.30 p.m.

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

The Television Society

13th March. "Training in Television Servicing", by G. C. Barker, to be held at Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2, at 7 o'clock.

VALVE CHART FOR TAYLOR VALVE TESTERS

Compiled and published by Taylor Electrical Instruments Ltd., Montrose Avenue, Slough, Bucks. Price 7s. 6d. (including postage). A new publication containing over 5,000 settings.

NEW TITLE FOR I.E.E. SECTION

At a recent meeting of the I.E.E. Council it was decided to change the name of the Radio and Telecommunication Section. In future it will be known as the "Electronics and Communications Section".

OBITUARY

Mr. Thomas Lydwell Eckersley, F.R.S., B.A., B.Sc., M.I.E.E., F.I.R.E., died on 15th February 1959 at the age of 72. He was best known for his pioneering work on the propagation of radio waves. He served in the Royal Engineers in World War I and joined the Marconi Company in 1919. In 1940 he joined the staff of the Air Ministry and in 1942 became Chief Scientific Adviser to the Inter-Services Ionosphere Bureau.

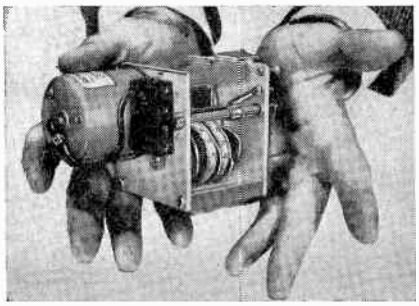
He retired because of ill-health in 1946 but continued as consultant to the Marconi Company.

New Products

Miniature Multiple Circuit Timer

The Electrical Remote Control Co. Ltd. has introduced a new miniature synchronous timer, type AMS, which consists essentially of an electromagnetically-operated clutch, continuously and independently adjustable calibrated time-setting dials (each associated with a snap-action changeover switch), an adjustable mechanism to re-set the timer within 0·5 second, and a miniature synchronous motor.

The timer can be supplied with either 1,



2 or 3 changeover switches and corresponding time-setting dials. The switching capacity at 250 V a.c. is 10 A (with a non-inductive load) and 6 A (with an inductive load).

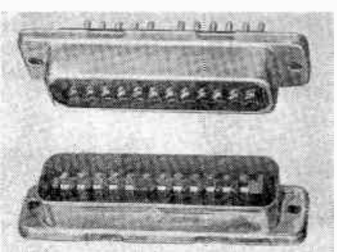
Measuring $2\frac{3}{4}$ in. × $3\frac{1}{2}$ in. × $6\frac{1}{2}$ in., the timer is suitable for operating from a 230 or 110-V, 50 or 60-c/s single-phase supply. Its power consumption is 10 VA. Transformers and d.c./a.c. conversion units can be provided for operation on higher or lower a.c. supplies and 6–220-V d.c. supplies respectively.

Ten standard timing ranges are available, extending from 0–30 seconds to 0–48 hours.

Electrical Remote Control Co. Ltd.,
Elremco Works, Harlow New Town, Essex.

Miniature Connectors

The Contact Connectors Division of Modern Acoustics Ltd. announce that they have added 23- and 13-way models to their 'Modac' range of electrical connectors. Both connectors are polarized and each contact is



capable of carrying up to 10 A at 250 V (maximum).

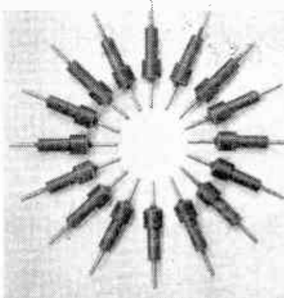
In all 'Modac' connectors reliable electrical connections are maintained by means of a series of spring banks which, on closure of plug and socket, are forced apart by a bevelled blade. This action 'wipes' the surface of the contact blade, which is then gripped (in a pincer fashion) to give a secure connection with a multiple of contact points.

The photograph shows an uncoupled 23-way connector of length $2\frac{21}{32}$ in.

Modern Acoustics Ltd.,
Contact Components Division, Manor Way,
Boreham Wood, Herts.

Moulded Feed-Through Insulator

An improved type of feed-through insulator (W.7001), comprising a solid hot-tin-dipped conductor pin securely bonded into an alkyd body, has been produced by Harwin Engineers Ltd. The makers state that the increased pin size ensures that true concentricity is maintained, and prevents



earth faults resulting from distortion during the moulding process.

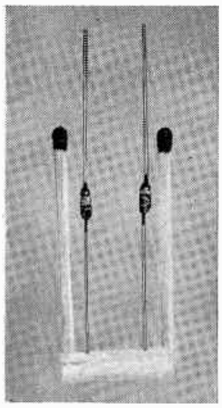
Details from the makers' advance specification include:

Working voltage: 500 V (r.m.s.).
Current capacity: 2 A.
Operating temperature:
80 °C (constant)
100 °C (intermittent).

Harwin Engineers Ltd.,
Rodney Road, Portsmouth, Hants.

Gold-Bonded Germanium Diode

The Mullard gold-bonded germanium diode, type OA7, intended for use with transistors in high-speed computing circuits, features low hole-storage characteristics. Of single-ended all-glass construction, it is suitable for dip soldering into printed-wiring circuits. Measuring 15 mm long by 5·2 mm diameter, it has a p.i.v. of 25 V and a maximum forward current of 50 mA



(mean or peak). At 25 °C ambient temperature, the forward voltage drop at a forward current of 30 mA is 560 mV. The inverse current is 1·9 μA at an inverse voltage of −25 V.

An alternative, sub-miniature version, type OA47, in double-ended construction, is also available. This has similar characteristics, but measures only 7·6 mm long by 3·5 mm diameter.

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

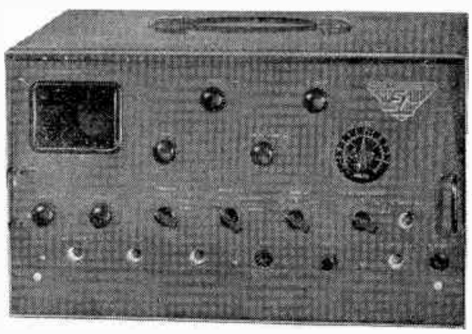
General-Purpose Oscilloscope

A light-weight general-purpose oscilloscope, type 5, has been produced by Erskine Laboratories Ltd.

Its $2\frac{3}{4}$-in. cathode-ray tube (a four-anode type) is positioned horizontally and is viewed through a 45° mirror. This arrangement has certain advantages over conventional designs. It allows a viewing hood to be confined within the instrument, permits a more convenient functional layout of internal circuits and obviates the cramping of front panel controls. Also, the depth of the instrument is no longer dictated by the length of the cathode-ray tube.

By using a stabilized h.t. supply, direct readings of waveform time and amplitude against calibrated graticules (regardless of supply fluctuations) are allowed.

The delay circuit incorporated permits the



viewing of any part of a repetitive waveform up to 20,000 μsec from the synchronizing edge.

All circuit components are contained in a low-temperature section of the instrument and adequate ventilation is provided in such a way that the dust-cover is devoid of vent holes.

Details from the makers' specification include:

Time-base: 9 pre-set positions (3–30,000 μsec).
Delay: 20–20,000 μsec.
Sensitivity: 30 V/cm (X), 45 V/cm (Y).
Y amplifier: Bandwidth 0–4 Mc/s (6 dB) on all ranges; Constant gain (x 100); Input (five ranges), 0·5–500 V.
Synchronization: Better than 0·5 V at Y input.
Trigger: Better than 3 V (r.m.s. sine wave).
Dimensions: 17 in. (wide) × 10 in. (high) × 11 in. (deep).
Weight: 30 lb.

Erskine Laboratories Ltd.,
Scalby, Scarborough, Yorks.

New Epoxide Resin

Bakelite Ltd. have produced a new grade of epoxide resin (DR.19120) which, with a phthalic anhydride hardener (DQ.19121), gives a hot-curing system suitable for potting and casting in the electrical industry. The system has a long pot life, low viscosity when molten, and the castings have very good electrical properties which are retained at elevated temperatures.

The low initial viscosity of the system at 120 °C gives good penetration between components in potting and enables large proportions of inert filler to be incorporated into the mix. Long liquid life at the same temperature allows adequate time for de-gassing under vacuum and only slight exotherm is developed on curing.

Typical characteristics of the system are as follows:

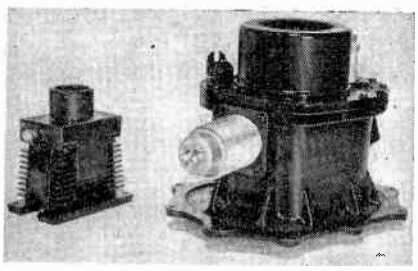
Resin/hardener ratio: 100/30.
Initial viscosity at 120 °C: 160 centipoise.
Working life at 120 °C: 1½ hours.
Setting time at 120 °C: 3½ hours.

Bakelite Ltd.,
12–18 *Grosvenor Gardens, London, S.W.*1.

Light-Weight Transformer

A range of light-weight high-voltage resin-cast transformers and chokes, suitable for use in airborne electronic equipment, has been developed by G.E.C.

The saving in size and weight of a 23-kV

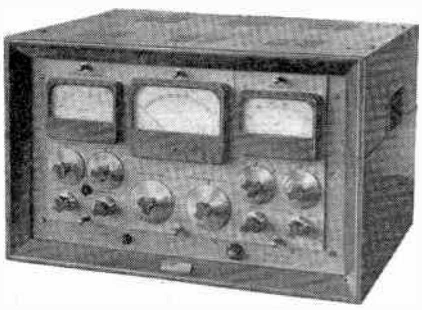


pulse transformer for operation at 70 °C can be seen by comparison with a conventional oil-filled version (shown on the right of the photograph) of equivalent performance. In this case, a weight reduction of better than 4 : 1 has been achieved.

The General Electric Co. Ltd.,
*Magnet House, Kingsway, London, W.C.*2.

Ratemeter

The Ekco type N600 ratemeter is a precision counting instrument for use with all forms of radiation counters. It incorporates a complete spectrum-analysis facility in a discriminator unit which has an automatic or manually-swept single channel. Ranges of fixed or proportional gate widths are provided, some of which are narrow enough to scan and resolve gamma spectra while others are wide enough to select a given peak which results in an improvement of signal-to-noise ratio in conventional counting of known sources. The integrating time constant is adjustable and a special circuit maintains a constant mean probable error (1–10%) which is indicated on a dial.



Some of the details from the makers' specification are given below:

Input sensitivity: 0·005–2 V (negative-going); 5–100 V (positive-going).
Amplifier: Gain, 25 to 1,000; Bandwidth, 1 Mc/s; Rise time, 0·3 μsec.
Pulse Height Analyser: Threshold variable between 5 and 50 V, and 5 and 100 V.
Ratemeter ranges: 3 c/s–100 kc/s.
Resolution: 3 μsec.
Dimensions: 19 in. × 10½ in. × 14 in.
Weight: 70 lb.

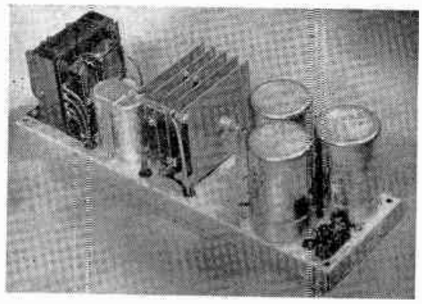
Ekco Electronics Ltd.,
Southend-on-Sea, Essex.

Low-Voltage D.C. Supplies

Advance Components Ltd. have produced a series of eight low-voltage power supplies which provide d.c. outputs ranging from 6 to 48 V at currents up to 4 A, thus making them suitable for use in transistorized equipment where a low-voltage and high-current d.c. supply is required.

The circuit employed comprises a constant-voltage transformer (of the makers' design) followed by a high efficiency semiconductor bridge rectifier and high-capacitance filter.

The output voltage is stabilized within ±1% at full load, with supply-voltage varia-



tions up to ±15%. The stabilization at half full load is generally better than ±1·5%, and the ripple voltage is less than 1·5% (r.m.s.) of the total output volts.

All models can be operated from a 200, 215, 230 or 250-V 50-c/s mains supply and have dimensions not exceeding 17 in. (length) × 5·25 in. (depth) × 6·5 in. (height).

Advance Components Ltd.,
Roebuck Road, Hainault, Ilford, Essex.

High-Speed Flash Tubes

A new range of high-speed cathode-ray flash tubes which produce a flash of the same order of intensity as a gas-filled electronic flash tube, as used in photography, has been introduced by the Electronics Department of Ferranti Ltd.

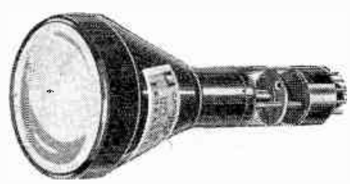
In contrast to the normal behaviour of the cathode-ray tube, these tubes light up all over simultaneously, and have the additional advantage that flashes can be repeated much more rapidly. Flashes of up to one million per second are obtainable. It is also possible to produce long flashes as well as flashes which contain an internal fine structure consisting of a multiplicity of smaller flashes.

They are likely to be of value as high-speed stroboscopes and in high-speed photographic work; in the operation of high-speed photo-electric devices and in Xerographic processes.

Another possible use for these tubes lies in the measurement of short intervals of time or electrical pulses which occur in a random manner and the lamp can be arranged to flash at the beginning and end of a given interval of time or pulse. Alternatively, two different coloured lamps could be used to mark these limits.

Three different types are available: the CL60 triode series (3½ in. diam.), the CL70 miniaturized series (1 in. diam.) and the CE1 diode series (6 in. diam.). They can be supplied with phosphors of all colours including those of ultra-violet and infra-red.

Ferranti Ltd.,
Hollinwood, Manchester, Lancs.



Abstracts and References

COMPILED BY THE RADIO RESEARCH ORGANIZATION OF THE DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH AND PUBLISHED BY ARRANGEMENT WITH THAT DEPARTMENT

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publishers concerned.*

ACOUSTICS AND AUDIO FREQUENCIES

534 : 061.3 — 661

All-Union Acoustical Conference.—V. A. Krasil'nikov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 105–106.) Report on a conference held in Moscow, June 1957, at which 150 papers were read on propagation in inhomogeneous media, radiation and diffraction, waves of finite amplitude, ultrasonics, musical and physiological acoustics and speech investigations.

534.2 — 662

Amplitude and Phase Fluctuations in a Spherical Wave.—V. N. Karavaĭnikov. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 165–176.) Mathematical analysis of fluctuations produced by inhomogeneities of the medium.

534.2 — 663

Correlation of Field Fluctuations.—L. A. Chernov. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 192–194.) Formulae are derived which establish a relation between the correlation function of the field fluctuation and the autocorrelation functions of the amplitude and phase fluctuations. See also 320 of 1957.

534.2-14 — 664

Diffraction and Radiation of Acoustic Waves in Liquids and Gases: Part 2.—M. D. Khaskind. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 92–99.) General expressions are derived for the average value of hydrodynamic forces and moments acting on a body in the presence of diffraction and radiation. Part 1: 3665 of 1958.

534.21 — 665

Waveguide Sound Propagation in One Type of Stratified Inhomogeneous Medium.—Yu. L. Gazarian. (*Akust Zh.*, April/June 1957, Vol. 3, No. 2, pp. 127–141.) An expression is derived for the field of a spherical harmonic point source in a medium in which the sound velocity varies according to Epstein's law (see 319 of 1957). A solution is given for waveguide-type propagation in an inhomogeneous half-space with a totally reflecting boundary.

534.21 — 666

Acoustic Field in a Medium with a Homogeneous Surface Layer.—A. N. Barkhatov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 13–18.) Experimental investigation of the propagation of sound through a space bounded by a homogeneous surface layer over a medium with a constant negative gradient of sound velocity.

534.21 — 667

Sound Amplitude Fluctuations in a Turbulent Medium.—B. A. Suchkov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 85–91.) Report of an experimental investigation of the fluctuation of sound waves propagated through atmospheric layers near the ground.

534.21 — 668

On the Absorption of Sound Waves of Finite Amplitude.—K. A. Naugol'nykh. (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 115–124.) Review of theory and comparison of results of calculations with experimental data, showing that waveform distortion leads to a marked increase in absorption. In water for example at 100 kc/s the absorption coefficient doubles for a pressure increase of the order of 0·01 atm.

534.21 — 669

Attenuation of a Sound Beam Traversing a Layer of Discontinuity in Sound Velocity.—A. N. Barkhatov & I. I. Shmelev. (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 125–127.) A note on the experimental determination of the attenuation of a sound wave traversing a transition layer between two homogeneous media. The application of geometrical theory to the phenomena is considered.

534.21 — 670

Rayleigh-Type Waves on Cylindrical Surfaces.—I. A. Victorov. (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 131–136.) Mathematical treatment of the propagation of Rayleigh-type elastic waves along a convex and a concave cylindrical surface.

534.21-8-14 — 671

Ultrasonic Absorption in Viscous Liquids.—I. G. Mikhaĭlov. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 177–182.) Ultrasonic absorption was measured in castor oil and other oils in the frequency range 0·26–30 Mc/s. Results are tabulated.

534.232 — 672

On the Theory of Piezoelectric Transducers.—K. V. Goncharov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 37–46.) Investigation of the frequency characteristics of X-cut quartz plates, steel, Al, fused quartz and Mg. The effect of an adhering layer is considered in relation to its thickness and acoustic properties.

534.232-8 **673**

Distributed Transducer.—M. Greenspan & R. M. Wilmotte. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 528–532.) In an array of transducers separated by inactive material the input voltages at successive inputs are delayed so that the speed of the electric wave travelling towards the load equals the speed of sound in the transducer material. High power and a wide frequency band can thus be obtained. See also 3012 of 1954 (Rabinow & Apstein).

534.232.001.4 : 534.522.1 **674**

Visualization of Mode Conversion of an Ultrasonic Beam in Fused Quartz.—V. J. Hammond & R. Carter. (*Nature, Lond.*, 20th Sept. 1958, Vol. 182, No. 4638, p. 790.) The process is useful in investigating methods of bonding transducers to fused quartz for use in delay lines.

534.26 **675**

Sound Scattering on Inhomogeneous Surfaces.—Yu. P. Lysanov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 47–50.) Description of a mathematical method for the solution of a system of *n* equations determining the complex amplitude of waves scattered from a flat surface with periodically varying acoustic conductivity for the case of normal incidence.

534.26 **676**

Scattering of Sound by a Thin Rod of Finite Length.—L. M. Lyamshev. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 51–58.) Mathematical analysis of the scattering of a plane monochromatic sound wave by a thin finite rod of circular cross-section shows that vibrations of the rod can produce an angular variation of the scattering characteristics.

534.374-8 **677**

Ultrasonic Interference Filters with Variable Transmission Frequencies.—B. D. Tartakovskiĭ. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 183–191.) The general theory of ultrasonic interference filters of single- and multi-layer type is developed. See 912 of 1953 (Curtis & Hadley).

534.522.1 **678**

Diffraction of Light by Large-Amplitude Ultrasonic Waves.—I. G. Mikhaĭlov & V. A. Shutilov. (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 174–183.) The light intensity distribution is calculated according to the diffraction maxima for different waveforms. The results of calculations taking account of phase modulation are in good agreement with experimental data. See also *ibid.*, April/June 1957, Vol. 3, No. 2, pp. 203–204.

534.6 : 621.385.83 : 537.228.1 **679**

An Electronic-Acoustical Converter.—Yu. B. Semennikov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 73–84.) Methods of rendering an acoustical field visible are noted and a detailed treatment is given of the image-converter tube in which a piezoelectric plate is scanned by an electron beam. See 2184 of 1956 (Oshchepkov et al.).

534.781 **680**

Masking of English Words by Prolonged Vowel Sounds.—J. J. O'Neill & J. J. Dreher. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 539–543.) Results of tests are analysed.

534.79 **681**

Temporary Threshold Shift and Masking for Noise of Uniform Spectrum Level.—J. D. Miller. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 517–522.) An empirical relation between masking and temporary threshold shift is examined experimentally.

534.79 **682**

Temporary Threshold Elevation Produced by Continuous and 'Impulsive' Noises.—W. Spieth & W. J. Trittipoe. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 523–527.)

534.833 **683**

Surface Absorption of Sound in Internally Lined Ducts.—R. Piazza. (*Alta Frequenza*, Feb. 1958, Vol. 27, No. 1, pp. 44–53.)

534.843 **684**

The Distribution of Normal Modes of Vibration in a Rectangular Room according to the Frequency Spectrum and Direction.—Ma Da-Yu (D. Y. Maa). (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 168–173.) The angular distribution approaches a random one for an increase in the dimensions of the room, for a shift of the signal spectrum in the direction of higher frequencies, or for a widening of the frequency band.

534.844 : 534.6 **685**

Testing the 'Echo Parameter' Criterion in Room Acoustics by means of Measurements of Syllable Intelligibility.—H. Niese. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1957, Vol. 66, No. 3, pp. 70–83.) Measurements of the 'echo parameter' (see 2658 of 1957) at various points in four different halls are compared with subjective intelligibility tests at the same points. A close correlation between the results of objective and subjective tests is found.

534.845 **686**

Acoustic Properties of some Types of Sound-Absorbing Material.—Z. N. Baranova & K. A. Velizhanina. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 99–103.) The results of investigations are tabulated.

534.845 : 534.414 **687**

Slit Resonators as Low-Frequency Sound Absorbers.—D. G. Ragavan. (*J. Instn Telecommun. Engrs, India*, Sept. 1958, Vol. 4, No. 4, pp. 213–219.) Theoretical values of resonance frequency, bandwidth and maximum absorption agree fairly well with values obtained in test chambers and studios.

534.846 **688**

Acoustical Design of the Alberta Jubilee Auditoria.—T. D. Northwood & E. J. Stevens. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 507–516.) Impedance-tube and reverberation-chamber data were obtained for materials and components of the two auditoria. Measurements were made in the halls before their completion and concluded by a test concert.

534.851.089 (083.74) **689**

I.R.E. Standards on Recording and Reproducing: Methods of Calibration of Mechanically Recorded Lateral Frequency Records, 1958.—(*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1940–1946.) Standard 58 I.R.E. 19. S1.

621.395.61 **690**

The Effect of Mechanical Vibrations on the Response of Various Types of Microphone.—A. Chiesa. (*Alta Frequenza*, Feb. 1958, Vol. 27, No. 1, pp. 54–60.)

621.395.61 **691**

The Effect on a Receiving System of a Set of Independent Noise Sources Located on the Surface of a Sphere of Finite Radius.—V. I. Klyachkin. (*Akust. Zh.*, April/June 1958, Vol. 4, No. 2, pp. 153–160.) The concepts of concentration coefficient and directivity characteristic, as applied to a receiving system in the field of a large number of independent noise sources distributed over a continuous surface, are discussed, and concentration coefficients for several types of receiving system are determined.

621.395.623.7 : 534.831 **692**

Methods of Generating High-Intensity Sound with Loudspeakers for Environmental Testing of Electronic Components Subjected to Jet and Missile Engine Noise.—J. K. Hilliard & W. T. Fiala. (*J. acoust. Soc. Amer.*, June 1958, Vol. 30, No. 6, pp. 533–538.)

