ELECTRONIC & RADIO ENGINEER

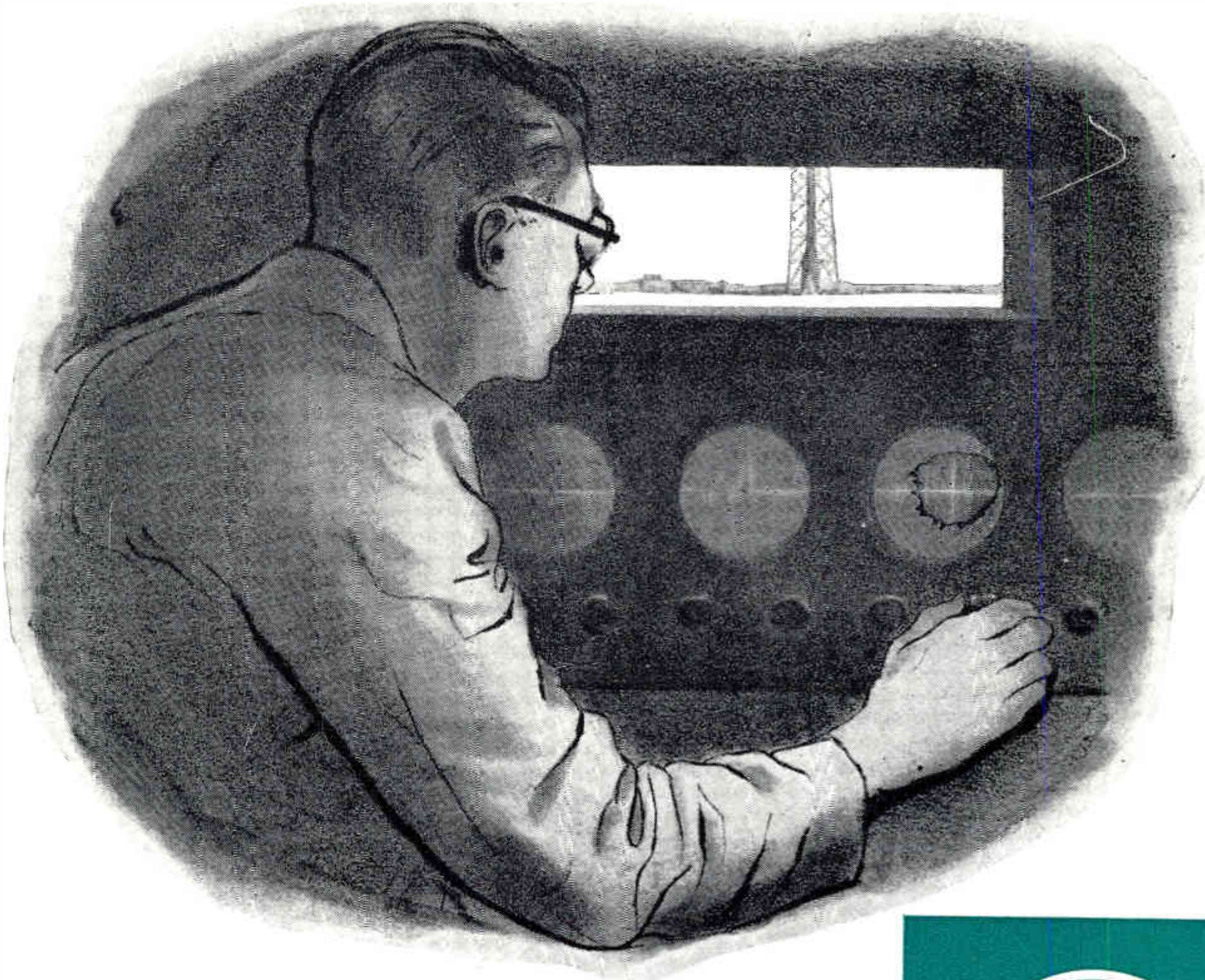
Incorporating **WIRELESS ENGINEER**

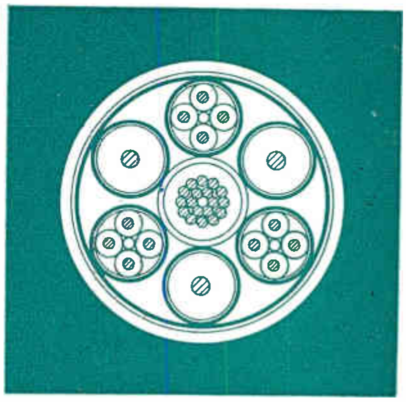
In this issue

Two Short Low-Power Ferrite Duplexers
L.F. Random-Signal Generator
Voltage Standing-Wave Ratio Measurement
Linear Frequency Discriminator