621.395.623.8 **693**

On a Method for Increasing the Stability of Sound Amplification Systems.—L. N. Mishin. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 64–72.) Description of a system of acoustic feedback in which the phase of the feedback is varied continuously. A mathematical justification is given for the choice of a particular value of phase deviation. Experimental results are given.

621.395.625.3 **694**

Braking Action in Magnetic-Tape Recorders.—G. Hartmann. (*Elektronische Rundschau*, Feb. 1958, Vol. 12, No. 2, pp. 45–49.) Consideration of the mechanics of the braking action shows that a constant braking moment is desirable. The shortcomings of practical methods are discussed, and a purely electrical braking system is suggested whose characteristics closely approach the ideal.

AERIALS AND TRANSMISSION LINES

621.372.2 + 621.396.11 **695**

Transmission and Reflection of Electromagnetic Waves in the Presence of Stratified Media.—Wait. (See 939.)

621.372.2.09 **696**

Electric Waves on Delay Lines.—F. Borgnis. (*Elektrotech. Z., Edn A*, 1st June 1958, Vol. 79, No. 11, pp. 383–385.) Propagation conditions for surface waves along plane slow-wave structures are examined.

621.372.221 **697**

A Novel Construction Concept for Linear Delay Lines.—D. Elders. (*Trans. Inst. Radio Engrs*, March 1957, Vol. CP-4, No. 1, pp. 24–28. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1035.)

621.372.221 **698**

The Significance of Phase and Group Delay.—F. Kirschstein & H. Krieger. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 57–60.) The importance of the phase velocity in assessing distortion in coil-loaded transmission lines is shown experimentally.

621.372.8 **699**

Critical Cross-Sections in Irregular Waveguides. — B. Z. Katsenelenbaum. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Nov. 1958, Vol. 123, No. 1, pp. 53–56.) Mathematical analysis for the determination of the amplitude of any wave in a rectilinear irregular slanting waveguide with ideal walls.

621.372.8.002.2 **700**

Waveguide Manufacturing Techniques.—T. Beardow. (*Brit. Commun. Electronics*, Oct. 1958, Vol. 5, No. 10, pp. 772–778.) Survey and comparison of techniques based on fabrication, casting, metal deposition, and printing methods.

621.372.821 **701**

The Characteristic Impedance and Phase Velocity of High-Q Triplate Line.—K. Foster. (*J. Brit. Instn Radio Engrs*, Dec. 1958, Vol. 18, No. 12, pp. 715–723.) An exact solution is obtained for the impedance in the absence of the dielectric support sheet and an expression for the phase velocity is derived. Comparison with experimental results shows that the line parameters can be calculated with an error of about 1%.

621.372.823 **702**

An Approximate Theory for Determining the Characteristic Impedances of Elliptic Waveguide.—R. V. Harrowell. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 289–299.) Much similarity is found between the impedance behaviour of an elliptic waveguide sustaining an H_{C1} wave [see *J. appl. Phys.*, Sept. 1938, Vol. 9, No. 9, pp. 583–591 (Chu)] and a rectangular guide sustaining an H_{10} wave. For a fixed broad dimension in each case the total axial wall current is not changed by reducing the length of the short dimension, and the impedance is modified only by the change in maximum voltage across the guide.

621.372.829 **703**

Helix Waveguide Theory and Application.—H. G. Unger. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1599–1647.) Generalized telegraphist's equations are derived for the curved helix waveguide, and coefficients obtained for conversion from normal modes of this waveguide to normal modes of the metallic waveguide. A radial wave impedance at the helix interface is used to calculate the effect of composite jacket structures in three applications of circular-electric-wave transmission. The analysis is confirmed by measurement.

621.372.829 **704**

Attenuation of the TE_{01} Wave within the Curved Helix Waveguide.—D. Marcuse. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1649–1662.) The helix waveguide has a coating of lossy dielectric and is shielded by a metallic pipe. A perturbation method is used to calculate the change in the field pattern of the TE_{01} mode caused by bending this waveguide; the additional field components produce an e.m. field in the dielectric, which results in energy dissipation and attenuation. The attenuation can be reduced markedly by proper choice of dielectric thickness.

621.372.832.43 **705**

Modified Two-Hole Directional Coupler.—W. G. Voss. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, p. 28.) The modification described enables a two-hole, interference-type coupler to be tuned to give perfect directivity at any wavelength within a wide frequency range.

621.372.85 **706**

On the Theory of Anisotropic Obstacles in Waveguides.—W. Hauser. (*Quart. J. Mech. appl. Math.*, Nov. 1958, Vol. 11, Part 4, pp. 427–437.) "Variational principles for the approximate computation of the elements of the scattering matrix for anisotropic obstacles in waveguides are presented."

621.372.85 **707**

Variational Principles for Guided Electromagnetic Waves in Anisotropic Materials.—W. Hauser. (*Quart. appl. Math.*, Oct. 1958, Vol. 16, No. 3, pp. 259–272.) Approximate expressions are given for propagation in a waveguide partially filled with a material with tensor electromagnetic properties. Variational principles are used to obtain a solution for a rectangular guide containing an infinitely long ferrite slab.

621.396.67 : 621.372.54 **708**

A Three-Band Aerial Combining Network.—Fife. (See 733.)

621.396.67.001.57 **709**

A Microwave Model Equipment for Use in the Study of the Directivity Characteristics of Short-Wave Aerials.—D. W. Morris, E. W. Thurlow & W. N. Genna. (*P.O. elect. Engrs' J.*, July & Oct. 1958, Vol. 51, Parts 2 & 3, pp. 126–131 & 173–179.) Report on investigations of the directivity characteristics of a rhombic aerial and a horizontal array of dipoles. The effect of nearby metallic structures on the performance of aerials was also studied.

621.396.677.4 : 621.396.11 : 551.510.535 **710**

A New Antenna to Eliminate Ground-Wave Interference in Ionospheric Sounding Experiments.—R. S. Macmillan, W. V. T. Rusch & R. M. Golden. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 183–186.)

621.396.677.7 : 621.396.969.33 **711**

Slotted-Waveguide Array for Marine Radar.—H. G. Byers & M. Katchky. (*Electronics*, 5th Dec. 1958, Vol. 31, No. 49, pp. 94–96.) A rectangular waveguide with 126 inclined slots extends along the throat of a 30°-flare horn. A metal grating mounted across the horn aperture eliminates cross-polarization radiation, and full-length chokes between waveguide and horn prevent back radiation. Advantages over a reflector-type aerial are noted.

621.396.677.8 **712**

Split Reflector for Microwave Antennas.—R. L. Mattingley, B. McCabe & M. J. Traube. (*Electronics*, 19th Dec. 1958, Vol. 31, No. 51, pp. 86–88.) Impedance mismatch in pillbox (cheese) and other reflectors is reduced by dividing the reflector into two halves by a metal septum. Each half is fed by conjugate output ports of a short-slot hybrid coupler having suitable phase correction.

621.396.677.85 **713**

On the Axial Phase Anomaly for Microwave Lenses.—G. W. Farnell. (*J. opt. Soc. Amer.*, Sept. 1958, Vol. 48, No. 9, pp. 643–647.) Measurements of axial phase anomaly made on solid-dielectric lenses at 3 cm λ give results which generally agree with those calculated from scalar diffraction theory.

621.396.677.85 **714**

Experimental Investigation of a Homogeneous Dielectric Sphere as a Microwave Lens.—R. N. Assaly. (*Canad. J. Phys.*, Oct. 1958, Vol. 36, No. 10, pp. 1430–1435.) The emergent beam can be rotated through 360° by movement of the source alone. See also 996 of 1957 (Bekefi & Farnell).

AUTOMATIC COMPUTERS

681.142 : 061.4 **715**

Electronic Computer Exhibition.—(*Wireless World*, Jan. 1959, Vol. 65, No. 1, pp. 17–21.) Short notes on special features of computers shown in London from 28th November to 4th December 1958.

681.142 : 061.4 **716**

Learning Machines.—(*Wireless World*, Jan. 1959, Vol. 65, No. 1, pp. 8–9.) A conditional - probability computer and character-recognition machines, exhibited at the symposium on 'The Mechanization of Thought Processes' held at the National Physical Laboratory from 24th to 27th November 1958, are discussed.

681.142 : 537.227 **717**

Ferroelectric Storage Devices.—S. Morleigh. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 678–684.) The switching characteristics of ferroelectric materials, as required for use in digital storage systems, are examined. The properties of single-crystal $BaTiO_3$ capacitors are investigated experimentally.

681.142 : 621.318.57 718

A Track Switching System for a Magnetic Drum Memory.—D. D. Majumder. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 702–705.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.077.6 : 621-526 719

Phase-Selective Gate rejects Quadrature.—B. Fennick. (*Electronics*, 19th Dec. 1958, Vol. 31, No. 51, pp. 92–93.) A phase-reference voltage controls two unmatched diodes which conduct only when the in-phase component is at maximum and the quadrature component at minimum. A circuit designed for a 400-c/s servo amplifier is described.

621.314.22 720

Transformer Design for Zero Phase Shift.—N. R. Grossner. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CP-4, No. 3, pp. 82–85. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 384.)

621.316.825 721

Bounds for Thermistor Compensation of Resistance and Conductance.—A. B. Soble. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CP-4, No. 3, pp. 96–101. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 384.)

621.318.57 : 621.314.7 722

An Introduction to the Use of Transistors in Inductive Circuits.—A. F. Newell. (*Mullard tech. Commun.*, Nov. 1958, Vol. 4, No. 35, pp. 157–160.) The mechanism of a delayed switch-off effect, which is due to avalanche multiplication, is described, and a method of avoiding it in relay switching circuits is given.

621.318.57 : 621.314.7 723

Transistor Switch Design.—A. Gill. (*Electronics*, 5th Dec. 1958, Vol. 31, No. 49, p. 97.) Switching parameters are tabulated for eight types of transistor.

621.318.57 : 621.387 724

An Experimental Gas-Diode Switch.—A. D. White. (*Bell Lab. Rec.*, Dec. 1958, Vol. 36, No. 12, pp. 446–449.) The a.c. impedance is a stable negative resistance of about 225 Ω for frequencies up to 30 kc/s.

621.319.4 725

Dielectric Films in Aluminium and Tantalum Electrolytic and Solid Tantalum Capacitors.—J. Burnham. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CP-4, No. 3, pp. 73–82. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 384.)

621.319.4 : 621.318.134 726

Ferrite-Cored Capacitors.—R. Davidson. (*Research, Lond.*, Sept. 1958, Vol. 11, No. 9, pp. 367–370.) Details are given of British capacitors with ferrite loading for improving their attenuation characteristics.

621.372.029.3 (083.7) 727

I.R.E. Standards on Audio Techniques: Definitions of Terms, 1958.—(*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1928–1934.) Standard 58 I.R.E. 3. S1.

621.372.2 728

A Review on the Analysis of Transients in Electrical Circuits using the Laplacian Transformation.—P. R. Rao. (*J. Instn Telecommun. Engrs, India*, Sept. 1958, Vol. 4, No. 4, pp. 209–212.)

621.372.44 : 621.316.8 729

General Power Relationships for Positive and Negative Nonlinear Resistive Elements.—R. H. Pantell. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1910–1913.) The method developed by Manley & Rowe (2988 of 1956) for the treatment of reactive elements is extended to resistors. Relations are derived which yield modulation efficiency, efficiency of harmonic generation and stability criteria.

621.372.5 730

How Quickly does a Twin-T Respond?—M. Price. (*Canad. Electronics Engng*, Sept. 1958, Vol. 2, No. 9, pp. 40–41.) The response to three different input waveforms is examined using the Laplace transformation. In all cases the output transient is negligible after 1½ cycles of the resonance frequency of the network.

621.372.54 731

Filter Attenuation Characteristics.—M. D. Johnson & D. A. G. Tait. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 710–711.) The use of hyperbolic functions is avoided, and formulae are derived which permit slide-rule computation.

621.372.54 : 621.376.3 732

The Magnitude of the Permissible Circuit Impedances of the Filters in I.F. Amplifiers for F.M. Systems.—E. G. Woschni. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1957, Vol. 66, No. 3, pp. 63–67.) Application of results obtained in 2672 of 1958.

621.372.54 : 621.396.67 733

A Three-Band Aerial Combining Network.—S. L. Fife. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 720–722.) Three prototype filter sections of low-pass, band-pass and high-pass characteristics respectively for bands 1, 2 and 3 are combined in a masthead unit to provide a common output for the three aerial inputs.

621.372.543 734

Normalized Input-Admittance Curves of Two-Stage Band Filters and the Smallest Mismatch Circles for a Given Frequency Range.—H. Hein. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 85–91.)

621.372.543 735

Band Filters with Electronic Bandwidth Control.—C. Kurth. (*Elektronische Rundschau*, Feb. 1958, Vol. 12, No. 2, pp. 39–44.) The design of a two-stage filter with amplification in the feedback line is considered.

621.372.543 736

Null-Point Band Filters and their Theoretical Treatment. — E. Trzeba. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1957 & Jan. 1958, Vol. 66, Nos. 3 & 4, pp. 90–94 & 95–107.)

621.372.553 : 621.397.61 737

The Frequency and Time Characteristics of All-Pass Filters for Delay Equalization in Television Transmission.—H. Dobesch. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1957, Vol. 66, No. 3, pp. 67–70.) The basic circuits given by Bünemann (2240 of 1956) are considered.

621.372.632 : 621.3.072.6 738

Some Properties of a Frequency Stabilizing Circuit.—L. L. Campbell. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CS-5, No. 2, pp. 10–12. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1760.)

621.373 739

The Wave-Mechanical Damped Harmonic Oscillator.—K. W. H. Stevens. (*Proc. phys. Soc.*, 1st Dec. 1958, Vol. 72, No. 468, pp. 1027–1036.) A wave-mechanics treatment is given of damped harmonic motion for the charge in a tuned circuit and the e.m. field in a resonant cavity. The classical frequency appears in the time-dependence of the eigenfunctions, and the classical damping as a decay in the eigenvalues.

621.373.14 : 621.396.96 740

High-Q Echo Boxes.—A. Cunliffe & R. N. Gould. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, pp. 29–33.) An investigation of the occurrence of unwanted modes in a tunable H_0 cylindrical cavity. Suggestions for the suppression of these modes are given.

621.373.421.13 741

Generations of Oscillations with Equally Spaced Frequencies in a Given Band.—D. Makow. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CS-5, No. 2, pp. 13–20. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1760.)

621.373.43 : 621.396.96 742

The Application of Pulse-Forming Networks.—A. Graydon. (*Trans. Inst. Radio Engrs*, March 1957, Vol. CP-4, No. 1, pp. 7–13. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1035.)

621.373.431 743

Miller Sweep Circuit.—C. S. Speight. (*Wireless World*, Jan. 1959, Vol. 65, No. 1, pp. 34–36.) Circuits are described in which both Miller integrator and Puckle flyback circuits are combined to produce a very linear time base.

621.373.431 : 621.314.7 744

Bistable Circuits using Unijunction Transistors.—T. P. Sylvan. (*Electronics*, 19th Dec. 1958, Vol. 31, No. 51, pp. 89–91.) The design and operation of various circuits are explained. Use of the negative-resistance region as one stable state, decreases power requirements and increases switching speed. The application to ring-counter circuits is shown.

621.373.52 745

Transistor 20-kc/s Oscillator with 50 mW Output.—J. F. Berry & L. E. Jansson. (*Mullard tech. Commun.*, Nov. 1958, Vol. 4, No. 35, pp. 122–127.) "A design procedure is described for transistor oscillators employing external feedback and an *LC* resonator. An example is given which delivers an output of 50 mW at 20 kc/s, and which operates from a 9-volt battery. It is suitable for use as a bias oscillator for dictation machines."

621.373.52 746

Graphical Designing of Transistor Oscillators.—W. R. McSpadden & E. Eberhard. (*Electronics*, 5th Dec. 1958, Vol. 31, No. 49, pp. 90–93.) A method is described which is simple but yields results accurate enough for most engineering design calculations. A design example is given for a crystal-controlled oscillator to operate at 1 Mc/s.

621.375.018.756 747

Summary of the Theory of Wide-Band Distributed Amplifiers Suitable for the Amplification of Very Short Pulses.—D. Dosse. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 61–68.)

621.375.024 : 621-52 748

D.C. Amplifiers for Control Systems.—L. S. Klivans. (*Electronics*, 21st Nov. 1958, Vol. 31, No. 47, pp. 96, 98.)

621.375.121.2 : 621.396.621.22 749

An Electronic Multicoupler and Antenna Amplifier for the V.H.F. Range.—K. Fischer. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CS-5, No. 3, pp. 43–48. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1958, Vol. 46, No. 2, p. 515.)

621.375.2.024 750

A Triode-Connected Pentode with Stabilized Anode Current.—B. C. Cox. (*J. sci. Instrum.*, Dec. 1958, Vol. 35, No. 12, pp. 471–472.) Drift in d.c. coupled circuits caused by heater supply voltage variation is minimized by the use of a triode-connected pentode stabilized by variation of grid current with emission.

621.375.2.132.3 751

The Influence of the Output Time-Constant of a Cathode Follower.—C. Edwards. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 712–714.) A nomogram is derived for determining the maximum pulse amplitude, as a function of rise time and output time constant, that may be applied without causing positive grid current.

621.375.23 : 621.317.089.2 752

A Low-Capacitance Input Circuit.—J. C. S. Richards. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 706–708.) Methods of reducing the effect of coaxial-cable capacitance are discussed. A probe containing only passive elements is used with a single triode-pentode valve feedback amplifier to give an input capacitance of 5 pF and a gain of unity over a bandwidth of 2 Mc/s.

621.375.4 : 621.318.57 753

Properties of Hook Transistors in Switching and Amplifying Circuits.—L. M. Vallese. (*J. Brit. Instn Radio Engrs*, Dec. 1958, Vol. 18, No. 12, pp. 725–732.) An analysis is given of the most significant properties of hook and *p-n-p-n* transistors in common-base, common-emitter and common-collector configurations.

621.375.4.029.3 754

Stagger-Tuned Transistor Video Amplifiers.—V. H. Grinich. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. BTR-2, No. 3, pp. 53–56. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 253.) See also 69 of January.

621.375.4.078 755

Transistor Circuits based on the Half-Supply-Voltage Principle.—B. G. Dammers, A. G. W. Uitjens & W. Ebbinge. (*Electronic Applic.*, Aug. 1958, Vol. 18, No. 3, pp. 85–98.) Analysis shows that temperature stabilization of a transistor can be effected by a preceding d.c. coupled amplifying stage stabilized on the half-supply-voltage principle. Practical circuits are described for a.f. amplifiers and stabilized supply units. See also 71 of January (Dammers et al.).

621.375.9 : 538.569.4.029.64 756

Analysis of the Emissive Phase of a Pulsed Maser.—H. H. Theissing, F. A. Dieter & P. J. Caplan. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1673–1678.) A discussion of the emission from a matched cavity at paramagnetic resonance. Results are given for transverse moment, output field, output power, and power gain as a function of time.

621.375.4+621.373.52 757

Transistor Circuit Engineering. [Book Review]—R. F. Shea (Ed.). Publishers: John Wiley, New York, and Chapman & Hall, London, 1957, 468 pp., 95s. (*Nature, Lond.*, 20th Sept. 1958, Vol. 182, No. 4638, pp. 756–757.)

GENERAL PHYSICS

537+538] : 061.3 758

Report of the Meeting of the Swiss Physical Society.—(*Helv. phys. Acta*, 20th Nov. 1957, Vol. 30, No. 6, pp. 457–494.) The text is given of papers read at a meeting held at Neuchâtel, September 1957, including the following:—

(*a*) Electrical Properties of Silver Selenide Ag_2Se.—G. Busch & P. Junod (p. 470, in French).

(*b*) Oscillatory Magnetic Resistance Variation of *n*-Type InSb at Low Temperatures and High Field Strengths.—G. Busch, R. Kern & B. Lüthi (pp. 471–472, in German).

(*c*) The Field Parameters of Galvano- and Thermo-magnetic Effects in Ferromagnets.—G. Busch, F. Hulliger & R. Jaggi (pp. 472–474, in German).

(*d*) Hall and Righi-Leduc Effects in Ferromagnetics.—D. Rivier (pp. 474–478, in French).

(*e*) Effect of a Cubic Electric Field on the Fundamental Level of the Gd^{+++} Ion.—R. Lacroix (pp. 478–480, in French).

(*f*) Hyperfine Splitting in the Paramagnetic Resonance of Pr^{3+} in Ceramic $LaAlO_3$.—H. Gränicher, K. Hübner & K. A. Müller (pp. 480–483, in German).

(*g*) Improvements in an NH_3 Maser.—J. Bonanomi, J. De Prins, J. Herrmann & P. Kartaschoff (pp. 492–494, in German).

537.122 759

Magnetic Susceptibility of an Electron Gas at High Density.—K. A. Brueckner & K. Sawada. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 328–329.)

537.222 760

The Electrostatic Interaction of Two Arbitrary Charge Distributions.—M. E. Rose. (*J. Math. Phys.*, Oct. 1958, Vol. 37, No. 3, pp. 215–222.)

537.226.1 761

Molecular Theory of the Dielectric Constant.—L. Jansen. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 434–444.) A theory of the static dielectric constant is developed on a quantum-mechanical basis. By introducing the concept of 'local field' a molecular version of the general theory is obtained. It is shown that such a molecular theory is fundamentally ineffective in accounting for the observed results within the experimental accuracy.

537.311.31 : 530.145 762

Quantum Theory of the High-Frequency Conductivity of Metals.—M. Ya. Azbel'. (*Zh. eksp. teor. Fiz.*, April 1958, Vol. 34, No. 4, pp. 969–983.) Development of the theory of conductivity of a metal in a high-frequency e.m. field and a constant magnetic field. The amplitude of the quantum oscillations in the high-frequency case is usually larger than in the static case.

537.311.62 763

Quantum Oscillations of the High-Frequency Surface Impedance.—M. Ya. Azbel'. (*Zh. eksp. teor. Fiz.*, May 1958, Vol. 34, No. 5, pp. 1158–1168.) A quantum-mechanical formula is derived and cases involving constant magnetic fields both parallel and inclined to the surface are considered. From an experimental study of surface impedance in a strong magnetic field the shape of the Fermi surface and the velocity of the electrons in it can be determined.

537.533 : 621.385.029.6 764

Solution to the Equations of Space-Charge Flow by the Method of the Separation of Variables.—P. T. Kirstein & G. S. Kino. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1758–1767.) The steady-state behaviour of electron beams with high space-charge densities is analysed.

The equations for irrotational, electrostatic laminar space-charge flow are set up in terms of the action function; these equations are reduced by separation of variables in cylindrical polar coordinates. The method may be extended to include the effect of magnetic fields.

537.56 **765**

Energy Spectrum of Plasma Electrons.—G. Medicus. (*Elektrotech. Z., Edn A*, 1st June 1958, Vol. 79, No. 11, pp. 373–374.) Brief report on results obtained by the graphical method described earlier (731 of 1957).

537.56 **766**

Thermal Conductivity of an Electron Gas in a Gaseous Plasma.—T. Sekiguchi & R. C. Herndon. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 1–10.) Experimental techniques used with Ne and He plasmas are described. Conductivity values obtained by two separate methods are in agreement and consistent with theory.

537.56 : 538.56 **767**

Kinetic Theory of Magnetohydrodynamic Waves.—K. N. Stepanov. (*Zh. eksp. teor. Fiz.*, May 1958, Vol. 34, No. 5, pp. 1292–1301.) Analysis of the propagation of magnetohydrodynamic waves in an ionized gas when the wave frequency is much greater than the frequency of 'short-range' collisions. See also 2418 of 1957.

537.562 **768**

Microwave Method for Measuring the Probability of Elastic Collision of Electrons in a Gas.—J. L. Hirshfield & S. C. Brown. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1749–1752.) A plasma in a d.c. magnetic field has a transverse conductivity component whose reactive part depends on the magnetic field. By measuring the magnetic field necessary to make the reactive part zero the probability of the elastic collision of electrons in He is obtained.

537.562 : 538.56 **769**

Scattering of Microwave Radiation by a Plasma Column.—F. I. Boley. (*Nature, Lond.*, 20th Sept. 1958, Vol. 182, No. 4638, pp. 790–791.) Experiments conducted at 10 cm λ to determine the angular distribution of the e.m. radiation scattered by the positive column of a mercury discharge under resonant conditions are described. Results support the theory of Mackinson & Slade (3532 of 1954). See also 418 of 1958 (Dattner).

538.24 : 621.372.413 **770**

Microwave Faraday Rotation: Design and Analysis of a Bimodal Cavity.—A. M. Portis & D. Teaney. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1692–1698.) An equivalent circuit is developed for the cavity, and the coupling between degenerate modes is expressed in terms of the susceptibility tensor of the material producing the rotation. The results are compared with experimental data.

538.566 **771**

Propagation of Plane Electromagnetic Waves in Inhomogeneous Media.—H. Osterberg. (*J. opt. Soc. Amer.*, Aug. 1958, Vol. 48, No. 8, pp. 513–521.) The laws of propagation along the z direction are derived for infinite inhomogeneous media in which dielectric constant and electrical conductivity are functions of z and the magnetic permeability is constant. Homogeneous media are treated as special cases.

538.566 **772**

Electromagnetic Scattering by Thin Conducting Plates at Glancing Incidence.—J. S. Hey & T. B. A. Senior. (*Proc. phys. Soc.*, 1st Dec. 1958, Vol. 72, No. 468, pp. 981–995.) A large signal is scattered back from a thin plate illuminated edge-on by a field whose magnetic vector is normal to the plate. Experimental measurements show that the currents in the plate are predominant near the edges; their relation to theoretical results is discussed.

538.566 : 516.6 **773**

A Helical Coordinate System and its Applications in Electromagnetic Theory.—R. A. Waldron. (*Quart. J. Mech. appl. Math.*, Nov. 1958, Vol. 11, Part 4, pp. 438–461.) The system described enables problems involving helical symmetry to be solved exactly.

538.566 : 535.42 **774**

The Edge Condition in Diffraction Problems.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 16th June 1958, Vol. 246, No. 24, pp. 3324–3325.) The diffraction of an e.m. wave by a perfectly conducting solid is considered.

538.566 : 535.43] + 534.26 **775**

Theory of Wave Scattering on Periodically Uneven Surfaces.—Yu. P. Lysanov. (*Akust. Zh.*, Jan./March 1958, Vol. 4, No. 1, pp. 3–12.) Description of six approximate mathematical methods for calculating the scattering of sound or e.m. waves over the sea or uneven ground. 79 references.

538.566.2 : 538.22 **776**

Magnetic Double Refraction of Microwaves in Paramagnetics.—F. S. Imamutdinov, N. N. Neprimerov & L. Ya. Shekun. (*Zh. eksp. teor. Fiz.*, April 1958, Vol. 34, No. 4, pp. 1019–1021.) Investigation at 9 375 Mc/s of the rotation of the plane of polarization of an H_{11} wave in a circular waveguide containing a paramagnetic salt, and its dependence on the intensity of the static magnetic field perpendicular to the direction of wave propagation.

538.569.4 : 538.221 **777**

Spin-Wave Analysis of Ferromagnetic Resonance in Polycrystalline Ferrites.—E. Schlömann. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 242–256.) Dipolar interaction is taken into account by means of the spin-wave formalism. Crystalline anisotropy and the polycrystalline nature of the material cause the homogeneous mode of precession to interact with spin waves whose wavelength is of the order of, or larger than, the average linear grain size. The theory predicts a very strong frequency- and shape-dependence of the line width when the homogeneous mode is approximately degenerate with long-wavelength spin waves propagating normal to the d.c. field.

538.569.4 : 538.222 **778**

Paramagnetic Resonance.—J. S. van Wieringen. (*Philips tech. Rev.*, 31st May 1958, Vol. 19, No. 11, pp. 301–313.) The quantum-mechanical theory of the phenomenon is discussed with reference to results of experimental investigations.

538.569.4.029.6 : 539.2 **779**

Technical Applications of Microwave Physics.—D. J. E. Ingram. (*Research, Lond.*, Oct. 1958, Vol. 11, No. 10, pp. 401–407.) Particular reference is made to electron-resonance techniques.

538.569.4.029.64 **780**

Paramagnetic Resonance Spectrum of Gadolinium in Hydrated Lanthanum Trichloride.—M. Weger & W. Low. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1526–1528.) "The paramagnetic resonance spectrum of Gd^{3+} in $LaCl_3.7H_2O$ was measured and found to agree quite well with a spin Hamiltonian with dominant coefficients $b_2^0 = \pm 0{\cdot}0131$ cm^{-1}, $b_2^2 = \mp 0{\cdot}0075$ cm^{-1} at room temperature, and $b_2^0 = \pm 0{\cdot}0099$ cm^{-1}, $b_2^2 = \mp 0{\cdot}0115$ cm^{-1} at liquid air temperature."

539.2 **781**

Theory of Plasma Resonance in Solids.—P. A. Wolff. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 66–69.) The modes of a confined plasma are studied for simple geometries. Modes are closely spaced in frequency and unresolvable unless the sample size is comparable to the Debye length. Observation in small samples is made difficult by line broadening due to surface scattering but might be possible in a suitably designed experiment.

539.2 : 548.0 **782**

Fourier Coefficients of Crystal Potentials.—J. Callaway & M. L. Glasser. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 73–77.) A method is developed for the calculation of the Fourier coefficients of the electrostatic potential of a given distribution of valence electrons in a solid, taking full account of the nonspherical character of the atomic polyhedron.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.14 : 538.69 : 523.165 **783**

Interplanetary Magnetic Field and its Control of Cosmic-Ray Variations.—J. H. Piddington. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 589–596.) A model interplanetary magnetic field is described which may explain some features of solar cosmic-ray increases and also fluctuations in the primary radiation. An attempt is made to show how localized solar fields may create the field, which should be largely radial in form.

523.15 **784**

A Theorem on Force-Free Magnetic Fields.—L. Woltjer. (*Proc. nat. Acad. Sci., Wash.*, 15th June 1958, Vol. 44, No. 6, pp. 489–491.) A variational principle is proved which provides a more direct and satisfactory approach than that given in 3788 of 1958 (Chandrasekhar & Woltjer).

523.164.32 **785**

Nonuniformity in the Brightness of the Sun's Disk at Sunspot Minimum.—J. C. Bhattacharyya. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 43–44.) A comparison is made between the nonuniformity, as derived from ionospheric measurements during eclipses, observed in 1944 and 1954. The results suggest that there are significant differences for the two epochs.

523.164.32 : 523.75 **786**

Observation of a Solar Flare at 4·3-mm Wavelength.—R. J. Coates. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, p. 861.) Scans made across the sun on 25th–27th September 1957 with a radio telescope (1126 *i* of 1958) are compared with corresponding scans for a quiet sun.

523.164.32 : 551.510.535 **787**

Solar Radiation on Decimetre Waves as an Index for Ionospheric Studies.—M. R. Kundu & J. F. Denisse. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 176–178.) The solar noise radiation flux at 10·7 cm λ is compared with other indices of solar activity. It seems to be as good as any other index for ionospheric studies on a monthly time scale and better for shorter time intervals.

523.164.4 **788**

The Trapping of Cosmic Radio Waves beneath the Ionosphere.—G. R. Ellis. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 61–71.) When there is a horizontal gradient in the critical frequency of a layer, incoming extraterrestrial radiation may be trapped between the layer and the ground and propagated over large distances. This would account for the reception of cosmic noise at frequencies lower than the local critical frequency.

523.164.4 : 523.74 **789**

Sudden Cosmic Noise Absorption associated with the Solar Event of 23 March 1958.—K. A. Sarada. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 192–194.)

550.38 **790**

The Relationships between the Secular Change and the Non-dipole Fields.—K. Whitham. (*Canad. J. Phys.*, Oct. 1958, Vol. 36, No. 10, pp. 1372–1396.) The drift and decay contributions to the secular variation have been estimated from isomagnetic and isoporic charts for 1955 in Canada. The westward drift in recent years was found to be significantly smaller than the world-wide average. Relations are obtained between the Gaussian coefficients in the spherical harmonic analyses of the earth's main field and the secular variation. It is shown that one half of this variation is produced by westward drift and that the decay terms are unimportant.

550.389.2 : 551.510.535 **791**

Electron Density Profiles in the Ionosphere during the I.G.Y.—(*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 195–197.) See 438 of February (Smith-Rose).

550.389.2 : 629.19 **792**

Use of Artificial Satellites to Explore the Earth's Gravitational Field: Results for Sputnik II (1957β).—R. H. Merson & D. G. King-Hele. (*Nature, Lond.*, 6th Sept. 1958, Vol. 182, No. 4636, pp. 640–641.)

550.389.2 : 629.19 **793**

Seasonal Illumination of a Circumpolar Earth Satellite at its Extreme-Latitude Orbit Point.—W. N. Abbott. (*Nature, Lond.*, 6th Sept. 1958, Vol. 182, No. 4636, pp. 651–652.)

550.389.2 : 629.19 **794**

Polyhedral Satellite for More Accurate Measurement of Orbit Data of Earth Satellites.—D. R. Herriott. (*J. opt. Soc. Amer.*, Sept. 1958, Vol. 48, No. 9, pp. 667–668.) A 270-face polyhedron would reflect to an observer a light pulse of intensity more than 3 000 times that from a sphere.

550.389.2 : 629.19 **795**

Rotation of Artificial Earth Satellites.—R. N. Bracewell & O. K. Garriot. (*Nature, Lond.*, 20th Sept. 1958, Vol. 182, No. 4638, pp. 760–762.) Radio signals received from Sputnik I were subject to deep regular fading with a semiperiod of about 4 s, probably due to free motion of the satellite about its centroid. These fluctuations were less noticeable in the field strength record of Sputnik III. The advantages of a disk-shaped satellite and methods of eliminating rotational effects are discussed.

550.389.2 : 629.19 **796**

The Faraday-Rotation Rate of a Satellite Radio Signal.—S. A. Bowhill. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 175–176.)

550.389.2 : 629.19 **797**

An Irregularity in the Atmospheric Drag Effects on Sputniks II and III (Satellites 1957β, 1958δ_1 and 1958δ_2).—D. G. King-Hele & D. M. C. Walker. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, pp. 860–861.)

550.389.2 : 629.19 **798**

Satellite Tracking by H.F. Direction Finder.—J. L. Wolfe. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 155–164.) A twin-channel c.r. d.f. with an Adcock aerial was used on the 20-Mc/s signals from satellites 1957α and β. The results obtained and their accuracy are discussed.

550.389.2 : 629.19 : 551.510.535 **799**

Comparison of Phase Difference and Doppler Shift Measurements for Studying Ionospheric Fine Structure using Earth Satellites.—M. C. Thompson, Jr, & D. M. Waters. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, p. 1960.)

551.510.535 **800**

On the Electron Production Rate in the F_2 Region of the Ionosphere.—S. Datta. (*Indian J. Phys.*, Oct. 1958, Vol. 32, No. 10, pp. 483–491.) The F_2 region, between 'bottom' and height of maximum density, is divided into four equal columns, the mean production rate in each being calculated. The results show a diurnal variation of the rate with a single peak at about half an hour before noon.

551.510.535 **801**

Drift Observations Evaluated by the Method of 'Similar Fades'.—E. Harnischmacher & K. Rawer. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 1–16.) A regular interference model is considered as opposed to the usual purely random model. Some of the observed features are well explained by a model which lies between these two views, provided a finite lifetime is assumed for the irregularities. Large changes of drift velocity can be explained by assuming small vertical velocities.

551.510.535 **802**

Bifurcations in the F Region at Baguio, 1952–1957.—V. Marasigan. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 26–31.) A five-year statistical survey of bifurcations at Baguio, Phillipines, is presented, and it is shown that the condition for bifurcation is mainly governed by the parameters h_m and y_m of the F_2 layer; h_m depends on latitude whilst y_m depends on solar activity.

551.510.535 **803**

The Diurnal Variation of f_0F_2 near the Auroral Zone during Magnetic Disturbances.—B. Maehlum. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 187–190.)

551.510.535 **804**

Horizontal Drifts and Temperature in the Lower Part of Region E.—W. J. G. Beynon & G. L. Goodwin. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 180–182.) Drift velocities deduced from the fading of c.w. signals at oblique incidence are related to results obtained using other techniques. A minimum velocity at 80–90 km is indicated and a possible connection with a temperature minimum at the same height is discussed.

551.510.535 **805**

Height Gradient of Electron Loss in the F Region.—V. Marasigan. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 107–112.) A theoretical expression is derived for an exponential height gradient of the electron loss coefficient in the F region on the assumption that this gradient completely accounts for F_1–F_2 bifurcation. Five models are investigated; they are second-power, linear, cosine, parabolic and quasi-parabolic distributions of electron density with height.

551.510.535 **806**

The Electron Distribution in the Ionosphere over Slough: Part 2—Disturbed Days.—J. O. Thomas & A. Robbins. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 131–139.) The results are analysed of the distribution with height for three months in a year of low, and three months in a year of high sunspot number. It is shown that the variation in $h'F_2$ is not a quantitative guide to the changes in height of the F_2 layer during storms. The electron distributions for five individual storms are thoroughly investigated. Some important ionospheric changes have been noticed during a world-wide sudden impulse. Part 1: 1734 of 1958 (Thomas et al.).

551.510.535 **807**

On Instrument Effects in Ionosphere Data.—H. J. Albrecht. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 173–175.)

551.510.535 : 523.164.4 **808**

Abnormal Ionospheric Behaviour at 10 Metres Wavelength.—M. Krishnamurthi, G. S. Sastry & T. S. Rao. (*Curr. Sci.*, Sept. 1958, Vol. 27, No. 9, pp. 332–333.) During observations at Hyderabad, India, of cosmic radio noise total reflection of cosmic noise was observed on three occasions near sunrise. The effect is assumed to be caused by a locally high concentration of matter in the upper ionospheric layers.

551.510.535 : 621.396.11 **809**

On the Approximate Daytime Constancy of the Absorption of Radio Waves in the Lower Ionosphere.—S. Chapman & K. Davies. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 86–89.) The lower part of the D layer is due to photo-detachment of electrons from negative ions the concentration of which is large and nearly constant from day to night. The optical depth is small for the photo-detachment radiation but large for the radiation which can ionize neutral particles.

551.510.535 : 621.396.11 **810**

Ionospheric Absorption over Delhi.—B. V. T. Rao & M. K. Rao. (*J. Instn Telecommun. Engrs, India*, Sept. 1958, Vol. 4, No. 4, pp. 205–208.) An analysis of further measurements made at 5 Mc/s until October 1957. See also 3461 of 1958 (Mazumdar).

551.510.535 : 621.396.11 : 621.396.677.4 **811**

A New Antenna to Eliminate Ground-Wave Interference in Ionospheric Sounding Experiments.—R. S. Macmillan, W. V. T. Rusch & R. M. Golden. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 183–186.)

551.594.5 : 621.396.11 **812**

Radio Reflections on Low Frequencies from 75–90 km Height during Intense Aurora Activity.—W. Stoffregen. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 167–169.)

551.594.6 **813**

Simultaneous Recording of Atmospherics on Four Different Frequency Bands in the Low-Frequency Region.—M. W. Chiplonkar & V. N. Athavale. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 32–37.) The number and intensity of atmospherics are recorded simultaneously on four narrow bands at 85, 125, 175 and 455 kc/s. Results show atmospherics with large field strengths to be less frequent than those with small field strengths on all bands, and the field strength of atmospherics to be a maximum in the 125-kc/s band.

551.594.6 : 551.594.221 **814**

The Relationship between Atmospheric Radio Noise and Lightning.—F. Horner. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 140–154.) In Europe atmospherics in a bandwidth of 300 c/s at 10 kc/s have median amplitude, amplitude range and frequency occurrence in accordance with that expected from lightning discharges to the ground. In Australia atmospherics from tropical storms were found to consist of numerous pulses the largest of which probably originated in ground strokes. The origin of the smaller pulses is not clear. Some observations on atmospherics in the h.f. band are also described.

551.594.6 : 621.3.087.4/.5 **815**

Waveforms of Atmospherics with Superimposed Pulses Recorded with an Automatic Atmospherics Recorder.—B. A. P. Tantry & R. S. Srivastava. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 38–42.) The recorder design is briefly outlined and observations are discussed in which 'stepped' pulses from one lightning discharge are superimposed on the waveform from a second discharge. See also 3824 of 1958 (Tantry).

LOCATION AND AIDS TO NAVIGATION

621.396.932.1 **816**

A New Method of Component Determination in Radio Direction-Finding for Coherent Waves.—H. Gabler & M. Wächtler. (*Elektrotech. Z., Edn A*, 1st June 1958, Vol. 79, No. 11, pp. 385–388.) The three ellipses produced by three suitably spaced crossed-loop aerials with a two-channel c.r. d.f. equipment are superimposed on the screen. Bearings free from night-effect can be obtained even under unfavourable site conditions.

621.396.933.1 **817**

New V.H.F. Direction-Finding Equipment.—(*Brit. Commun. Electronics*, Sept. 1958, Vol. 5, No. 9, p. 681.) A brief description is given of automatic equipment using a rotating Adcock aerial. The operational range is about 100 miles for aircraft flying at 10 000 ft and radiating 5 W.

621.396.96 **818**

Radar Systems with Electronic Sector Scanning.—D. E. N. Davies. (*J. Brit. Instn Radio Engrs*, Dec. 1958, Vol. 18, No. 12, pp. 709–713.) The application to radar of a system previously described in relation to underwater acoustic echo-ranging [3676 of 1958 (Tucker et al.)] is discussed. Information rate is much higher than that of a conventional radar system.

621.396.96 : 551.51 **819**

Radar Echoes from Atmospheric Inhomogeneities.—R. F. Jones. (*Quart. J. R. met. Soc.*, Oct. 1958, Vol. 84, No. 362, pp. 437–442.) A quantitative examination of the possibility of radar echoes being caused by scattering or reflection from atmospheric inhomogeneities shows this to be theoretically possible but necessitating large changes in refractive index.

621.396.962.33 **820**

Decca Doppler and Airborne Navigation.—T. Gray & M. J. Moran. (*Brit. Commun. Electronics*, Oct. 1958, Vol. 5, No. 10, pp. 764–771.) The techniques of transmission and reception used in Doppler systems are examined. The Decca Doppler sensor uses a self-coherent pulsed 'Janus' system with a symmetrical four-beam aerial configuration. A qualitative description of the Decca Integrated Airborne Navigation system (D.I.A.N.) is given.

621.396.963 **821**

Accurate Method for Correction of Slant-Range Distortion in High-Altitude Radars and a Contribution to the Optics of Reflecting Conical Surfaces.—L. Levi. (*J. opt. Soc. Amer.*, Oct. 1958, Vol. 48, No. 10, pp. 680–686.)

621.396.967 : 621.396.65 **822**

Microwave Links for Radar Networks.—Sutherland. (See 976.)

621.396.969.33 : 621.396.677.7 **823**

Slotted-Waveguide Array for Marine Radar.—Byers & Katchky. (See 711.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 **824**

Internal Photoeffect and Exciton Diffusion in Cadmium and Zinc Sulphides.—M. Balkanski & R. D. Waldron. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 123–135.) The absorption spectra of CdS and ZnS and the photoconductivity produced by illumination at a distance from the electrodes were studied to determine the mechanisms of photon-electron interaction and energy transport in these materials.

535.215 **825**

Photoemissive, Photoconductive, and Optical Absorption Studies of Alkali-Antimony Compounds.—W. E. Spicer. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 114–122.) Experimental methods and results are given. Values for the band gaps and electron affinities are tabulated and the conductivity types as indicated by photo-emission data are listed.

535.215 : 546.47-31 **826**

Photo-properties of Zinc Oxide with Ohmic and Blocking Contacts.—H. J. Gerritsen, W. Ruppel & A. Rose. (*Helv. phys. Acta*, 20th Nov. 1957, Vol. 30, No. 6, pp. 504–512.) Report of measurements of primary and secondary photocurrents in fine-grain ZnO layers.

535.215 + 535.37] : 546.482.21 **827**

The Mechanism of Energy Transfer in Cadmium Sulphide Crystals.—I. Broser & R. Broser-Warminsky. (*J. Phys. Chem. Solids*, Sept. 1958, Vol. 6, No. 4, pp. 386–400. In German.) The transfer of the energy of excitation from the point of absorption to non-irradiated regions over relatively large distances may be explained by the scattering and re-absorption of the incident light or the luminescent radiation generated in the crystal. The hypothesis of energy conduction by excitons is not confirmed.

535.215 : 546.482.21 **828**

Photoconductivity and Crystal Size in Evaporated Layers of Cadmium Sulphide.—J. M. Gilles & J. Van Cakenberghe. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, pp. 862–863.) The effect of heating a layer of CdS crystals in contact with a film of evaporated Ag to 500°–600°C in an inert atmosphere is discussed.

535.215 : 546.482.21 **829**

Progress in Cadmium Sulphide.—L. L. Antes. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CP-4, No. 4, pp. 129–132. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 801.)

535.215 : 546.817.231 : 539.23 **830**

Photoconductivity in Chemically Deposited Films of Lead Selenide.—D. H. Roberts & J. E. Baines. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 184–189.)

535.37 **831**

Electrophotoluminescent Amplification.—R. E. Halsted. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1706–1708.) An alternating electric field applied parallel to an excitation gradient in some sulphide phosphors can produce a pronounced characteristic modulation of the photoluminescent emission. Experimental results on this effect are discussed.

535.376 : 546.472.21 **832**

Electroluminescence of Zinc Sulphide Phosphors as an Equilibrium Process.—W. Lehmann. (*J. opt. Soc. Amer.*, Sept. 1958, Vol. 48, No. 9, pp. 647–653.) Predictions of phosphor characteristics based on the equilibrium condition relating monomolecular collision excitation and bimolecular recombination processes are compared with experimental results.

535.376 : 546.472.21 **833**

The Significance of Boundary Layers and Polarization Fields for Electroluminescence.—D. Hahn & F. W. Seemann. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 486–503.) Luminescence of ZnS-based phosphors was investigated using excitation by square pulses of alternating or single polarity. See also 2156 of 1957.

535.376 : 546.472.21 **834**

Cathodo-electroluminescence Phenomena in ZnS Phosphors.—H. Gobrecht, H. E. Gumlich, H. Nelkowski & D. Langer. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 504–510.) Two powder phosphors, one electroluminescent and the other nonelectroluminescent but with pronounced enhancement effect, are investigated.