Three shillings
and sixpence

AUGUST 1958 Vol 35 *new series* No 8





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Testing time for guided missiles — with cables providing the nervous system for the control equipment . . . BICC design and manufacture a wide variety of control cables for both ground and airborne use. Standard types are also available for use with ancillary equipment such as ground radar, centimetre radio links and closed circuit television.

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BICC Polypole Couplers are also particularly suitable for use with ground control equipment, since they ensure a tough, permanent, moisture-resistant assembly which virtually eliminates the possibility of conductor breakages at the coupler.

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BICC control cables

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U.H.F. MEASURING EQUIPMENT

Trade Mark

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

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

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

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

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

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

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

BRIEF CHARACTERISTICS

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

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

DIMENSIONS: 7½" × 5½" × 5½".

NET WEIGHT: 8¼ lbs.

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

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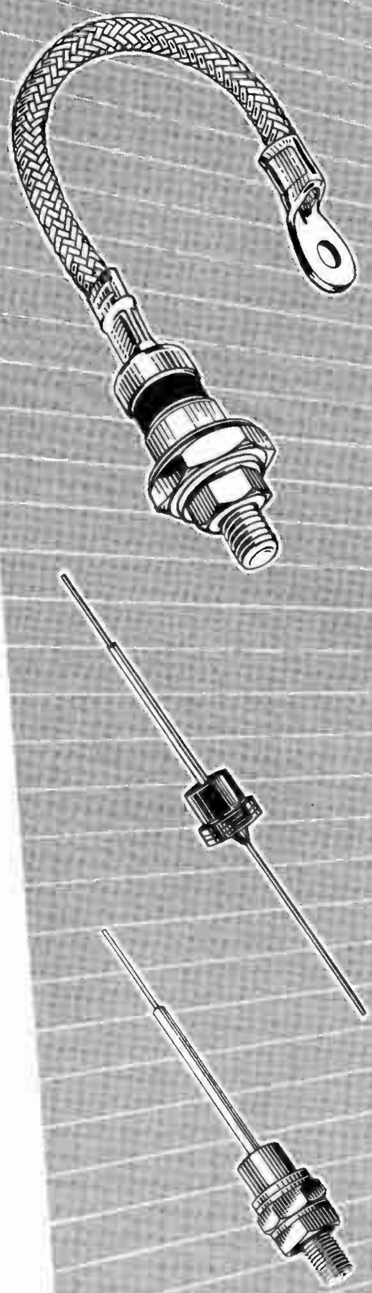

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Ferranti offer a range of hermetically sealed rectifiers with peak inverse voltage ratings up to 300 volts, and rectified current from **100 mA.** to **30 amps.** from a single convection cooled diode.