535.376 : 546.472.21 **835**

Electroluminescence of ZnS Single Crystals with Cathode Barriers.—D. R. Frankl. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1540–1549.) Certain ZnS crystals show electroluminescence predominantly near the cathode. The major emission from Cu-activated crystals occurs as a burst of light when the exciting voltage is suddenly removed, the burst being quenched by a voltage in the initial direction or enhanced by a voltage in the opposite direction. These results, as well as the emission peaks obtained under sinusoidal voltage excitation, are explained in terms of ionization of luminescent centres in a barrier region.

537.226/.227 **836**

Dielectric and Thermal Study of $(NH_4)_2SO_4$ and $(NH_4)_2BeF_4$ Transitions.—S. Hoshino, K. Vedam, Y. Okaya & R. Pepinsky. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 405–412.)

537.226 **837**

Electrical Dispersion Phenomena in Inhomogeneous Dielectrics.—R. Parker & M. S. Smith. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 354–361.) The theory of Koops (162 of 1952) is generalized to include an arbitrary distribution of layer thicknesses and orientations. Observed deviations from the dispersion equations are ascribed to inhomogeneities in the current density vector. The evaluation of surface-layer parameters is discussed.

537.226 **838**

The Ternary Systems BaO-TiO_2-SnO_2 and BaO-TiO_2-ZrO_2.—G. H. Jonker & W. Kwestroo. (*J. Amer. ceram. Soc.*, 1st Oct. 1958, Vol. 41, No. 10, pp. 390–394.)

537.226 : 621.315.612 **839**

Lightweight Ceramic Materials as High-Frequency Dielectrics.—J. L. Pentecost & P. E. Ritt. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CP-4, No. 4, pp. 133–135. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 801.)

537.227 **840**

Room-Temperature Ferroelectricity in Lithium Hydrazinium Sulphate, $Li(N_2H_5)SO_4$.—R. Pepinsky, K. Vedam, Y. Okaya & S. Hoshino. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1467–1468.) A method of producing (and protecting) large crystals is described together with details of an investigation of the ferroelectric behaviour and crystallographic structure.

537.227 **841**

Ammonium Hydrogen Sulphate: a New Ferroelectric with Low Coercive Field.—R. Pepinsky, K. Vedam, S. Hoshino & Y. Okaya. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1508–1510.) Details are given of the ferroelectric behaviour of $(NH_4)HSO_4$ in the range −3°C to −119°C, together with details of the crystallographic structure. A particular feature is the low coercive field which is about 150 V/cm at −13°C.

537.227 : 539.2 **842**

Polarization Fluctuations in a Ferroelectric Crystal.—R. E. Burgess. (*Canad. J. Phys.*, Nov. 1958, Vol. 36, No. 11, pp. 1569–1581.) The theory of thermal fluctuations of electrical polarization in a ferroelectric crystal is considered, with special attention to temperatures in the neighbourhood of the Curie point. By thermodynamical analysis of a dipole model, the fluctuations and their spectral density are related respectively to the real and imaginary parts of the susceptibility of the crystal.

537.227 : 546.431.824-31 **843**

Transition to the Ferroelectric State in Barium Titanate.—D. Meyerhofer. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2. pp. 413–423.) Besides raising the cubic-tetragonal transition about 15°C with an electric field along the cube-edge direction, an orthorhombic phase was induced above the Curie point by a field along a face-diagonal direction. Birefringence, polarization, and dielectric constant were measured above and below the Curie point as functions of field strength and field direction.

537.227 : 546.431.824-31 **844**

Polarization Changes during the Process of Aging in Ferroelectrics of the $BaTiO_3$-Type.—Z. Pajak & J. Stankowski. (*Proc. phys. Soc.*, 1st Dec. 1958, Vol. 72, No. 468, pp. 1144–1146.) The dielectric-hysteresis loops of barium metatitanate ceramics become constricted with age and eventually form a narrow double loop. The aging process may be reversed by heating above the Curie point.

537.228.1 **845**

Some Piezoelectric Properties of Polycrystalline Solid Solutions $(Ba,Sr)TiO_3$, $Ba(Ti,Sn)O_3$ & $Ba(Ti,Zr)O_3$.—V. A. Bokov. (*Akust. Zh.*, April/June 1957, Vol. 3, No. 2, pp. 104–108.) In certain solid solutions the electromechanical response and effective piezoelectric modulus depend on the voltage of the polarizing field and temperature. This response and coefficient pass through a maximum, due to domain orientation, particularly in solid solutions of the type $Ba(Ti,Sn)O_3$ and $Ba(Ti,Zr)O_3$.

537.228.1 : 546.472.21 **846**

Theory of the Piezoelectric Effect in the Zincblende Structure.—J. L. Birman. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1510–1514.) A theory is developed which leads to an equation relating the macroscopic piezoelectric constant to the static and dynamic effective charges, and a lattice parameter which relates the internal strain to the external strain. An order-of-magnitude check of the theory is made on ZnS.

537.311.33 **847**

The Prediction of Semiconducting Properties in Inorganic Compounds.—C. H. L. Goodman. (*J. Phys. Chem. Solids*, Sept. 1958, Vol. 6, No. 4, pp. 305–314.) Various criteria are presented for predicting possible semiconductor behaviour in inorganic compounds, based on the requirements of saturated ionic-covalent bonding. It is shown that new series of semiconducting compounds can be derived from known ones by replacing one element by pairs from other groups of the periodic table, while keeping the valence-electron : atom ratio constant. These concepts are illustrated with particular reference to diamond-type lattices.

537.311.33 **848**

Calculations on the Shape and Extent of Space-Charge Regions in Semiconductor Surfaces.—G. C. Dousmanis & R. C. Duncan, Jr. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1627–1629.) Curves showing the potential as a function of distance inside a semiconductor are obtained by numerical integration of the Poisson equation. The results apply to all semiconductors.

537.311.33 **849**

The Dependence of Minority-Carrier Lifetime on Majority-Carrier Density.—D. M. Evans. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1962–1963.) The minority-carrier lifetime in Ge and Si is inversely proportional to the majority-carrier density only if the recombination levels lie near the appropriate band edge and/or the semiconductor is only weakly extrinsic.

537.311.33 **850**

Influence of Crystal Lattice Vibrations on the Production of Electron-Hole Pairs in a Strong Electric Field.—L. V. Keldysh. (*Zh. eksp. teor. Fiz.*, April 1958, Vol. 34, No. 4, pp. 962–968.) The probability for the production of an electron-hole pair in a semiconductor is calculated considering electron-phonon interaction. Other processes which may influence the diffusion of a valence electron into the conduction band are considered, in particular, the absorption of several phonons which may be decisive in relatively weak fields.

537.311.33 **851**

Semiconductor Surface Potential and Surface States from Field-Induced Changes in Surface Recombination.—G. C. Dousmanis. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 369–380.) The variations in surface recombination are detected by changes in the reverse current of large-area 'back-surface' diodes. Observation of a maximum of surface recombination in terms of applied field provides a reference point from which the zero-field value of the surface potential can be evaluated, and the dependence of the surface recombination velocity on the surface potential can be established.

537.311.33 **852**

Junction Capacitance and Related Characteristics using Graded Impurity Semiconductors.—L. J. Giacoletto. (*Trans. Inst. Radio Engrs*, July 1957, Vol. ED-4, No. 3, pp. 207–215. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1760.)

537.311.33 **853**

The *K*-Edge Structures of the Elements of Various $A_{III}B_V$ Compounds.—W. Eberbeck. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 412–424.) The fine structure of the *K* X-ray absorption spectrum was investigated for Ga in GaP, GaAs, and GaSb, for As in GaAs and InAs, for Zn in ZnS, and for pure Ge.

537.311.33 : 537.32 **854**

The Effect of Strain on the Thermoelectric Properties of a Many-Valley Semiconductor.—J. R. Drabble. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 362–372.) A general treatment is given for nondegenerate materials. It is shown that the absence of effects other than those due to carrier transfer between valleys can be checked by the condition that the trace of the conductivity tensor is unaltered by the strain. Under these conditions the conductivity and thermoelectric power tensors can be combined to give an expression involving only the relative shifts in the minimum energies of the valleys.

537.311.33 : 538.632 **855**

Hall Effect in Semiconductor Compounds.—M. J. O. Strutt. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, pp. 2–10.) The effect in InAs and InSb, and the influence of probe shape and position are discussed. Applications to several types of wattmeter, to oscillators, to a flux-density meter and to a receiver mixer stage are described.

537.311.33 : 538.632 **856**

Hall Mobility of Carriers in Impure Nondegenerate Semiconductors.—M. S. Sodha & P. C. Eastman. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, p. 44.) The case of low temperatures and high impurity concentrations is considered.

537.311.33 : [546.28+546.289 **857**

Lattice Vibrational Spectra of Si and Ge.—H. Cole & E. Kineke. (*Phys. Rev. Lett.*, 15th Nov. 1958, Vol. 1, No. 10, pp. 360–361.)

537.311.33 : [546.28+546.289 **858**

Theoretical Surface Conductivity Changes and Space Charge in Germanium and Silicon.—V. O. Mowery. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1753–1757.) Graphs are given showing the surface conductivity and space charge as functions of resistivity for various values of surface potential.

537.311.33 : 546.28 **859**

The Measurement of Surface Recombination Velocity on Silicon.—A. H. Benny & F. D. Morten. (*Proc. phys. Soc.*, 1st Dec. 1958, Vol. 72, No. 468, pp. 1007–1012.) The surface recombination velocity on a Si filament can be determined by the measurement of the spectral distribution of the photoconductivity. On *n*-type Si the surface recombination velocity can be varied from less than 75 to greater than 7 500 cm.sec^{-1} by changing the ambient gas.

537.311.33 : 546.28 **860**

Microwave Spin Echoes from Donor Electrons in Silicon.—J. P. Gordon & K. D. Bowers. (*Phys. Rev. Lett.*, 15th Nov. 1958, Vol. 1, No. 10, pp. 368–370.) The relaxation time in Si containing Li and P has been measured at a frequency of 23 kMc/s using a heterodyne paramagnetic spectrometer. Results indicate that a spin-echo device could provide a storage system in which each element could store more than 10^4 bits of information.

537.311.33 : 546.28 **861**

Neutron-Bombardment Damage in Silicon.—G. K. Wertheim. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1500–1505.) Neutron-bombardment damage in Si is compared to electron-bombardment effects which have been previously analysed (3515 of 1958). A discrete energy level 0·27 eV above the valence band is produced by both neutrons and electrons. A spectrum of energy levels running from 0·16 eV below the conduction band toward the middle of the gap is ascribed to a defect pair with variable spacing.

537.311.33 : 546.28 **862**

Recombination Properties of Gold in Silicon.—G. Bemski. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1515–1518.) The capture of electrons in *p*-type Si occurs through the Au donor level with a capture cross-section of $3{\cdot}5 \times 10^{-15}$cm^2 at 300°K; this capture cross-section varies as $T^{-2.5}$ between 200° and 500°K. In *n*-type Si the electron capture cross-section is 5×10^{-16}cm^2 and is independent of temperature; the hole capture cross-section is 1×10^{-15}cm^2 at 300°K and varies as T^{-4}. The capture in this case occurs through the Au acceptor level.

537.311.33 : 546.28 : 535.215 **863**

Quantum Yield of Photoionization in Silicon.—V. S. Vavilov & K. I. Britsyn. (*Zh. eksp. teor. Fiz.*, May 1958, Vol. 34, No. 5, pp. 1354–1355.) Results of an investigation of the photo-effect in *p-n* junctions obtained by thermal diffusion of phosphorus indicate an increase in quantum yield and the presence of impact ionization by liberated carriers.

537.311.33 : [546.28 + 546.289] : 535.39-15 **864**

Antireflection Coatings for Germanium and Silicon in the Infrared.—J. T. Cox & G. Hass. (*J. opt. Soc. Amer.*, Oct. 1958, Vol. 48, No. 10, pp. 677–680.)

537.311.33 : 546.289 **865**

Production of Dislocations in Germanium by Thermal Shock.—R. S. Wagner. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1679–1682.) The results indicate that the imperfections are not generated during growth, but occur afterwards, as a result of thermal shock when the growth is terminated.

537.311.33 : 546.289 **866**

Effect of Monoenergetic Fast Neutrons on *n*-Type Germanium.—S. L.

Ruby, F. D. Schupp & E. D. Wolley. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1493–1496.) The density of vacancy-interstitial pairs produced per unit of fast-neutron flux on *n*-type Ge has been measured experimentally using monoenergetic neutrons. The observed changes in resistivity and Hall coefficients indicate charge carrier removal rates per unit of neutron flux much lower than that calculated from current models.

537.311.33 : 546.289 **867**

Photoconductive Response of Single-Crystal Germanium Layers prepared by the Pyrolytic Decomposition of GeI_2.—D. C. Cronemeyer. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1730–1735.) The photo-response measured at 77°K was found to extend to about 6 μ, in agreement with the thermal activation energy. Doping of the Ge with Au and Ag was attempted but no positive identification of the doping agents was obtained from the photoconductive response.

537.311.33 : 546.289 **868**

Radiation-Induced Recombination Centres in Germanium.—O. L. Curtis, Jr, J. W. Cleland & J. H. Crawford, Jr. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1722–1729.) The effect of energetic particle bombardment on the minority carrier lifetime has been measured. It is concluded that a defect state located 0·20 eV below the conduction band dominates the recombination process.

537.311.33 : 546.289 **869**

Radiative Recombination in Germanium.—P. H. Brill & R. F. Schwarz. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 330–333.) Dependence on excess-carrier density and on equilibrium-carrier density was studied by simultaneous measurements of output radiation and of photoconductivity as functions of incident light intensity. Results confirm the theory.

537.311.33 : 546.289 **870**

Magnetoresistance in *n*-Type Germanium at Low Temperatures.—R. A. Laff & H. Y. Fan. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 317–321.) The effective anisotropy parameter K decreased from $\approx$ 20 at 300°K to $\approx$ 6 at 20°K but values near 20 were again obtained at 7°K and 4·2°K. Introduction of compensating acceptors showed that the decrease of K was due to anisotropic scattering by ionized impurities. Below 7°K, scattering was controlled by neutral impurities with essentially isotropic relaxation time.

537.311.33 : 546.289 **871**

Infrared Absorption by Conduction Electrons in Germanium.—H. J. G. Meyer. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 298–308.) A theory is developed taking into account the multi-valley structure of the conduction band and present knowledge of the scattering mechanism. Explicit calculations are possible at all relevant wavelengths and temperatures.

537.311.33 : 546.289 **872**

Oscillation of the Electrical Resistance of *n*-Type Germanium in Strong Pulsed Magnetic Fields.—I. G. Fakidov & E. A. Zavadskiĭ. (*Zh. eksp. teor. Fiz.*, April 1958, Vol. 34, No. 4, pp. 1036–1037.) Note on measurements of the resistance variation in single-crystal Ge in a transverse pulsed magnetic field up to 120 kG at temperatures 300, 77 and 20°K.

537.311.33 : 546.289 **873**

Surface Effects in Electron-Irradiated Ge at 80°K.—W. E. Spear. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 362–369.) Results are given of photoconductivity, surface-conductance, Hall- and field-effect measurements made before and after electron irradiation. Use of both low- and high-energy irradiation distinguishes between volume and surface effects. Surface rather than volume effects are responsible for photoconductivity changes beyond the fundamental absorption edge.

537.311.33 : 546.289 : 535.215 **874**

Delayed Electron Emission and External Photo-effect of Germanium after Electron Bombardment. — K. Seeger. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 453–470.) Emission was investigated in the temperature range 103°–670°K. An interpretation of the effect as being due to slow surface states is discussed.

537.311.33 : 546.289 : 538.569 **875**

Experimental Evidence for Carriers with Negative Mass.—G. C. Dousmanis, R. C. Duncan, Jr, J. J. Thomas & R. C. Williams. (*Phys. Rev. Lett.*, 1st Dec. 1958, Vol. 1, No. 11, pp. 404–407.) Cyclotron resonance experiments have been carried out on Ge crystals at 4°K. A spectrum of heavy holes has been revealed which may be assigned to carriers of negative mass.

537.311.33 : 546.289.231 **876**

The Crystal Structure of Germanium Selenide GeSe.—A. Okazaki. (*J. phys. Soc. Japan*, Oct. 1958, Vol. 13, No. 10, pp. 1151–1155.)

537.311.33 : 546.47-31 **877**

An Approach to Intrinsic Zinc Oxide.—W. Ruppel, H. J. Gerritsen & A. Rose. (*Helv. phys. Acta*, 20th Nov. 1957, Vol. 30, No. 6, pp. 495–503.) Measurements have been made on layers of finely divided ZnO powder the resistivity of which (10^{17} Ω.cm) is closer to that of intrinsic ZnO than that of single crystals. The formation of depletion layers and the rectifying action observed with ohmic and blocking contacts are discussed.

537.311.33 : 546.47-31 **878**

The Preparation of Zinc Oxide Single Crystals with Defined Impurity Content.—G. Bogner & E. Mollwo. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 136–143. In German.) A method is described by which ZnO crystals of a definite composition can be deposited from the gas phase. Measurements of the composition, conductivity, and infrared absorption of crystals doped with Cu or In are reported.

537.311.33 : 546.47-31 **879**

The Concentration and Mobility of Electrons in Zinc Oxide Single Crystals with Defined Impurity Content.—H. Rupprecht. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 144–154 In German.) The conductivity and Hall effect in synthetic crystals of doped ZnO have been measured in the temperature range 65–700°K. Results are similar to those for other semiconductors such as Ge and Si.

537.311.33 : 546.47-31 **880**

Field Effect and Photoconductivity in ZnO Single Crystals.—G. Heiland. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 155–168.) Report and discussion of the results of measurements made on ZnO crystals with surface conductivity varying over a wide range.

537.311.33 : 546.681.19 **881**

The Preparation and Properties of Gallium Arsenide Single Crystals.—J. M. Whelan & G. H. Wheatley. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 169–172.) Apparatus is described for the preparation and purification of GaAs by horizontal zone refining, and for the production of single crystals by the floating-zone method. Resistivities up to 10^6 Ω.cm have been obtained. Hall coefficient and conductivity data for *n*-type samples are given. Electron mobility, the energy gap, and the position of an impurity level are deduced.

537.311.33 : 546.681.19 **882**

Diffusion, Solubility, and Electrical Behaviour of Copper in Gallium Arsenide.—C. S. Fuller & J. M. Whelan. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp, 173–177.)

537.311.33 : 546.682.18 **883**

Optical Properties of *n*-Type InP.—R. Newman. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1518–1521.)

537.311.33 : 546.682.19 : 537.312.9 **884**

Piezoresistance Constants of *n*-Type InAs.—A. J. Tuzzolino. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, p. 30.) Results of measurements from 77°K to 300°K are consistent with a spherical conduction-band model.

537.311.33 : 546.72.23 **885**

$FeSe_2$, a Semiconductor containing Iron.—G. Fischer. (*Canad. J. Phys.*, Oct. 1958, Vol. 36, No. 10, pp. 1435–1438.) The preparation of this compound is described; the results of electrical-resistivity and Hall-coefficient measurements and X-ray examination are summarized.

537.311.33 : 546.812.221 **886**

The Semiconductor Properties of Synthetic Herzenbergite (SnS) Crystals.—H. Gobrecht & A. Bartschat. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 511–522.)

537.311.33 : 546.824-31 : 537.312.9 **887**

Piezoresistivity in Reduced Single-Crystal Rutile (TiO_2).—L. E. Hollander, Jr. (*Phys. Rev. Lett.*, 15th Nov. 1958, Vol. 1, No. 10, pp. 370–371.) The effect of stress on resistivity has been measured at different temperatures.

537.311.33 : 621.314.63 **888**

Alloy Junctions in Semiconducting Devices.—D. F. Taylor. (*Research, Lond.*, Sept. 1958, Vol. 11, No. 9, pp. 335–338.) Outline of theory on which methods of preparing alloy junctions are based.

537.311.33 : 621.318.57 **889**

Current Build-Up in Semiconductor Devices.—W. Shockley & J. Gibbons. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1947–1949.) Assuming that the minority-carrier densities and their associated currents have an exponential rise during the switching action, a solution to the partial differential equations can be found from which the efficiencies of various designs of switching devices may be derived.

538.22 **890**

An Interpretation of the Magnetic Properties of the Perovskite-Type Mixed Crystals La_{1-x} Sr_x $CoO_{3-\lambda}$.—J. B. Goodenough. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 287–297.)

538.221 **891**

Interatomic Distances in Ferromagnetics.—F. M. Gal'perin. (*Zh. eksp. teor. Fiz.*, April 1958, Vol. 34, No. 4, pp. 1000–1003.)

538.221 **892**

The Change of Spontaneous Magnetization with Hydrostatic Pressure.—D. Gugan. (*Proc. phys. Soc.*, 1st Dec. 1958, Vol. 72, No. 468, pp. 1013–1026.)

538.221 **893**

Dipolar Energy. Application to the Magnetostatic Energy of α-Fe_2O_3.—F. Bertaut. (*C. R. Acad. Sci., Paris*, 16th June 1958, Vol. 246, No. 24, pp. 3335–3337.) The magnetic anisotropy and weak ferromagnetism of α-Fe_2O_3 cannot be explained in terms of dipolar energy.

538.221 **894**

Sendust Flake—a New Magnetic Material for Low-Frequency Application.—W. M. Hubbard, E. Adams & J. F. Haben. (*Trans. Inst. Radio Engrs*, March 1957, Vol. CP-4, No. 1, pp. 2–6. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1035.)

538.221 : 539.231 **895**

Magnetic, Electrical and Electron-Optical Investigations of the Thermal Transformation of Cathode-Sputtered Nickel Films.—L. Reimer. (*Z. Phys.*, 31st Oct. 1957, Vol. 149, No. 4, pp. 425–431.)

538.221 : 621.3.042.2 **896**

Cube-Oriented Magnetic Sheet.—(*J. Metals, N.Y.*, Aug. 1958, Vol. 10, No. 8, pp. 507–511.)

A Major Advance in Magnetic Materials.—G. W. Wiener & K. Detert (pp. 507–508).

Magnetic Properties of Cube-Textured Transformer Sheet.—J. L. Walter, W. R. Hibberd, Jr, H. C. Fiedler, H. E. Grenoble, R. H. Pry & P. G. Frischmann (pp. 509–511).

538.221 : [621.318.124 + 621.318.134 **897**

Data on Ferrite Core Materials.—A. C. Hudson & E. J. Stevens. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 718–719.) Charts are presented which permit comparison of the permeability and loss at low levels for various ferrites over the frequency range 100 kc/s–100 Mc/s.

538.221 : [621.318.124 + 621.318.134 **898**

Effect of Neutron Irradiation on the Curie Temperature of a Variety of Ferrites.—E. I. Salkovitz, G. C. Bailey & A. I. Schindler. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1747–1748.) No significant change in Curie temperature after irradiation was found for either magnetically soft or hard ferrites.

538.221 : 621.318.124 **899**

Magnetization Studies and Possible Magnetic Structure of Barium Ferrate III.—W. E. Henry. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 326–327.)

538.221 : 621.318.134 **900**

Modified Rotational Model of Flux Reversal.—E. M. Gyorgy. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1709–1712.) The mechanism of flux reversal in square-loop ferrites is analysed showing that the flux in a toroid can be reversed by rotation without prohibitively large demagnetizing fields. This agrees with observations. See also 534 of 1958.

538.221 : 621.318.134 **901**

Magnetic Anisotropy Constant of Yttrium-Iron Garnet at 0°K.—B. R. Cooper. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 395–396.)

538.221 : 621.318.134 **902**

Low-Temperature Transition of Magnetic Anisotropy in Nickel-Iron Ferrite.—N. Menyuk & K. Dwight. (*Phys. Rev.*, 15th Oct. 1958, Vol. 112, No. 2, pp. 397–405.) A study of the magnetic properties of single-crystal samples of Ni-Fe ferrite has revealed an abrupt transition in the magnetic anisotropy characteristics at 10°K. Predictions based on a proposed model are in accord with experimental findings.

538.221 : 621.318.134 **903**

An Investigation into the Magnesium and Magnesium - Manganese Ferrite System.—L. C. F. Blackman. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 373–384.) "The influence of both manganese and firing conditions on certain physical and chemical properties of iron-deficient magnesium ferrite has been investigated. The results suggest that Fe^{2+} and Mn^{3+} can coexist, at least to a certain extent, in the solid state. It is also shown that the excess of magnesia, which is added to improve the microwave performance of the materials, plays an important role in the chemistry of the ferrite phase."

538.221 : 621.318.134 **904**

Ferrimagnetism in the System Na_2O-ZnO-Fe_2O_3.—A. H. Mones & E. Banks. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 267–270.)

538.221 : 621.318.134 **905**

Origin of Weak Ferromagnetism in Rare-Earth Orthoferrites.—R. M. Bozorth. (*Phys. Rev. Lett.*, 15th Nov. 1958, Vol. 1, No. 10, pp. 362–363.)

538.221 : 621.318.134 : 538.569.4 **906**

Ferromagnetic Resonance in Polycrystalline Ferrites with Large Anisotropy: Part 1—General Theory and Application to Cubic Materials with a Negative Anisotropy Constant.—E. Schlömann. (*J. Phys. Chem. Solids*, Aug. 1958, Vol. 6, Nos. 2/3, pp. 257–266.)

538.221 : 621.318.134 : 538.569.4 **907**

Ferrimagnetic Resonance in Yttrium-Iron Garnet at Liquid Helium Temperatures.—J. F. Dillon, Jr. (*Phys. Rev.*, 15th Sept. 1958, Vol. 111, No. 6, pp. 1476–1478.) Experiments performed on single-crystal spheres at 24 kMc/s are described. The paper gives in some detail the low-temperature results on line width and the variation of the resonance field with crystal direction and temperature.

538.221 : 621.318.134 : 538.569.4 **908**

Magnetostatic Modes in Ferrimagnetic Spheres.—J. F. Dillon, Jr. (*Phys. Rev.*, 1st Oct. 1958, Vol. 112, No. 1, pp. 59–63.) Experiments on Y-Fe garnet show that the positions of the modes are not functions of crystal direction as previously reported. Specimens produced by tumbling procedures may deviate slightly from sphericity and in these experiments truly spherical samples were used. Evidence has been given of the effect of dielectric inhomogeneities in the neighbourhood of the sample in a ferrimagnetic-resonance experiment.

538.222 : 538.569.4 **909**

Paramagnetic-Resonance Spectrum of Gadolinium in Single Crystals of Thorium Oxide.—W. Low & D. Shaltiel. (*J. Phys. Chem. Solids*, Sept. 1958, Vol. 6, No. 4, pp. 315–323.) Measurements at 290° and 90°K at 3 cm λ give line positions and intensities which can be explained by a crystalline field of cubic symmetry. It is suggested that oxygen vacancies are randomly distributed throughout the crystal.

539.2 : 537.311.31 **910**

Electronic Band Structures of the Alkali Metals and of the Noble Metals and their α-Phase Alloys.—M. H. Cohen & V. Heine. (*Advances Phys.*, Oct. 1958, Vol. 7, No. 28, pp. 395–434.)

621.315.616.96 **911**

Electrical Properties of Epoxy Resins.—C. F. Pitt, B. P. Barth & B. E. Godard. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CP-4, No. 4, pp. 110–113. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 801.)

MATHEMATICS

517.942 : 538.566 **912**

Method of Liouville Applied to Weber's Equation.—R. Meynieux. (*C. R. Acad. Sci., Paris*, 9th June 1958, Vol. 246, No. 23, pp. 3208–3210.) Applications include the propagation of a plane e.m. wave across an ionospheric layer.

517.942 : 538.566 **913**

Gamma-Function Approximations Applied to Solutions of Weber's Equation.—R. Meynieux. (*C. R. Acad. Sci., Paris*, 16th June 1958, Vol. 246, No. 24, pp. 3312–3314.)

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) : 529.786 : 525.35 **914**

Comparison of Astronomical Time Measurements with Atomic Frequency Standards.—J. P. Blaser & J. De Prins. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, pp. 859–860.) Comparisons have been made in Neuchâtel between 'atomic' time determined by integration of the frequency of a quartz clock calibrated against an NH_3 maser, and astronomical time determined by means of photographic zenith tube and astrolabe. Plotting U.T.1 against atomic time shows the variation in the rate of rotation of the earth, which is in agreement with the measurements of Essen et al. (3195 of 1958).

621.3.018.41(083.74) : 621.396.11 **915**

Comparison of an Ammonia Maser with a Caesium Atomic Frequency Standard.—J. P. Blaser & J. Bonanomi. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, p. 859.) Note of a comparison made in Neuchâtel by means of MSF standard-frequency transmissions during 1957 and 1958 of the frequencies of an NH_3 maser and the Cs resonator of the National Physical Laboratory. Two different methods were used: (*a*) differentiation of the phase of MSF time signals at 5 and 10 Mc/s; (*b*) direct frequency comparison using the 60-kc/s transmission. See also 1208 of 1958 (Essen et al.).

621.3.018.41(083.74) : 621.396.11 : 551.510.535 **916**

Frequency Variations in Short-Wave Propagation.—T. Ogawa. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1934–1939.) Frequency measuring equipment, with an accuracy within 1×10^{-8} is described. Observations were made on 5- and 10-Mc/s standard-frequency transmissions from station JJY and the results are discussed in relation to ionospheric conditions.

621.317.1.029.6 **917**

Reciprocity in Radio-Frequency Measurements.—G. D. Monteath. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, pp. 18–20.) A discussion of the advantages of interchangeability of source and detector in measurements of radiation patterns and impedance.

621.317.3 : 538.632 **918**

Alternate Current Apparatus for Measuring the Ordinary Hall Coefficient of Ferromagnetic Metals and Semiconductors.—J. M. Lavine. (*Rev. sci. Instrum.*, Nov. 1958, Vol. 29, No. 11, pp. 970–976.) Apparatus for measurements at 1 000 c/s is described having a sensitivity of 10^{-18} W, a noise level of 10^{-9} V and a voltage resolution of 1 in 10^5 for use with sample impedances ranging from less than 1 Ω to several thousand ohms.

621.317.3.018.7 : 534.78 **919**

A Sampling Comparator.—A. Fischmann-Arbel. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 685–689.) The instrument described compares the instantaneous values of two waveforms at constant intervals of time. Applications of the comparator circuit in a delta modulator and a binary quantizer are detailed.

621.317.326 **920**

Measurement of the Peak Value of a High-Voltage Pulse.—G. Giralt. (*C. R. Acad. Sci., Paris*, 9th June 1958, Vol. 246, No. 23, pp. 3227–3229.) Two methods involving the use of a ballistic galvanometer are described: one in which the rectified current through a standard capacitor is measured [see 1851 of 1957 (Lagasse & Giralt)], the other a blocking method, using a single half-wave rectifier and a circuit-breaking device, applicable for pulses of any waveform.

621.317.335.3.029.6 : 621.317.733 **921**

Bridge Method for Microwave Dielectric Measurements.—S. H. Glarum. (*Rev. sci. Instrum.*, Nov. 1958, Vol. 29, No. 11, pp. 1016–1019.) The length of a short-circuited section of liquid-dielectric-filled coaxial line, is adjusted until its input admittance, measured using a microwave bridge, is purely resistive. The dielectric constant, between 1 and 3 kMc/s, obtained by this method compares well with published values.

621.317.335.3.029.64 **922**

The Measurement of the Dielectric Properties of Liquids in an H_{01} Resonator.—J. S. Dryden. (*J. sci. Instrum.*, Dec. 1958, Vol. 35, No. 12, pp. 439–440.) A thin-walled silica cup with a flat metal disk at the bottom has proved suitable as a sample container. A method of correction for the meniscus is described.

621.317.34 + 621.317.38] .029.62 **923**

Triple V.H.F. Reflectometer.—G. H. Millard. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, pp. 11–13.) The instrument is designed for use in the frequency bands 41–68 Mc/s and 88–95 Mc/s for powers of 1 kW or less.

621.317.39 : 531.76 **924**

A Catapult End-Speed Recorder.—J. R. Pollard. (*Brit. Commun. Electronics*, Sept. 1958, Vol. 5, No. 9, pp. 676–680.) A description is given of electronic equipment for indicating and recording in printed form the launch speed attained by a steam catapult.

621.317.733 : 621.314.7 **925**

A Wide-Band Bridge Yielding Directly the Device Parameters of Junction Transistors.—J. Zawels. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 21–25. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 932.)

621.317.75 : 621.372.413 **926**

Microwave Double-Sweep Method for Analysis of Time-Dependent Cavity Characteristics.—S. Ruthberg. (*Rev. sci. Instrum.*, Nov. 1958, Vol. 29, No. 11, pp. 999–1003.) A frequency-sweep f.m. search signal and a frequency-sweep receiver are used to obtain a pulse whose shape is sensitive to cavity resonance. Frequency shift is determined from pulse shape and position along the received signal trace.

621.317.755 : 621.385.832 **927**

Oscilloscope Tube with Travelling-Wave Deflection System and Large Field of View.—W. F. Niklas & J. Wimpffen. (*J. Brit. Instn Radio Engrs*, Nov. 1958, Vol. 18, No. 11, pp. 653–660.) A tube is described having a balanced helix system as deflection plates and with a large viewing field ($1\frac{3}{8}$ in. × 4 in.). Typical operating conditions and performance are described and the relative merits of magnetic and e.s. focusing in a high-speed oscilloscope are discussed in detail.

621.317.799 : 621.314.7 **928**

Sweep Equipment Displays Transistor Beta.—R. Zuleeg & J. Lindmayer. (*Electronics*, 5th Dec. 1958, Vol. 31, No. 49, pp. 100–101.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

612.1 : 621.396.62-519 **929**

Radio Control of Ventricular Contraction in Experimental Heart Block.—M. Verzeano, R. C. Webb, Jr, & M. Kelly. (*Science*, 24th Oct. 1958, Vol. 128, No. 3330, pp. 1003–1005.) A method is described for the stimulation of the ventricular myocardium by transmitting a pulse-modulated 2·5-Mc/s carrier which is detected by a receiver enclosed in the chest of the animal under test.

612.84 : 621.397.5 **930**

Eye Fixations Recorded on Changing Visual Scenes by the Television Eye-Marker.—J. F. Mackworth & N. H. Mackworth. (*J. opt. Soc. Amer.*, July 1958, Vol. 48, No. 7, pp. 439–445.) The corneal reflection of a light is picked up by a television camera and is superimposed upon a television monitor displaying the scene which the subject views on a separate screen. The resulting light spot can be made to lie accurately, within one or two degrees, on the part of the scene being regarded by the subject.

621.3.087.5 **931**

High-Fidelity Video Recording using Ultrasonic Light Modulation.—L. Levi. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1958, Vol. 67, No. 10, pp. 657–661. Discussion.) A method is described for recording video information with bandwidths up to 20 Mc/s on photographic film, using a piezoelectric ultrasonic transducer.

621.3.087.9 : 621.395.625.3 **932**

Tones find Data in High-Speed Tape Systems.—R. Wasserman & P. Hurney. (*Electronics*, 21st Nov. 1958, Vol. 31, No. 47, pp. 92–95.) A digital timing generator, operating during recording, and an associated search unit enable selected data to be extracted from multichannel magnetic-tape systems.

621.365.5 : 621.316.726.078.3 **933**

Frequency Stability of R.F. Heating Generators.—J. Verstraten. (*Electronic Applic.*, Aug. 1958, Vol. 18, No. 3, pp. 122–128.) A cavity resonator of large dimensions is used as a tank circuit. Details are given of a 2-kW generator whose frequency remains constant to within 0·5 % of 27·12 Mc/s.

621.384.62 **934**

New Design makes 10-MeV Particle Accelerator Possible.—J. L. Danforth. (*Canad. Electronics Engng*, July 1958, Vol. 2, No. 7, pp. 18–21.) See 935 below.

621.384.62 **935**

10-MeV Particle Accelerator will Aid Nuclear Research at Chalk River.—H. E. Gove. (*Canad. Electronics Engng*, July 1958, Vol. 2, No. 7, pp. 14–17.) A general description of the tandem Van de Graaff accelerator and associated equipment.

621.385.833 **936**

Investigations of the Remote-Focus Cathode of Steigerwald.—F. W. Braucks. (*Optik., Stuttgart*, April 1958, Vol. 15, No. 4, pp. 242–260.) Control voltage, and aperture, intensity, position and diameter of the smallest cross-section of the beam are discussed in relation to the design parameters of the system.

621.385.833 **937**

System Design of Asymmetric Unipotential Electron Lenses.—K. J. Hanssen. (*Optik, Stuttgart*, May 1958, Vol. 15, No. 5, pp. 304–317.) See also 1558 of 1957 (Everitt & Hanssen).

621.385.833 **938**

Shadow-Casting Carbon Films used in Electron Microscopy.—A. Oberlin & C. Tchoubar. (*C. R. Acad. Sci., Paris*, 16th June 1958, Vol. 246, No. 24, pp. 3329–3332.) Carbon films are shown to be preferable to metal for shadow techniques as they have less granulation.

PROPAGATION OF WAVES

621.396.11 + 621.372.2 **939**

Transmission and Reflection of Electromagnetic Waves in the Presence of Stratified Media.—J. R. Wait. (*J. Res. nat. Bur. Stand.*, Sept. 1958, Vol. 61, No. 3, RP 2899, pp. 205–232.) "A general analysis is presented for the electromagnetic response of a plane stratified medium consisting of any number of parallel homogeneous layers. The solution is first developed for plane-wave incidence and then generalized to both cylindrical and spherical-wave incidence. Numerical results for interesting special cases are presented and discussed. The application of the results to surface-wave propagation over a stratified ground is considered in some detail."

621.396.11 **940**

An Example of Guided Propagation in the Mediterranean.—L. Boithias & P. Misme. (*Ann. Télécommun.*, April 1957, Vol. 12, No. 4, pp. 126–132.) Results are analysed of field-strength measurements at 10 cm λ made on board ship, at distances up to 180 nautical miles from the transmitter, together with meteorological data provided by a tethered radiosonde. Three types of ducting are suggested.

621.396.11 : 551.510.535 **941**

Variations in the Direction of Arrival of High-Frequency Radio Waves.—J. E. Titheridge. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 17–25.) The effect of propagation in a non-horizontally uniform ionosphere is investigated. The equations are solved for linear and parabolic layers and used to calculate layer tilts necessary to produce the large diurnal bearing changes in the arrival of short waves at Auckland, New Zealand.

621.396.11 : 551.510.535 **942**

Very-Long-Distance Ionospheric Propagation.—N. C. Gerson. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 169–172.) A brief discussion on the effectiveness of propagation in the spherical shell between the earth and the ionosphere, and also magnetic (whistler-mode) propagation.

621.396.11 : 551.510.535 **943**

Long-Distance Single-F-Hop Transmission.—W. Dieminger, H. G. Möller & G. Rose. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 191–192.)

621.396.11 : 551.510.535 **944**

Magneto-ionic Fading in Pulsed Radio Waves Reflected at Vertical Incidence from the Ionosphere.—C. A. Reddy, B. R. Rao & M. S. Rao. (*J. Brit. Instn Radio Engrs*, Nov. 1958, Vol. 18, No. 11, pp. 669–675.) The difference in phase paths of the two interfering magneto-ionic components is calculated on the basis of ray theory assuming a parabolic electron-density distribution for the F_2 region. The calculated fading frequencies agree fairly well with the observed values. A method of deducing the semi-thickness of the F_2 region from these fading frequencies is described and some results are given.

621.396.11 : 551.510.535 **945**

A New Type of Fading Observable on High-Frequency Radio Transmissions Propagated over Paths Crossing the Magnetic Equator.—K. C. Yeh & O. G. Villard, Jr. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1968–1970.) The received energy is split into two independently fading components of comparable strength, separated in frequency by some tens of c/s. It is suggested that the phenomenon is caused by a combination of conventional and tilt-supported propagation across the evening equatorial height bulge in the F layer of the ionosphere.

621.396.11 : 551.510.535 : 621.3.018.41 (083.74) **946**

Frequency Variations in Short-Wave Propagation.—Ogawa. (See 916.)

621.396.11 : 621.396.812 **947**

Some Relations between the Bearing and Amplitude of a Fading Radio Wave.—H. A. Whale & L. M. Delves. (*J. atmos. terr. Phys.*, Dec. 1958, Vol. 13, Nos. 1/2, pp. 72–85.) A correlation ratio connecting the changes in amplitude and bearing of a fading wave made up of several randomly phased components is derived. Experimental and theoretical conclusions have been compared.

621.396.11.029.45 **948**

Propagation of Very-Low-Frequency Pulses to Great Distances.—J. R. Wait. (*J. Res. nat. Bur. Stand.*, Sept. 1958, Vol. 61, No. 3, RP 2898, pp. 187–203.) The space between the earth and the ionosphere is represented as a sharply bounded waveguide with concentric spherical boundaries. The concept of phase and group velocity is discussed and applied to determine the influence of the propagation medium on the shape of the envelope of a quasi-monochromatic pulse. An alternative method is described that is applicable to wide-band sources containing many spectral components.

RECEPTION

621.372.632.029.6 **949**

Design Considerations in a Wide-Band Microwave Mixer and I.F. Preamplifier.—J. C. Rennie. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CS-5, No. 2, pp. 21–25.)

621.396.621.22 : 621.375.121.2 **950**

An Electronic Multicoupler and Antenna Amplifier for the V.H.F. Range.—K. Fischer. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CS-5, No. 3, pp. 43–48. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1958, Vol. 46, No. 2, p. 515.)

621.396.66 : 621.316.726.078.3 **951**

A.F.C. in Band-II F.M. Receivers: using a Junction Diode.—G. D. Browne. (*Mullard tech. Commun.*, Nov. 1958, Vol. 4, No. 35, pp. 152–157.) The reactance variation obtained by varying the Ge-diode reverse voltage provides a frequency control operating successfully above 10 Mc/s.

621.396.666 **952**

Laboratory Tests on Kahn's Theory of Anti-fading Reception.—G. Bronzi, (*Alta Frequenza*, Feb. 1958, Vol. 27, No. 1, pp. 17–43.) Analysis of an experimental comparison of methods of double-diversity reception using strongest-signal selection or the ratio-squarer combining system [541 of 1955 (Kahn)] shows the advantages of the latter method.

621.396.81 **953**

Correlation Measurements in the Short-Wave Range.—J. Grosskopf, M. Scholz & K. Vogt. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 91–95.) Report on measurements of the correlation coefficient in diversity reception using rhombic and dipole aerials.

621.396.82 **954**

Radio Interference: Part 5—Industrial, Scientific and Medical Apparatus and Radiating Receivers.—C. W. Sowton & G. A. C. R. Britton. (*P.O. elect. Engrs' J.*, Oct. 1958, Vol. 51, Part 3, pp. 202–205.) International frequency allocations for industrial, scientific and medical apparatus are discussed and examples of the field strength of radiation at fundamental and harmonic frequencies are tabulated. The free-radiation frequency of 27·12 Mc/s with harmonic attenuation has been adopted for medical apparatus. Methods for measuring the radiation fields of receivers are outlined. These methods have been accepted by most countries. Part 4: 3964 of 1958 (Macpherson).

621.396.822 : 621.397.62 **955**

Design Considerations in the Reduction of Sweep Interference from Television Receivers.—A. M. Intrator. (*Trans. Inst. Radio Engrs*, April 1956, Vol. BTR-2, No. 1, pp. 1–5.)

STATIONS AND COMMUNICATION SYSTEMS

621.376.5 **956**

Using Markerless Pulse Trains to Communicate.—M. Davidson, H. Joseph & N. Zucker. (*Electronics*, 21st Nov. 1958, Vol. 31, No. 47, pp. 89–91.) Three types of markerless pulse-train modulation are compared, and demodulating circuits for pulse-interval modulation are described.

621.376.5 : 621.391 **957**

Statistics of Regenerative Digital Transmission.—W. R. Bennett. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1501–1542.) Specific problems considered include the properties of a digital message pulse train as a random noise source, the effect of time jitter in a received pulse train on the recovered analogue signal, and the derivation of the pulse repetition frequency from a pulse train by shock excitation of a tuned circuit.

621.391 **958**

The Central Concepts of Communication Theory for Infinite Alphabets.—I. Fleischer. (*J. Math. Phys.*, Oct. 1958, Vol. 37, No. 3, pp. 223–228.)

621.391 : 621.376 **959**

Demodulation and Detection.—D. A. Bell. (*Electronic Radio Engr*, Jan. 1959, Vol. 36, No. 1, pp. 21–24.) A discussion of the basic processes, particularly in relation to information theory.

621.394.3 : 621.396.65 **960**

Teleprinting over Long-Distance Radio Links.—A. C. Croisdale. (*P.O. elect. Engrs' J.*, July & Oct. 1958, Vol. 51, Parts 2 & 3, pp. 88–93 & 219–225.) Methods for the transmission of 5-unit telegraphy signals and techniques for reducing the mutilation of signals on radio links are described. Automatic error correction and the resulting improvement in performance are discussed with reference to C.C.I.T.T. recommendations.

621.396(667) **961**

Radio Communications in Ghana.—R. G. Sharpe & D. R. Gamlen. (*Brit. Commun. Electronics*, Oct. 1958, Vol. 5, No. 10, pp. 750–756.)

621.396.2 **962**

The Bandwidth Occupied by a Class A1 Transmission and its Determination.—J. Marique. (*Rev. HF, Brussels*, 1957, Vol. 3, No. 10, pp. 359–368.)

621.396.2 **963**

Performance of some Radio Systems in the Presence of Thermal and Atmospheric Noise.—A. D. Watt, R. M. Coon, E. L. Maxwell & R. W. Plush. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1914–1923.) "The performance of several basic types of communication systems are determined experimentally, and in some cases theoretically, under typical conditions with steady or fading carriers, and in the presence of thermal or atmospheric noise. The relative efficiency of various carriers and the interference factor of various types of noise are found to be dependent upon the characteristics of the particular communication system as well as the characteristics of the carrier and noise themselves. Methods are considered for calculating errors expected from a given system, based upon the amplitude distribution of the noise envelope."

621.396.2 : [629.19+523.3 **964**

Long-Distance Telecommunications by means of Satellites.—F. Vilbig. (*Elektrotech. Z., Edn A*, 1st June 1958, Vol. 79, No. 11, pp. 375–382.) Propagation conditions and operational problems relating to the use of the moon and of artificial satellites as reflectors or relay stations are discussed.

621.396.25 : 621.394 **965**

High-Speed Frequency-Shift Keying of L.F. and V.L.F. Radio Circuits.—H. G. Wolff. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CS-5, No. 3, pp. 29–42. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1958, Vol. 46, No. 2, p. 515.)

621.396.3 **966**

Automatic Error Correction.—P. R. Keller & L. K. Wheeler. (*Wireless World*, Jan. 1959, Vol. 65, No. 1, pp. 28–33.) A description, with block diagrams, of the 'Autoplex' two-channel time-division electronic equipment, and of the improvement in the error rate with automatic repetition. See also 2224 of 1958 (Keller) and 960 above.

621.396.4 : 621.376.5 **967**

The Timing of High-Speed Regenerative Repeaters.—O. E. De Lange. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1455–1486.) A simplified method for determining the performance of the timing portion of chains of regenerative repeaters in multichannel p.c.m. systems. Two types are considered, those in which a repetitive timing signal is sent on one channel, and those in which timing is obtained from the coded pulse trains. The effect of random noise is calculated and other transmission defects are discussed.

621.396.4 : 621.376.5 **968**

Experiments on the Timing of Regenerative Repeaters.—O. E. De Lange & M. Pustelnyk. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1487–1500.) Experiments performed with self-timed binary regenerative repeaters to determine the behaviour of the timing portion of a chain of such repeaters are described. The number of errors produced by the action of noise on the timing system was negligible compared with those produced by other effects of noise.

621.396.4 : 621.376.5 **969**

Timing in a Long Chain of Regenerative Binary Repeaters.—H. E. Rowe. (*Bell Syst. tech. J.*, Nov. 1958, Vol. 37, No. 6, pp. 1543–1598.) The power spectra and the total powers of the timing noise, spacing noise, and alignment noise, caused by the input noise at each repeater are determined for a long chain of regenerative repeaters using either tuned-circuit or locked-oscillator timing filters. Tuning error effects are studied for a repeater chain using locked-oscillator timing circuits.

621.396.41 **970**

The 'Third Method'.—J. F. H. Aspinwall. (*Wireless World*, Jan. 1959, Vol. 65, No. 1, pp. 39–43.) The method described by D. K. Weaver, Jr, (*Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, pp. 1703–1705) is compared with filter and phasing methods of generating s.s.b. signals, and its application to radio telephony is briefly described.

621.396.41 **971**

A Compatible Single-Sideband Modulation System.—L. R. Kahn. (*Proc. Radio Cl. Amer.*, March 1958, Vol. 34, No. 1, pp. 3–9.) A system is described which is compatible with the existing d.s.b. a.m. system. It also compares favourably in signal/noise ratio, spectrum economy and reduction of selective fading distortion with the conventional s.s.b. system, and it is suitable for aeronautical communications. Operational tests show that the greatest improvement in audio fidelity and signal/noise ratio occurs in narrow-bandwidth domestic receivers. See 3247 of 1958 (Costas) for mathematical analysis of the system.

621.396.41 : 551.510.52 **972**

Quadruple - Diversity Tropospheric Scatter Systems.—W. G. Long & R. R. Weeks. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CS-5, No. 3, pp. 8–19.)

621.396.41 : 621.396.65 **973**

Microwave Network now Spans Canada.—M. James. (*Canad. Electronics Engng*, Sept. 1958, Vol. 2, No. 9, pp. 28–31.) A general description of the TD-2 multiplex f.m. system. See also 3638 of 1958 (Curtis et al.).

621.396.5 **974**

Practical and Theoretical Design Considerations for Bridge Negative-Feedback Amplifiers in Carrier Telephony.—M. J. Cotterill & J. W. Halina. (*Trans. Inst. Radio Engrs*, Sept. 1957, Vol. CS-5, No. 2, pp. 26–31.)

621.396.65 **975**

Scatter Equipment Built for Canadian Use.—J. A. Grant. (*Canad. Electronics Engng*, July 1958, Vol. 2, No. 7, pp. 30–33.) Terminal and repeater f.m. equipment covering the range 755–980 Mc/s is described, with particular reference to the 'serrasoid' sawtooth modulator.

621.396.65 : 621.396.967 **976**

Microwave Links for Radar Networks.—J. W. Sutherland. (*Brit. Commun. Electronics*, Sept. 1958, Vol. 5, No. 9, pp. 688–695.) The operational advantages of remote presentation of radar information and the parameters of practical transmission systems are discussed in detail. The use of travelling-wave valves in i.f. and non-demodulating types of repeater is described.

621.396.7 : 621.396.11 **977**

Radio Propagation Transmitting Station WWI at Havana, Illinois.—(*Tech. News Bull. nat. Bur. Stand.*, Aug. 1958, Vol. 42, No. 8, pp. 154–155.) The station, which is centred at 40° 13·27′ N, 90° 1·39′ W, is intended mainly for research in v.h.f. transmission. Its twelve transmitters operate on 30–108 Mc/s, at 3–50 kW. Details of the aerial system are given.

SUBSIDIARY APPARATUS

621.311.62 : 621.314.7 **978**

Boosting Power-Transistor Efficiency.—J. W. Caldwell & T. C. G. Wagner. (*Electronics*, 21st Nov. 1958, Vol. 31, No. 47, pp. 86–88.) High efficiency is obtained by carefully controlling the instantaneous voltage and current through the transistor.

621.316.721 : 621.318.3 : 621.314.7 **979**

Precision Current Regulator using Transistors.—S. D. Johnson & J. R. Singer. (*Rev. sci. Instrum.*, Nov. 1958, Vol. 29, No. 11, pp. 1026–1028.) A regulator for electromagnets of field strength between 1 000 and 5 000 G is described which is accurate to within two parts in 10^5.

621.316.722 : 621.314.7 **980**

Transistor Voltage Regulators.—G. W. Meszaros. (*Bell Lab. Rec.*, Dec. 1958, Vol. 36, No. 12, pp. 442–445.) Design considerations are discussed for series-type regulators and the magnetic-amplifier type designed for the TRIDAC computer is described.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 : 621.3.018.782.4 **981**

Demonstration of Delay Distortion Correction by Time-Reversal Techniques.—B. P. Bogert. (*Trans. Inst. Radio Engrs*, Dec. 1957, Vol. CS-5, No. 3, pp. 2–7. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1958, Vol. 46, No. 2, p. 515.)

621.397.5 **982**

The Possibilities of Reduced Television Bandwidth.—S. Deutsch. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. BTR-2, No. 3, pp. 69–82. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 253.)

621.397.5 : 621.396.4 **983**

Study of Multichannel Sound Transmission from a Single Transmitter: Application to Bilingual Television.—L. Bourassin. (*Électronique, Paris*, May & July/Aug. 1957, Nos. 126 & 128/129, pp. 45–50 & 37–44.) Continuation of 3312 of 1957.

621.397.611.001.4 **984**

Two New B.B.C. Transparencies for Testing Television Camera Channels.—(*B.B.C. Engng Div. Monographs*, Nov. 1958, No. 21, pp. 5–17.)

Part 1—Requirements, Design and Use of the Colour Response, and the Gradation and Resolution Transparencies.—G. Hersee (pp. 5–13).

Part 2—The Manufacture of the B.B.C. Test Transparency No. 51.—J. R. T. Royle (pp. 14–17).

621.397.611.2 **985**

The Problem of Inertia Effects in Television Camera Tubes of the Vidicon Type.—C. Kunze. (*Hochfrequenztech. u. Elektroakust.*, Nov. 1957, Vol. 66, No. 3, pp. 84–89.) Measurements show that inertia effects are mainly due to incomplete re-charging of the picture elements [see also 2552 of 1956 (Heimann)]. The dependence of these effects on operating parameters, and methods of eliminating inertia effects, such as roughening the target surface, are discussed.

621.397.62 : 535.623 **986**

Tentative Methods of Measurement of Colour Television Receiver Performance.—S. P. Ronzheimer & R. J. Farber. (*Trans. Inst. Radio Engrs*, April 1956, Vol. BTR-2, No. 1, pp. 10–30.)

621.397.62 : 621.314.7 **987**

Transistors in Television Receivers.—B. R. Overton. (*J. Telev. Soc.*, July–Sept. 1958, Vol. 8, No. 11, pp. 444–468.) "A complete receiver, operating from a 12-V battery (consumption 12 W approximately), and employing transistors throughout is described. Special attention is given to the three major technical problems of incorporating transistors in television receivers, namely, r.f. and i.f. amplification, video drive and line scanning. The paper discusses some new techniques for dealing with these problems, in particular a system of scan magnification, hitherto undisclosed. Some attempt is made to forecast future trends."

621.397.62 : 621.372.54 **988**

Trap Improves TV Picture.—G. C. Field. (*Electronics*, 21st Nov. 1958, Vol. 31, No. 47, pp. 100, 102.) The use of a bifilar-T trap to suppress adjacent-channel interference in 40-Mc/s i.f. amplifiers is described.

621.397.621.2 **989**

Improvements in Television Receivers: Part 4—Stabilization of the Line Deflection Circuit by means of a V.D.R. Resistor.—B. G. Dammers, A. G. W. Uitjens, A. Boekhorst & H. Heyligers. (*Electronic Applic.*, Aug. 1958, Vol. 18, No. 3, pp. 118–121.) A simplified version of the circuit proposed in 3316 of 1957 makes the protection device proposed in 282 of January (Dammers et al.) unnecessary. A voltage-dependent resistor is used to control the line output valve. The scanning current is stabilized to about $\pm$ 2% for line voltage variations of $\pm$ 13%.

621.397.621.2 : 535.623 : 621.385.832 **990**

Error Correction in Mask-Type Colour Television Tubes.—S. H. Kaplan. (*J. Telev. Soc.*, July–Sept. 1958, Vol. 8, No. 11, pp. 470–480.) Corrections for the errors which increase as a function of scan angle, and optical exposure methods for correcting triad size and location errors, are described. A proposal to eliminate triad shape errors by means of a radially distorted aperture mask pattern is given in detail.

621.397.7 **991**

B.B.C. Install 'Translator' for Improved TV Reception.—(*Brit. Commun. Electronics*, Oct. 1958, Vol. 5, No. 10, p. 757.) Signals transmitted on one channel are converted, without demodulation, to another and automatically re-radiated at low power over line-of-sight paths to the area to be served. Equipment at Folkestone is briefly described.

621.397.7 : 621.315.212 **992**

Interconnection of Television Cable Links at the Carrier-Frequency Stage according to the C.C.I.F.—R. Hoffmann. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 96–99.) Summary of the C.C.I.F. recommendations of December 1956 concerning the transmission of television signals over international cable routes.

621.397.8 : 535.623 **993**

The Origin and Measurement of Level-Dependent Phase and Amplitude Fluctuations in the Transmission

of the Subcarrier in Colour Television.—J. Piening. (*Nachrichtentech. Z.*, Feb. 1958, Vol. 11, No. 2, pp. 70–77.) Causes of amplitude and phase distortion are investigated quantitatively.

VALVES AND THERMIONICS

621.314.63 : 546.289 **994**

New Types of Germanium Diodes and their Circuit Applications.—G. Grimsdell. (*Electronic Engng*, Dec. 1958, Vol. 30, No. 370, pp. 709–710.) The properties of gold-bonded diodes and small-area junction diodes are compared with those of point-contact types and the relative advantages and disadvantages are outlined.

621.314.7 **995**

On the Variation of Transistor Small-Signal Parameters with Emitter Current and Collector Voltage.—N. I. Meyer. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 329–337.) Addendum to 3285 of 1958.

621.314.7 **996**

The Characteristic Frequencies of a Junction Transistor.—J. M. Rollett. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 344–347.) The relations between characteristic frequencies and intrinsic parameters are considered using the model which assumes one-dimensional flow of carriers across the base region. The frequencies discussed are the cut-off frequencies in the common-base and common-emitter connections, and the frequency at which the common-emitter current gain is unity.

621.314.7 **997**

Structure-Determined Gain-Band Product of Junction-Triode Transistors.—J. M. Early. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1924–1927.) The fundamental frequency limitations are discussed and it is shown that mesa-type transistors for use in the microwave region are theoretically possible. The gain-bandwidth product is an order of magnitude better than that of field-effect or analogue transistors.

621.314.7 **998**

Theory of the *P-N* Junction Device using Avalanche Multiplication.—T. Misawa. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, p. 1954.)

621.314.7 **999**

The Internal Current Gain of Drift Transistors.—F. J. Hyde. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1963–1964.)

621.314.7 : 546.289 **1000**

High-Frequency Germanium Transistors.—J. S. Lamming. (*Research, Lond.*, Nov. 1958, Vol. 11, pp. 425–431.) Design theory and production techniques are reviewed.

621.314.7 : 546.289 **1001**

Very-High-Power Transistors with Evaporated Aluminium Electrodes.—H. W. Henkels & G. Strull. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. ED-4, No. 4, pp. 291–294. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1958, Vol. 46, No. 2, p. 515.)

621.314.7 : 621.317.733 **1002**

A Wide-Band Bridge Yielding Directly the Device Parameters of Junction Transistors.—J. Zawels. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 21–25. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 932.)

621.314.7 : 621.317.799 **1003**

Sweep Equipment Displays Transistor Beta.—R. Zuleeg & J. Lindmayer. (*Electronics*, 5th Dec. 1958, Vol. 31, No. 49, pp. 100–101.)

621.314.7 : 621.318.57 **1004**

A New High-Current Mode of Transistor Operation.—C. G. Thornton & C. D. Simmons. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 6–10. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 932.)

621.314.7 : 621.318.57 **1005**

The 'Thyristor'—a New High-Speed Switching Transistor.—C. W. Mueller & J. Hilibrand. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 2–5. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 931–932.)

621.314.7 : 621.318.57 **1006**

A Transistor with Thyratron Characteristics and Related Devices.—W. von Münch. (*J. Brit. Instn Radio Engrs*, Nov. 1958, Vol. 18, No. 11, pp. 645–552.) Details of production and electrical performance are given for a device produced by immersing a tungsten whisker in the collector of an *n-p-n*-junction transistor, of high base resistivity, during the alloy process. Devices with more than one output electrode are developed from this design. A special structure for triggering by radiation and a symmetrical switching transistor are also studied.

621.314.7.002.2 : 546.28 **1007**

Silicon Transistors.—J. T. Kendall. (*Research, Lond.*, Oct. 1958, Vol. 11, No. 10, pp. 381–386.) Industrial manufacturing techniques are described.

621.314.7.01 **1008**

On the Need for Revision in Transistor Terminology and Notation.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1949–1950.)

621.314.7.012.8 **1009**

Transistor Equivalent Circuits.—C. Moerder. (*Elektrotech. Z., Edn A*, 1st July 1958, Vol. 79, No. 13, pp. 469–472.) Derivation of equivalent circuits from fundamental considerations.

621.383.4 **1010**

Properties of Cadmium Sulphide Photoconductive Cells.—R. L. Williams. (*Canad. J. Phys.*, Nov. 1958, Vol. 36, No. 11, pp. 1536–1550.)

621.383.49 **1011**

The Characteristics of Evaporated CdS and CdSe Photistors.—D. A. Anderson. (*Canad. Electronics Engng*, July 1958, Vol. 2, No. 7, pp. 23–29.) The performance of various types of photoconductive cell is discussed.

621.383.8 : 535.215-15 **1012**

High-Sensitivity Crystal Infrared Detectors.—M. E. Lasser, P. Cholet & E. C. Wurst, Jr. (*J. opt. Soc. Amer.*, July 1958, Vol. 48, No. 7, pp. 468–473.) The characteristics of *n*-type Au-doped Ge photoconductive cells are tabulated, and comparisons are made with the photovoltaic *p-n*-junction InSb cell. A multiple-contact cell is described which can locate a target as well as detect it without a moving optical system [see also 3691 of 1957 (Wallmark)].

621.385 : 534.39 **1013**

Analysis of Microphony in Electron Tubes.—A. Stecker. (*Electronic Applic.*, Aug. 1958, Vol. 18, No. 3, pp. 99–117.)

621.385.029.6 **1014**

On the Coupling Coefficients in the 'Coupled-Mode' Theory.—A. Yariv. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1956–1957.) Perturbation theory is used to evaluate the coupling coefficients for the case of small coupling, taking the field and current values as those applicable to the no-coupling case.

621.385.029.6 **1015**

Effect of Beam Coupling Coefficient on Broad-Band Operation of Multicavity Klystrons.—S. V. Yadavalli. (*Proc. Inst. Radio Engrs*, Dec. 1958, Vol. 46, No. 12, pp. 1957–1958.) Wide-band multicavity klystrons are more efficient in the L band of frequencies than in the S band, and are better operated at higher voltages than synchronously tuned klystrons.

621.385.029.6 **1016**

Pulse-Modulated Beam Current Improves Operation of Mixer-Series Klystrons.—A. K. Scrivens. (*Canad. Electronics Engng*, Oct. 1958, Vol. 2, No. 10, pp. 32–34.)

621.385.029.6 **1017**

On the Design of the Transition Region of Axisymmetric, Magnetically Focused Beam Valves.—V. Bevc, J. L. Palmer & C. Süsskind. (*J. Brit. Instn Radio Engrs*, Dec. 1958, Vol. 18, No. 12, pp. 696–708.) A method is described for using an analogue computer to trace out electron trajectories. It is applied to the presentation of beam envelopes for Brillouin flow, periodic magnetic focusing and space-charge-balanced flow. By matching these with envelopes derived from the theory of the Pierce gun, it is possible to specify the dimensions of a gun for producing a required beam.

621.385.029.6 **1018**

A Relativistic Treatment of Space-Charge-Limited Current in a Planar Diode Magnetron before Cut-Off.—J. A. Bradshaw. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 300–306.) "Relativistic expressions, correct to first order in a current density parameter, are obtained from two cross-coupled integral equations for the potential functions in a planar diode magnetron. These expressions are evaluated and plotted in two figures, and compared with expressions given by Gold [3695 of 1957]."

621.385.029.6 **1019**

Travelling-Wave Valves.—J. Voge. (*Ann. Télécommun.*, March & April 1957, Vol. 12, Nos. 3 & 4, pp. 92–104 & 105–119.) A general survey of the development of the device. Carcinotrons and amplifiers are described and comparative tables of French and American types are given. Amplifier noise and methods for reducing it, and the characteristics of several linear accelerators are discussed. 28 references.

621.385.029.6 **1020**

Wave Matrices Applied to a Periodically Loaded Travelling-Wave Tube.—D. E. T. F. Ashby. (*J. Electronics Control*, Oct. 1958, Vol. 5, No. 4, pp. 338–343.) The derivation of the matrices is shown, and they are used to determine the conditions necessary for oscillation due to feedback caused by reflections from the periodic discontinuities. Oscillations may be produced unintentionally, and be mistaken for backward-wave oscillations.

621.385.029.6 : 621.318.2 **1021**

The Design of Periodic Magnetic Focusing Structures.—J. E. Sterrett & H. Heffner. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1, pp. 35–42. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 932.)

621.385.029.6 : 621.372.2 **1022**

A Note on the Dispersion of Interdigital Delay Lines.—F. Paschke. (*RCA Rev.*, Sept. 1958, Vol. 19, No. 3, pp. 418–422.) "It is shown that the effect of the backwall on the dispersion of an interdigital delay line can be taken into account by a lumped susceptance which periodically loads the 'ideal' line. Experimental results are in good agreement with the theory." See also 629 of 1957.

621.385.029.6 : 621.396.822 **1023**

Noise Wave Excitation at the Cathode of a Microwave Beam Amplifier.—W. R. Beam. (*Trans. Inst. Radio Engrs*, July 1957, Vol. ED-4, No. 3, pp. 226–234. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1760.) See also 4067 of 1957 (Knechtli & Beam).

621.385.029.62/.63 **1024**

Travelling-Wave Amplifiers and Backward-Wave Oscillators for V.H.F.—D. A. Dunn. (*Trans. Inst. Radio Engrs*, July 1957, Vol. ED-4, No. 3, pp. 246–264. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1760–1761.)

621.385.029.64 : 537.533 : 621.375.9 **1025**

Parametric Amplification of Space-Charge Waves.—A. Ashkin. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1646–1651.) An experimental investigation of the theory of Louisell & Quate (2273 of 1958), which shows that a signal imposed on a beam as a 'fast' or 'slow' space-charge wave can be exponentially amplified by strongly modulating the beam with a r.f. wave of twice the signal frequency. Over a 10-in. length of beam a 41-dB increase has been observed. With the high-level signal frequency lower than the signal frequency, an increase of 30 dB over a 9·2-in. length was observed.

621.385.032.213.13 **1026**

On the Conduction Mechanism of Oxide-Coated Cathode.—H. Mizuno. (*J. phys. Soc. Japan*, Oct. 1958, Vol. 13, No. 10, pp. 1234–1235.)

621.385.032.213.13 : 621.396.822 **1027**

Noise in Oxide Cathode Coatings.—H. J. Hannam & A. van der Ziel. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, pp. 1702–1705.) A discussion of noise measurements at 8 Mc/s and 30 c/s. The h.f. measurements show thermal noise at high and low cathode temperatures with a pronounced noise peak caused by shot noise where pore conduction changes to grain conduction. At l.f. the results show that the pores are inherently noisier than the grains.

621.385.032.213.13 : 621.396.822 **1028**

New Mechanism for the Generation of Flicker Noise.—F. Fisher & I. P. Valkó. (*J. appl. Phys.*, Dec. 1958, Vol. 29, No. 12, p. 1772.) The effect of cathode porosity on flicker noise in valves has been investigated and results confirm the theoretical predictions of Lindemann & van der Ziel (1284 of 1957).

621.385.032.213.63 **1029**

The Breakdown of Cathode Coatings.—B. Wolk. (*Sylvania Technologist*, Oct. 1957, Vol. 10, No. 4, pp. 106–110.) A report is given of an experimental study of the processing of cathode coatings under a set of controlled temperature and time conditions. Results indicate that thermal decomposition characteristics and particle size are related. Factors accounting for poor processing reliability are also considered.

621.385.032.24 **1030**

The Grid Emitting Properties of Titanium.—J. A. Champion. (*Brit. J. appl. Phys.*, Dec. 1958, Vol. 9, No. 12, pp. 491–495.) Experiments show that titanium is suitable as a screen-grid winding wire and for other electrode applications when used in the temperature range 700–900°C; above this cathode poisoning occurs.

621.385.032.26 **1031**

Dynamics of Electron Beams from Magnetically Shielded Guns.—A. Ashkin. (*J. appl. Phys.*, Nov. 1958, Vol. 29, No. 11, pp. 1594–1604.) Theoretical and experimental investigation of electron orbits and beam shapes. Current density variations and transverse velocity are measured with a beam analyser. In the limit of negligible space charge the observations are explained on the basis of simple energy considerations, taking account of the effect of transverse thermal velocities. In the presence of space charge the beam behaviour is qualitatively unchanged, provided that the magnetic field is higher than about three times the Brillouin field.

621.385.3 **1032**

New Electron Tubes for Wide-Band Amplifiers.—S. Edsman. (*Ericsson Rev.*, 1958, Vol. 35, No. 3, pp. 98–102.) A note on the characteristics and operation of a pentode Type 5847/404A, triode Type 5842/417A and tetrode Type 7150.

621.385.3 : 621.365.5 **1033**

Two Methods of Calculation for the Class-C Operation of Transmitter Valves.—N. Weyss. (*Elektrotech. u. Maschinenb.*, 1st Dec. 1958, Vol. 75, No. 23, pp. 633–638.) Oscillators for r.f. heating are considered.

621.385.832 : 621.317.755 **1034**

Oscilloscope Tube with Travelling-Wave Deflection System and Large Field of View.—Niklas & Wimpffen. (See 927.)

621.385.832 : 621.397.621.2 : 666.1 **1035**

A Method of Sealing the Window and Cone of Television Picture Tubes.—A. H. Edens. (*Philips tech. Rev.*, 31st May 1958, Vol. 19, No. 11, pp. 318–323; *Glass Ind.*, Oct. 1958, Vol. 39, No. 10, pp. 534–538.)

MISCELLANEOUS

061.6 : 621.396 **1036**

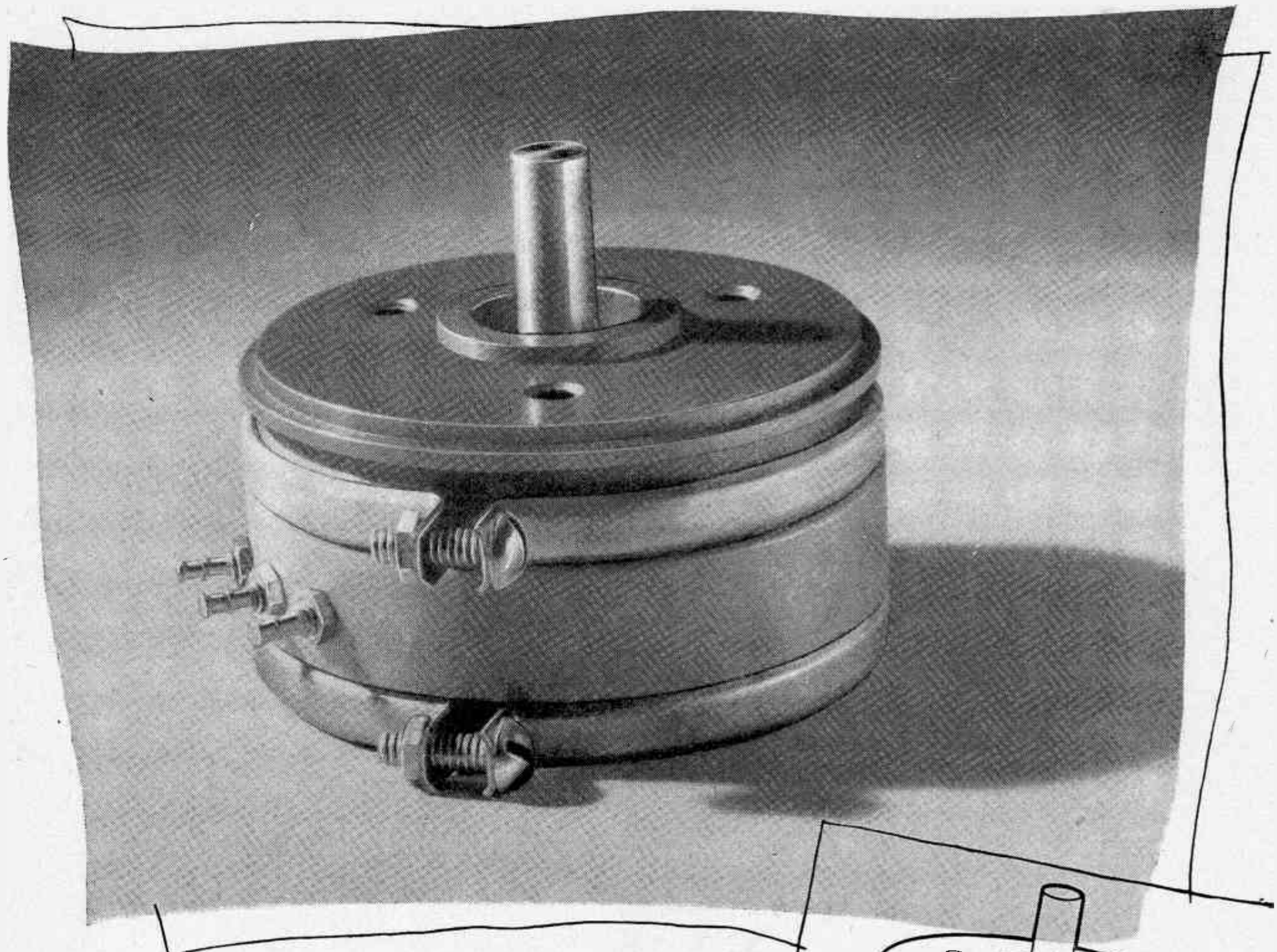
International Radio Organizations—some Aspects of their Work.—R. L. Smith-Rose. (*J. Brit. Instn Radio Engrs*, Nov. 1958, Vol. 18, No. 11, pp. 631–639.) An historical outline is given of the growth of communications and the resulting organizations set up to control international affairs. Details of the present work of the International Radio Consultative Committee (C.C.I.R.) are given and the activities of the International Scientific Radio Union (U.R.S.I.) are reviewed.

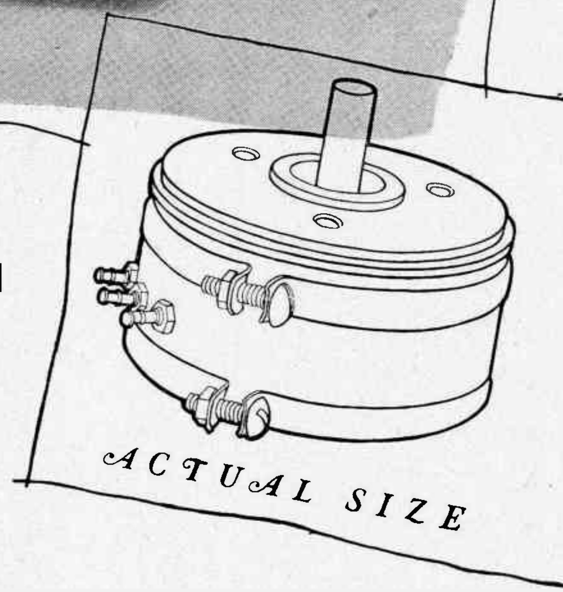
621.3.002.5 **1037**

Production Machinery for the Electronics Industry.—G. Sideris. (*Electronics*, 24th Oct. 1958, Vol. 31, No. 43, pp. 73–84.) A review dealing with the modernization of machinery, tools, and plant layout, and with advances in component production and equipment assembly techniques.

621.37/.38].004.1 **1038**

The Influence of Interaction Reliabilities.—M. A. Acheson. (*Sylvania Technologist*, July 1958, Vol. 11, No. 3, pp. 91–95.) Examples are given of simple electronic systems in which interaction between two or more reliabilities of parts influences the total reliability.





BECKMAN Instruments Limited

Series 'J' HELIPOT

A further example of the range of helical potentiometers manufactured by Beckman Instruments Limited in Scotland.

The Helipot Series J are high precision, single turn potentiometers.

Bush mounting or Servo mounting;
up to 8 sections can be ganged on a common shaft.

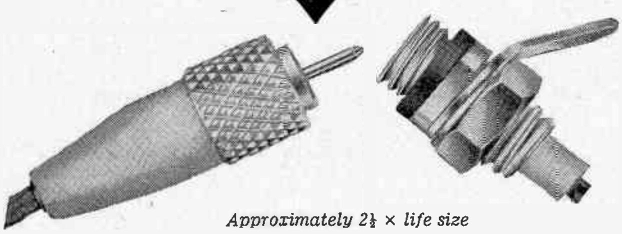
Please write for further details to:

WINSTON ELECTRONICS LIMITED

GOVETT AVENUE · SHEPPERTON · MIDDLESEX · *Walton-on-Thames 6321*

The very best connections ...secured by Plessey

SUB-MINIATURE COAXIAL CONNECTORS



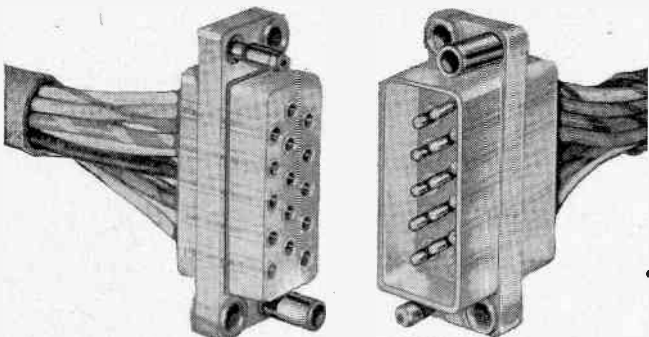
Approximately 2½ × life size

As a contribution towards increasingly compact equipment, Plessey have introduced this new, highest quality and fully comprehensive range to allow a new approach on applications hitherto restricted by the limitations of existing connectors.

Designed for the matched impedance coupling of high frequency coaxial cables operating in the super high frequency bands, these connectors—

* have a working voltage of 650 volts PEAK at sea level, and matched impedance coupling of 50 ohm lines is accommodated.
* have hard gold plated contacts on silver plate to give maximum performance with minimum voltage drop.

SERIES '110' (15- and 30-way) MINIATURE RECTANGULAR CONNECTORS



Developed specifically by Plessey to meet the demand for a safe, inexpensive connector for commercial applications, this new series embodies excellent electrical and mechanical characteristics, and the many unique features that make it really outstanding include:—

* Plug pins and socket inserts are polythene shrouded to dispense with gaskets and ensure insert anchorage.
* Mismating is prevented by corner pins and corner sockets.
* Extreme simplicity of wiring, demands less-skilled operation than the orthodox methods of soldering pins *in situ*.

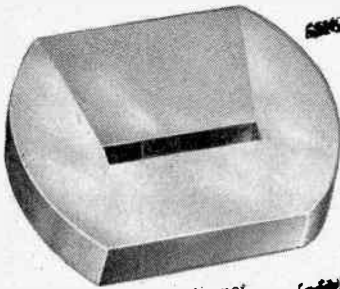
Plessey

ELECTRICAL CONNECTORS DIVISION · THE PLESSEY COMPANY LIMITED

CHENEY MANOR · SWINDON · WILTS · TELEPHONE: SWINDON 6251

Overseas Sales Organisation: PLESSEY INTERNATIONAL LIMITED · ILFORD · ESSEX

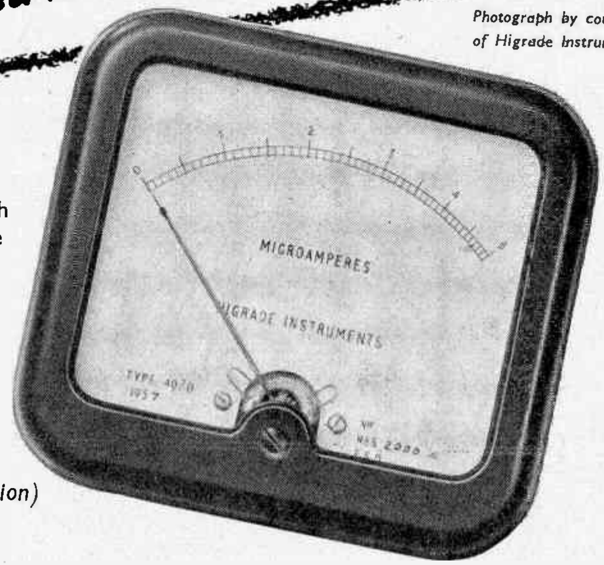
CW6



Actual size of Magnet

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

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*

bare and insulated

resistance wires

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

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

VACROM (Regd.)

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

EUREKA (Regd.)

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

insulated wires

THE LONDON ELECTRIC WIRE COMPANY AND SMITHS LIMITED · *LEYTON, LONDON E10*

L·E·W· VACTITE

bare wires

VACTITE WIRE COMPANY LIMITED
75 ST. SIMON STREET, SALFORD 3, LANCS

A Mains Voltage Fluctuation problem

A Unicam type S.P. 600 Spectrophotometer in use with an 'Advance' C.V.T.

in Absorption Analysis

The Spectrophotometer employs a particular optical/electronic system in which a light source, photocells and special amplifier play a vital part. Vital, because it is upon their correct performance characteristics that the accuracy of the instrument depends.

But their correct performance, in turn depends upon a stable supply voltage. How is this achieved?

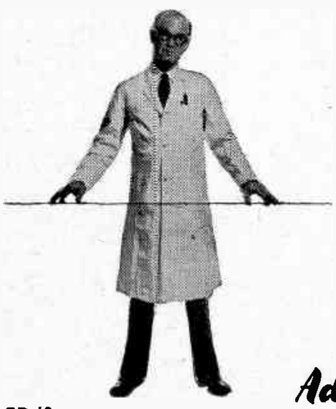
. . . straightened out by

In the Unicam S.P.600 Spectrophotometer illustrated, the tungsten lamp and valve heaters in the amplifier are supplied from the mains through an "Advance" Constant Voltage Transformer (seen on the right of photograph). As a result the accuracy of the instrument is unaffected by fluctuating supply voltage.

"Advance" Constant Voltage Transformers provide a.c. voltage stabilization of ± 1% for input variations of up to ± 15% at maximum load. For power requirements from 4 to 6000 watts, they are automatic and contain no moving parts.

Advance

CONSTANT VOLTAGE TRANSFORMERS



"VOLTAGE STABILIZATION"

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

POST THIS COUPON TO-DAY

To ADVANCE COMPONENTS LIMITED
ROEBUCK ROAD, HAINAULT, ILFORD, ESSEX

Please send me a copy of your booklet "Voltage Stabilization" by Advance

NAME ....

POSITION ....

COMPANY ....

ADDRESS ....

.... R13

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

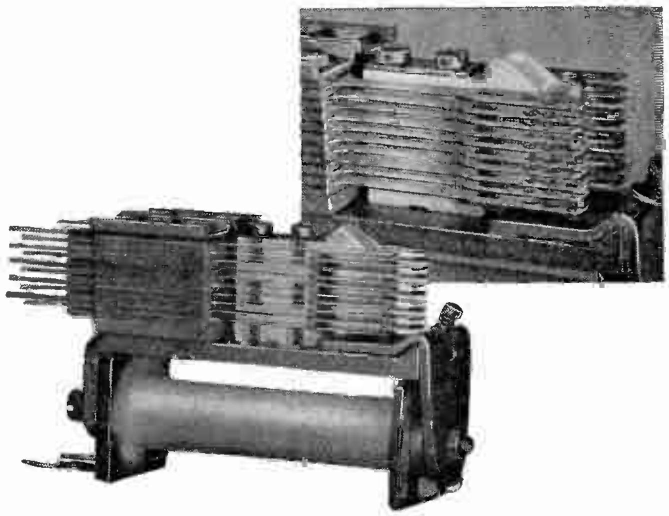
GD 13

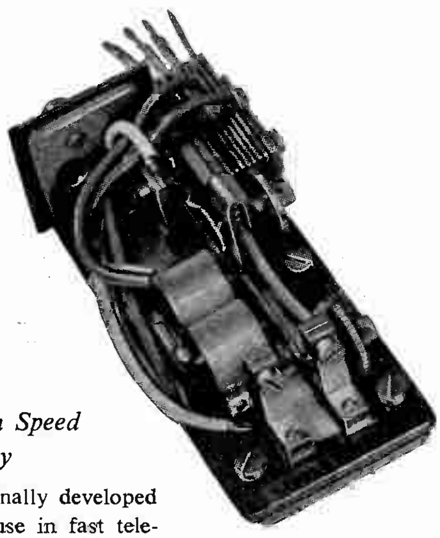
3 outstanding RELAYS

These relays are available for early delivery; your enquiries are welcomed

Comb Relay

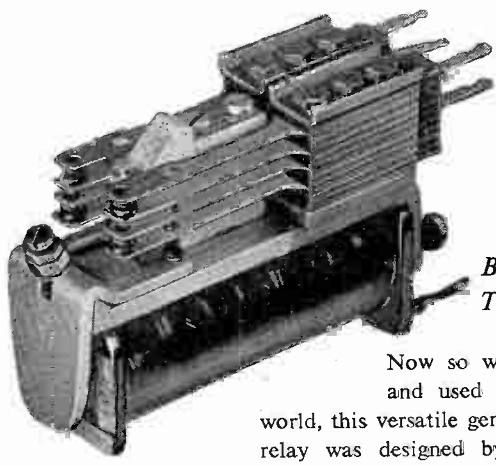
This is our latest development in relays and will give 100 million operations without the need for readjustment. Long life and reliability have led to its adoption by the British Post Office for use where continuous pulsing is required. A wide range of contact combinations is available with a maximum of 10 make or 10 break actions. It is suitable for switching light currents at 250 v.





High Speed Relay

Originally developed for use in fast telephone switching circuits, this high speed relay is of great value in automation systems. It can effect a single changeover action in as little as 1·5 milliseconds. A high speed relay is also available with two changeover actions.



B.P.O. 3000 Type

Now so widely known and used all over the world, this versatile general purpose relay was designed by us in the early thirties for the British Post Office. Many millions are now in use in both telephone and industrial applications. It is available in an extremely wide range of coil resistances and spring set combinations for practically all normal voltages and frequencies.

SIEMENS EDISWAN

SIEMENS EDISON SWAN LTD
An A.E.I. Company
Telecommunications Division P.D.7.
Woolwich, London, SE18
Telephone: Woolwich 2020 Extn. 621

P.D.7/1.

JUST RELEASED ! !

SIGNAL GENERATOR Model 68 A/M

with Meter for Monitoring Output

Exceptionally Wide Range—100 Kcs—240 Mcs

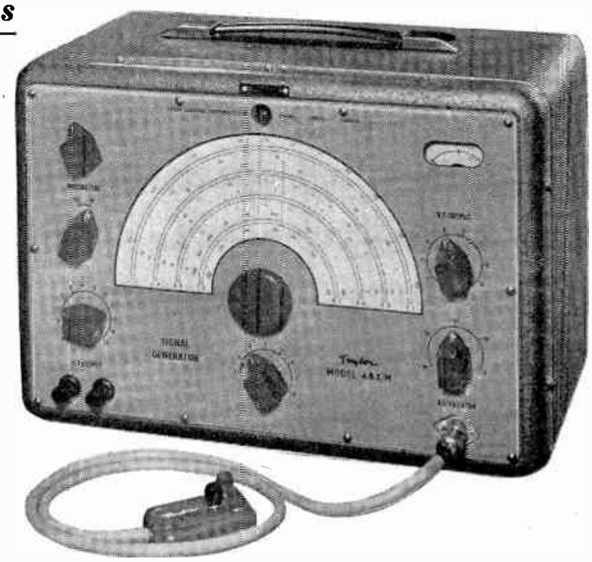
All on Fundamentals

Designed for use in the servicing of television and radio receivers, and may also be used in the C-W alignment of discriminators, ratio detectors and the RF and IF stages of FM receivers. A meter is incorporated for monitoring output.

A five-section attenuator with a non-inductive potentiometer ensures accurate attenuation and provides a constant source impedance over the whole frequency range. Radio frequency leakage is reduced to a minimum. Power-line leakage negligible. An added facility in the servicing of audio and video stages is afforded by a variable 400-cycle signal. **8 BANDS**: The following **Frequency Ranges**: 100 Kc/s-300 Kc/s, 300 Kc/s-1 Mc/s, 1 Mc/s-3 Mc/s, 3 Mc/s-10 Mc/s, 10 Mc/s-30 Mc/s, 30 Mc/s-55 Mc/s, 55 Mc/s-110 Mc/s, 110 Mc/s-240 Mc/s, all on fundamentals. **Scale length:** 58 in. (146 cms.). **Calibration Accuracy:** ±1%. **R.F. Output:** 100 mV normal.

Leakage: 3 μV. approx. **Attenuation:** Coarse: 5 steps of —20dB; Fine: variable to —20dB approx. (rather more on low frequency ranges). **AF Output:** Direct connection to the AF output, IV level maximum is provided. **Modulation:** 400 cycles sine wave, 30% depth. **Output Impedance:** 75 ohms, approx. via coaxial lead. Instrument supplied with dummy aerial. **Protection:** The instrument is protected by a mains fuse. **Weight:** 18 lb. **Dimensions:** 9 in. x 13 in. x 5 in. **Case:** Metal case grey hammer finish with matching Perspex panel.

Model 68A with identical specifications but not incorporating meter for monitoring output is available at **TRADE PRICE £23.7.6.**



TRADE PRICE £27. 12. 6. **Prompt Delivery**

Hire Purchase or Credit Sale Terms available

NOW AVAILABLE!
New edition of
VALVE MANUAL FOR TAYLOR TESTERS
Gives over 5,000 settings and characteristics. Price 7/6 post free!

Taylor

Write for full details

TAYLOR ELECTRICAL INSTRUMENTS LTD.
MONTROSE AVENUE, SLOUGH, BUCKS. Telephone Slough 21381

THE SMALLEST EVER RECHARGEABLE CELL

DEAC

PERMA-SEAL

For
Radios
Deaf-Aids
Guided Weapons
Tape Recorders
Shavers, Etc.

A miracle of engineering precision. This 20 mAh capacity, hermetically sealed, gas tight accumulator provides the smallest rechargeable power source available in this country.

It is the latest addition to the famous DEAC range of patented rechargeable cells which can be permanently "wired in" to work with the utmost efficiency under all conditions without maintenance.

The range includes Disc, Rectangular and Cylindrical types. The Disc-type cells can be stacked to form up to 12 volt single batteries and combinations of stacks give even higher voltages and capacities.

- ★ **No corrosion**
- ★ **No gassing**
- ★ **No maintenance**
- ★ **Unlimited shelf life**
- ★ **Robust and compact**
- ★ **From 20 mAh to 23 Ah**

G. A. STANLEY PALMER LIMITED
Maxwell House, Arundel Street, London, W.C.2
Telephone : Temple Bar 3721

A NEW HIGH SPEED OSCILLOSCOPE

MODEL 301 D.C. to 40 Mc/s.

Designed around the latest type of high sensitivity cathode ray tube, this instrument provides all the essential features of a wide band oscilloscope with accurate measuring facilities and an exceptionally wide range of sweep speeds.

The aim has been to get these features incorporated in the least complicated manner, resulting in a compact, rugged and reliable design produced to the highest standard of workmanship at an economical price.

CATHODE RAY TUBE

Five-inch diameter flat faced with a window area of 4 x 10 cms. Total accelerating voltage applied by means of a helical system is 10 KV, resulting in a small beam spot and a trace capable of resolving fine detail.

TIME BASE

Calibrated sweep times from 0·5 sec. to 0·1 μsec. per cm. accurate within 2% by direct reading of time per cm. An uncalibrated fine control increases the maximum sweep time to 1·2 sec./cm. or 12 secs. for a full sweep of 10 cms. Sweep expansion x 10 reduces the minimum sweep time to 10 mμsec./cm., equivalent to a sweep speed of 1 mm. per mμsec. Versatile and highly efficient triggering circuitry provides both repetitive and single stroke conditions.

Y AMPLIFIER

Band width .... D.C. to 40 Mc/s (−3dB)
Rise time .... 9 mμsec.
Sensitivity .... 1 cm. per 100 mV.

By switch control the sensitivity can be increased to 1 cm. per 10 mV over a bandwidth of 2·5 c/s to 20 Mc/s. A nine-step attenuator and a fine gain control extend the sensitivity range to approximately 1 cm. per 12 V. An RC probe is available with 10-1 reduction factor, extending the range to a minimum of 1 cm. per 120 V. A balanced signal delay is incorporated in the amplifier enabling the observation of pulse leading edges.

VOLTAGE MEASUREMENT

A calibrator is provided giving output of 40 mV and 400 mV accurate within 2%.

POWER SUPPLIES

Both H.T. and E.H.T. voltages are stabilised over a wide range of mains input voltage variations, and the power supplies are built into the case of the oscilloscope.

NAGARD LTD

18 AVENUE ROAD · BELMONT SURREY
TELEPHONE: VIGilant 9161-2

WOLSEY ELECTRONICS LTD.

Cray Ave., St. Mary Cray
Orpington, Kent

requires a

SENIOR DEVELOPMENT ENGINEER

for Antenna and RF development. St. Mary Cray is 25 Minutes from Holborn Viaduct. Applicants must be qualified GRAD Brit IRE, GRAD IEE or at least equivalent, preferably specialising in RF work. The Company is making excellent progress in its market due to a policy of continuous improvement in design and manufacture. The Lab is adequately equipped and is considered an important unit of the business. This is a happy firm with an informal but vigorous character. Candidates should expect to use continuous effort, which however will be properly appreciated by the management. Salary for this appointment from £1,000 upwards according to qualifications and ability. Applications originally in writing please to R. S. Roberts, M. Brit IRE, Sen MIRE, at above address.

YOUR FUTURE IN ELECTRONICS

TRANSISTORS—FERRITES—MAGNETICS
PRINTED CIRCUITS—POTTED CIRCUITS

and their application to all branches of

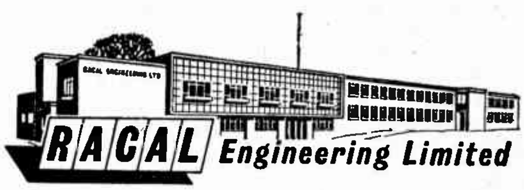
RADIO COMMUNICATIONS & ELECTRONIC INSTRUMENTATION

If you are interested in any of the above subjects and are anything from a Technician to a Senior Engineer/Physicist, staff vacancies are currently available in the salary scale

£500 to £2,500 p.a.

depending on qualifications and experience.

An excellent staff contributory superannuation scheme is in operation and New Town Housing can be made available to successful candidates. Applications, which will be treated in the strictest confidence, giving details of age, qualifications and experience to date, to the Technical Director,



WESTERN ROAD · BRACKNELL · BERKSHIRE

ELECTRONIC & RADIO ENGINEER

CLASSIFIED ADVERTISEMENTS

Advertisement Rates. 6d. *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 1959 issue is first post 23rd March 1959

SITUATIONS VACANT

INSTRUMENT ENGINEERING IN MEDICAL RESEARCH

THE Medical Research Council invite applications from Junior Engineers for a vacancy in the Instrument Laboratories at the National Institute for Medical Research. Staff in these laboratories undertake the development of specialized laboratory instruments and equipment in consultation or collaboration with various Scientific Divisions. The work is extremely varied and would be both congenial and satisfying to the right man who would require an interest in electronics, physics and general engineering. Applicants, preferably under 26 years of age, should possess an appropriate degree or City & Guilds Final (Telecomms.) or H.N.C. Salary will be on a scale:
JUNIOR Technical Officer, £510 x £35—£660 x £30—£780, with prospect of promotion to Technical Officer, £830—£1,150.
THE appointment will be superannuated and, after a provisional period of one year, established.
APPLICATION forms obtainable from Personnel Officer, National Institute for Medical Research, The Ridgeway, Mill Hill, N.W.7. [1305

BATTERSEA COLLEGE OF TECHNOLOGY LONDON, S.W.11
(A College of Advanced Technology)
DEPARTMENT OF PHYSICS

APPLICATIONS are invited for a Lecturer in Microwave Physics. Research experience essential; teaching experience desirable. The microwave section is concerned with research, for which good facilities are provided, and with the teaching of students in their final undergraduate and post-graduate years. Salary £1,298 x £31 10s.—£1,418, less statutory deductions. Further particulars from the Clerk to the Governing Body by whom applications should be received as soon as possible. [1301

NORWOOD TECHNICAL COLLEGE

LECTURER in Department of Telecommunications and Electronics for April, 1959, or as soon after as possible.
GOOD academic and professional qualifications and experience required; specialist knowledge or experience of computers of very high frequency techniques an advantage.
BURNHAM F.E. salary scale: £1,260 x £31 10s.—£1,417 10s. plus London Allowance £37 16s. or £50 8s.
APPLICATION forms from Principal at College, Knights Hill, S.E.27, to be returned by March 6, 1959. (249.) [1299

ELECTRO-ENCEPHALOGRAPHY RECORDIST, GRADE I

APPLICATIONS are invited from experienced technicians. The department also serves other hospitals in the district and a new research unit for neuro-surgery which opens shortly. Whitley Council salary scales and conditions apply. Apply, giving qualifications, experience, and quote two referees to Medical Superintendent, Parkside Hospital, Macclesfield. [1242

TECHNICAL Writer. An interesting opening occurs for a qualified electronic engineer, man or woman, able to write clear, concise English. Duties include preparation of sales literature, instruction manuals and technical articles on electronic laboratory instruments and industrial equipment. Apply, if possible with samples of your work, to Personnel Officer, Airmec Limited, High Wycombe, Bucks. [1297

MINISTRY OF TRANSPORT AND CIVIL AVIATION requires Electrical Engineers (Assistant Signals Officers) for aviation telecommunications and electronic navigational aids. Minimum age 23, 1st or 2nd Class degree in Physics or Engineering, or A.M.I.E.E. or A.F.R.Ae.S. (candidates with Parts I, II and III of A.M.I.E.E. or Parts I and II of A.F.R.Ae.S. or equivalent, or of very high professional attainment without these qualifications considered). Salary £665 (age 23) to £1,085 (age 34) maximum £1,250. Slightly lower outside London and for women. Five-day week. Further details and forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting D.70/9A. [1302

SITUATIONS VACANT

TECHNICAL WRITER
Leading Norwegian
TELECOMMUNICATION AND ELECTRONIC FIRM
seeks
Technical Writer
To take Charge and Build up the Handbook Section
Good prospects
Re-location expenses paid
Interviews will be arranged in England
NERA BERGEN A/S
Bergen, Norway
[1306

ELECTRONICS MAINTENANCE ENGINEERS

VICKERS RESEARCH LTD., Brooklands Road, Weybridge, have vacancies for engineers experienced in the repair of electronic test equipment.
THEY will be required to be conversant with a wide range of instruments and maintain them to a high standard of reliability and accuracy. Salary will be appropriate to experience and qualifications.
EMPLOYMENT will be at Weybridge at first and later will be transferred to Sunninghill, near Ascot.
SUITABLY experienced persons are invited to write to the above address for application forms. [1307

NIGERIAN BROADCASTING CORPORATION

ASSISTANT Engineer-in-Charge required on contract for two tours of 18 months in first instance. Salary scale (including 15% contract addition and inducement addition) £1,362 rising to £1,680. Gratuity at rate of £150 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 U.K. Liberal leave on full salary. Candidates should preferably be Graduates of the Institution of Electrical Engineers or equivalent. They should have wide theoretical and practical experience of low frequency amplifiers, low power transmitters, audio amplifiers, communication receivers and magnetic tape recording equipment, and be capable of maintaining and operating them. Knowledge of diesel generating plant 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/50274/EO. [1304

ROAD RESEARCH LABORATORY (Traffic and Safety Division), Langley, Bucks, requires a Scientific Officer Ref. A.87/8A and Assistant Experimental Officer (Physicist or Engineer) Ref. A.260/8A for work on development or electronic devices for experimental use in vehicles. Interest in or experience with transistor techniques and servo mechanisms an advantage. Qualifications: S.O. 1st or 2nd Honours degree in Physics or Engineering or equivalent qualifications. A.E.O. G.C.E. (Advanced) in two science or a science and mathematics subject. Over 22, pass degree, H.N.C. in Physics or Engineering, or equivalent qualifications generally expected. Salaries: S.O. £615/£1,080 (men). Normal prospects of promotion to Principal S.O. in mid-thirties (salary up to £2,000). A.E.O. £377 10s. (age 18)—£825 (men). Salaries exclusive of recent 3½ per cent increase. Five-day week. Forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1 (quote appropriate reference). [1303

PATENTS

PATENT No. 720702 "Piezoelectric Crystal Resonator Support" for sale or licence. Apply: Chatwin & Company, Chartered Patent Agents, 253 Gray's Inn Road, London, W.C.1. [1298

SCHOLARSHIPS

SCHOLARSHIPS

THE Council of The Institution of Electrical Engineers will consider for award this year, three Research Scholarships, two Graduate Scholarships and seven Student Scholarships for study at universities and technical colleges. Research Scholarships: value £100—£500 per annum. Graduate Scholarships: value £400 per annum. Student Scholarships: value £50—£200.
THE closing date for the receipt of applications is May 1, 1959, for Student Scholarships, and June 1, 1959, for Graduate and Research Scholarships.
FULL particulars of the conditions of award and nomination forms may be obtained on application to the Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2. [1300

STEEL SHELVING

100 bays of brand new adjustable STEEL SHELVING, 72″ high × 34″ wide × 12″ deep, stove enamelled, dark green, sent unassembled, six-shelf bay, £3 15s. 0d.

Sample delivered free; quantity discounts.

N. C. BROWN LTD.
EAGLE STEELWORKS
HEYWOOD, LANCS. Telephone 69018

TEST ENGINEERS
SERVICE ENGINEERS
PROTOTYPE WIREMEN
SOLARTRON

This Group of Companies, which has given the world some of the most significant advances in electronic instruments, is expanding rapidly and needs additional:

EXPERIENCED ELECTRONIC TESTERS and SERVICE ENGINEERS

capable of fault finding and working on a wide range of oscilloscopes, Servo test equipment and other instruments

EXPERIENCED PROTOTYPE WIREMEN

with a basic knowledge of circuitry, who are able to wire from circuit diagrams and incorporate modifications.

A non-contributory pension and life assurance scheme and a sickness benefit scheme are among the amenities offered by the Group. Annual holidays are increased by one day for each year of service to a maximum of three weeks after five years.

Applications, stating age, previous experience and present salary should be addressed to the:

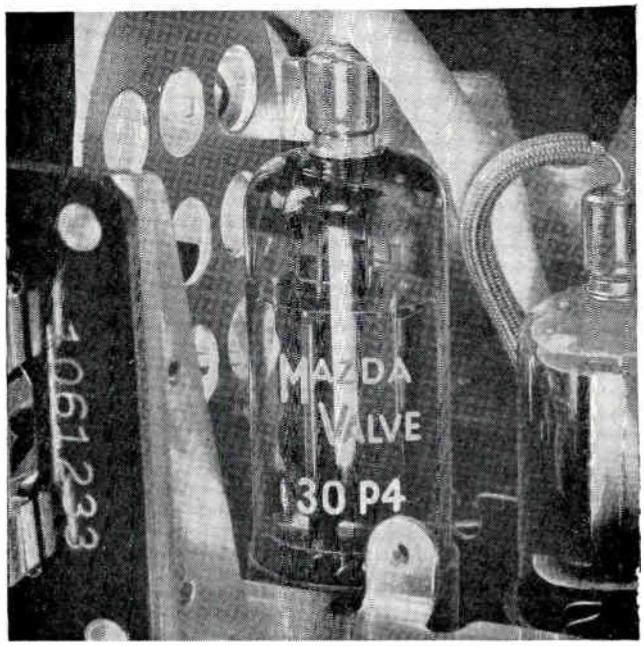
Personnel Officer,
THE SOLARTRON ELECTRONIC GROUP LTD.,
Thames Ditton, Surrey.
Quoting ref.: 322.

Index to Advertisers

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.

Introducing another outstanding Ediswan Mazda valve type 30P4

For the information of set designers we are publishing details of individual 0.3 amp. heater valves in our "First Preference" Range for TV circuits. If you are a TV manufacturer we shall be pleased to supply full technical details of our "First Preference" Range, together with a set of valves for testing, on receipt of your enquiry. The valve dealt with here is the Type 30P4, a line output beam tetrode, for a.c./d.c. Mains Television Receivers.

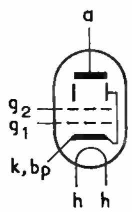


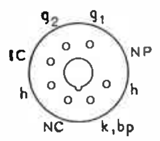
EDISWAN MAZDA 30P4

Line Output Beam Tetrode, for a.c./d.c. Mains Television Receivers

Heater Current (amps)	I_h	0.3
Heater Voltage (volts)	V_h	25

Base: International Octal (I07) Top Cap: Anode





VIEW OF FREE END

MAXIMUM DESIGN CENTRE RATINGS

Anode Dissipation (watts)	$P_{a(max)}$	10
Screen Dissipation (watts)	$P_{g2(max)}$	4
Cathode Current (mA)	$I_{k(max)}$	160
Anode Voltage (volts)	$V_{a(max)}$	400
Peak Anode Voltage—Pulse Positive* (kV)	$V_{a\ (pk)max}$	6†
Screen Voltage (volts)	$V_{g2(max)}$	250
Peak Screen Voltage—Pulse Negative* (volts)	$V_{g2(pk)max}$	2000
Heater to Cathode Voltage (volts, r.m.s.)	$V_{h-k(max)}$	200
Grid 1 to Cathode Circuit Resistance (MΩ)	$R_{g1-k(max)}$	1

* The pulse ratings are for Television Line Scan where the applied voltage pulse does not exceed 15% of one scanning cycle or 15 microseconds duration.

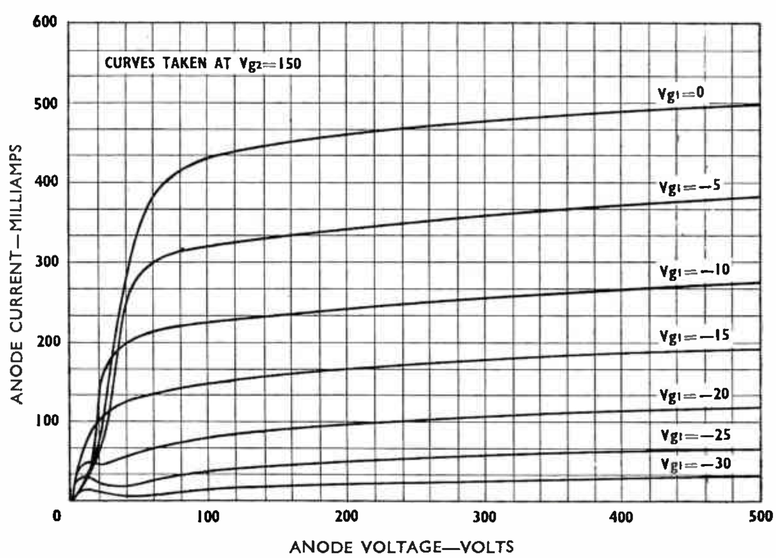
† The absolute rating of 7.5 kV must not be exceeded.

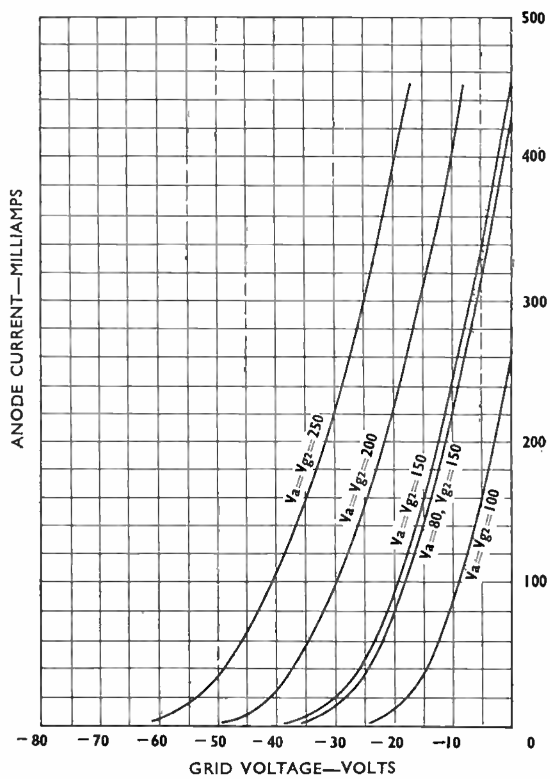
INTER-ELECTRODE CAPACITANCES

Grid 1 to Earth (pF)	c_{in}	20
Anode to Earth (pF)	c_{out}	10
Anode to Grid 1 (pF)	c_{a-g1}	0.3

MAXIMUM DIMENSIONS

Overall Length (mm)	113
Diameter (mm)	32
Seated Height (mm)	99





Characteristic Curves of Average Ediswan Mazda Valve, Type 30P4.

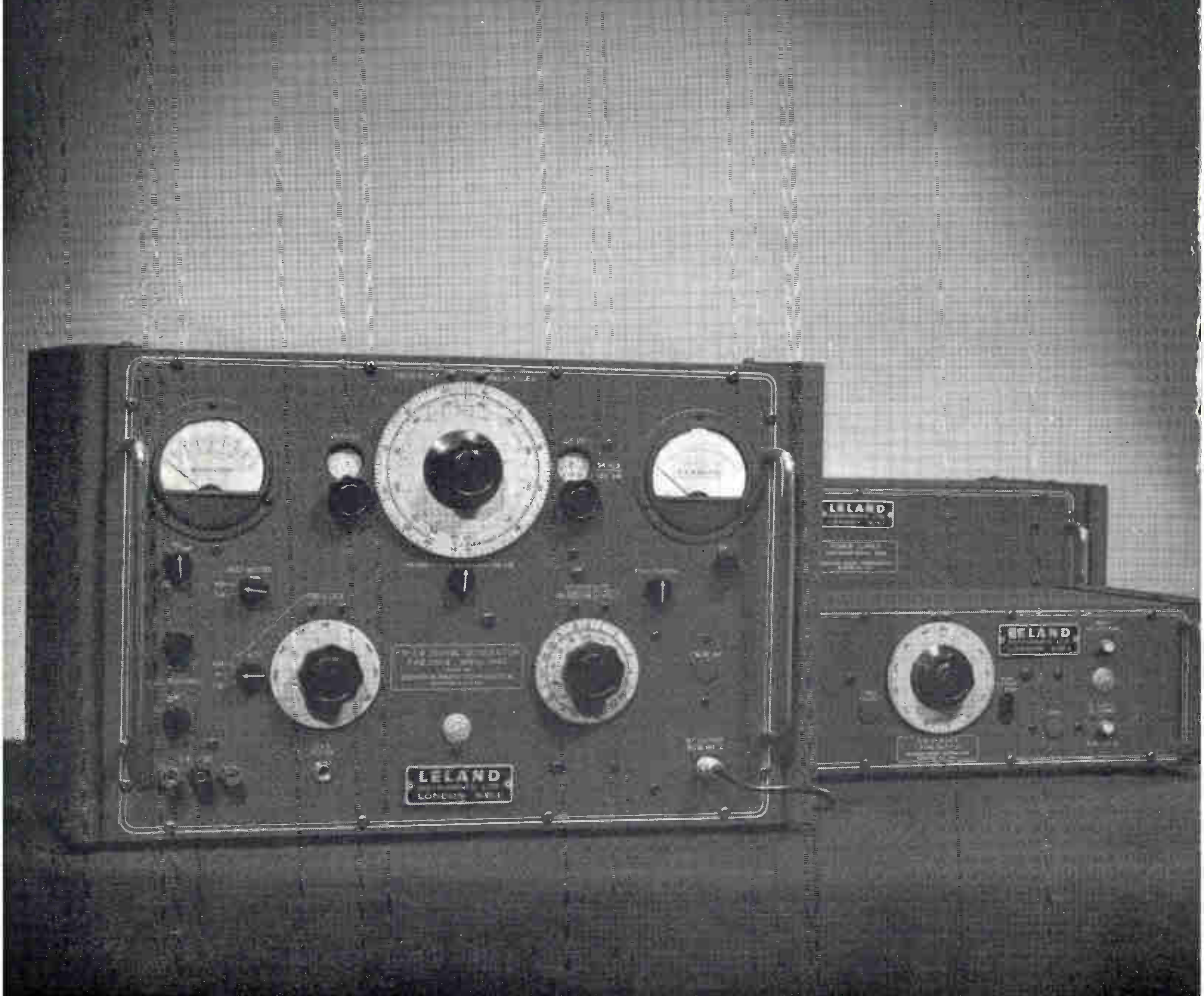
SIEMENS EDISON SWAN LIMITED *An A.E.I. Company*

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

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

EDISWAN MAZDA

CRC 15/35



F.M. *The latest* SIGNAL GENERATOR *by Boonton* — MODEL 202-E

Developed to meet the needs of engineers engaged in the design of F.M. and television receivers for operation within the frequency range 54 to 216 megacycles, this instrument embodies the results of twenty years' experience in F.M. generators, gained in the production of the well-known Boonton 150 and 202-B series.

★ FREQUENCY: 54-216 Mc/s may be extended down to 100 Kc/s by using the Univerter Type 207-E below.

★ R.F. OUTPUT VOLTAGE: Maximum open circuit voltage at front panel jack is approx. 0.4 volts. With output cable attached 0.2 volts nominal. Output impedance 50 ohms resistive at front panel jack. Minimum output 0.1 microvolts.

★ FREQUENCY MODULATION: Three deviation ranges, 0-24 Kc/s, 0-80 Kc/s, and 0-240 Kc/s, each continuously adjustable. FM distortion at 75 Kc/s is less than 2% and at 240 Kc/s less than 10%.

★ FIDELITY CHARACTERISTICS: Deviation sensitivity of the Frequency Modulation system as a function of frequency is flat within ±1 Db from 30 c/s to 200 Kc/s.

★ AMPLITUDE MODULATION: Internal AM is available from zero to 50% with meter calibration points at 30% and 50%. External modulation may be used over the range 0-50%. A front panel jack connects to the screen of the final stage for pulse and square wave modulation.

★ SPECIAL FEATURES: Incremental frequency range: The Δ F switch permits tuning in increments of ±5, ±10, ±15, ±20, ±25, ±30, ±50, ±60 Kc/s in the 108 to 216 Mc/s range — half these values in the 54 to 108 Mc/s range. A fine tuning control permits continuous tuning over a range of approximately ±20 Kc/s in the 108 to 216 Mc/s range, and ±10 Kc/s in the 54 to 108 Mc/s range.

Boonton instruments are available subject to import licence, from the SOLE CONCESSIONAIRES for the United Kingdom—

LELAND INSTRUMENTS LTD.

Electronic Engineers & Consultants

TELEPHONE: VICTORIA 3243 (FIVE LINES)

22 & 23 MILLBANK · WESTMINSTER · LONDON S.W.1