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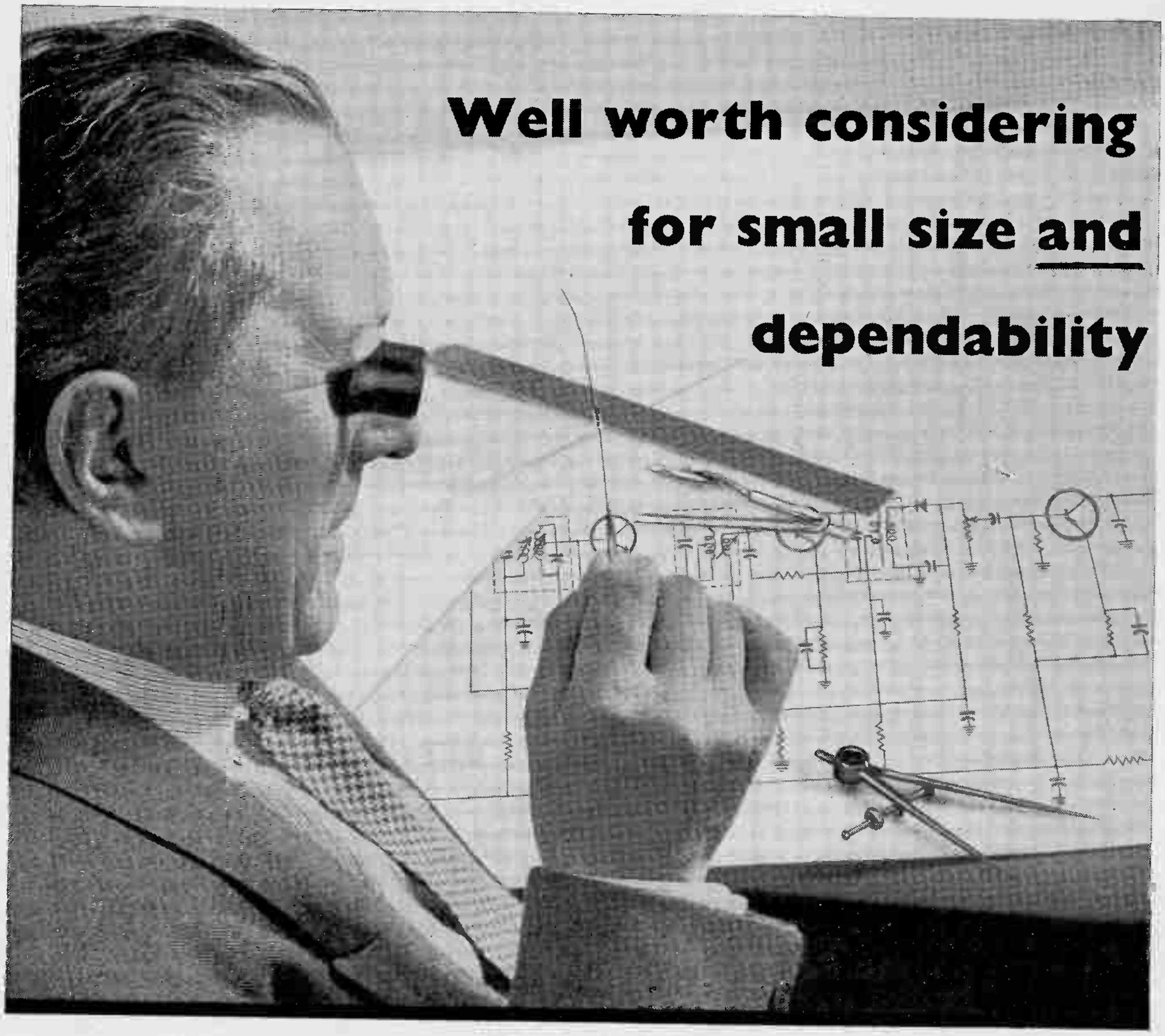

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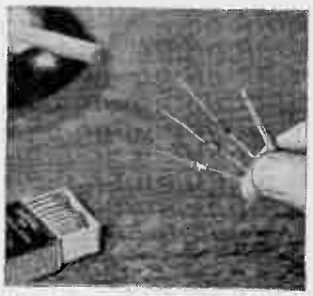
- ABILITY TO WITHSTAND TROPICAL CONDITIONS
- SMALLER DIMENSIONS
- VERY LONG LIFE

RATINGS: CONTINUOUS OPERATION AT 25°C. (77°F.)

TYPE	PEAK INVERSE VOLTAGE† V	MAX. INPUT CURRENT mA	MAX. RESISTANCE at + 1 volt ohms	MIN. RESISTANCE at − 50 volts kilohms
CV 448*	80	30	333	500
CG41–H	65	30	250	50
CG42–H	100	30	500	1,000
CG44–H	80	30	333	500
CG50–H	100	30	500	200

*Type CV 448 has been granted 'type approval'. †Corresponds to 1·2 mA inverse current.



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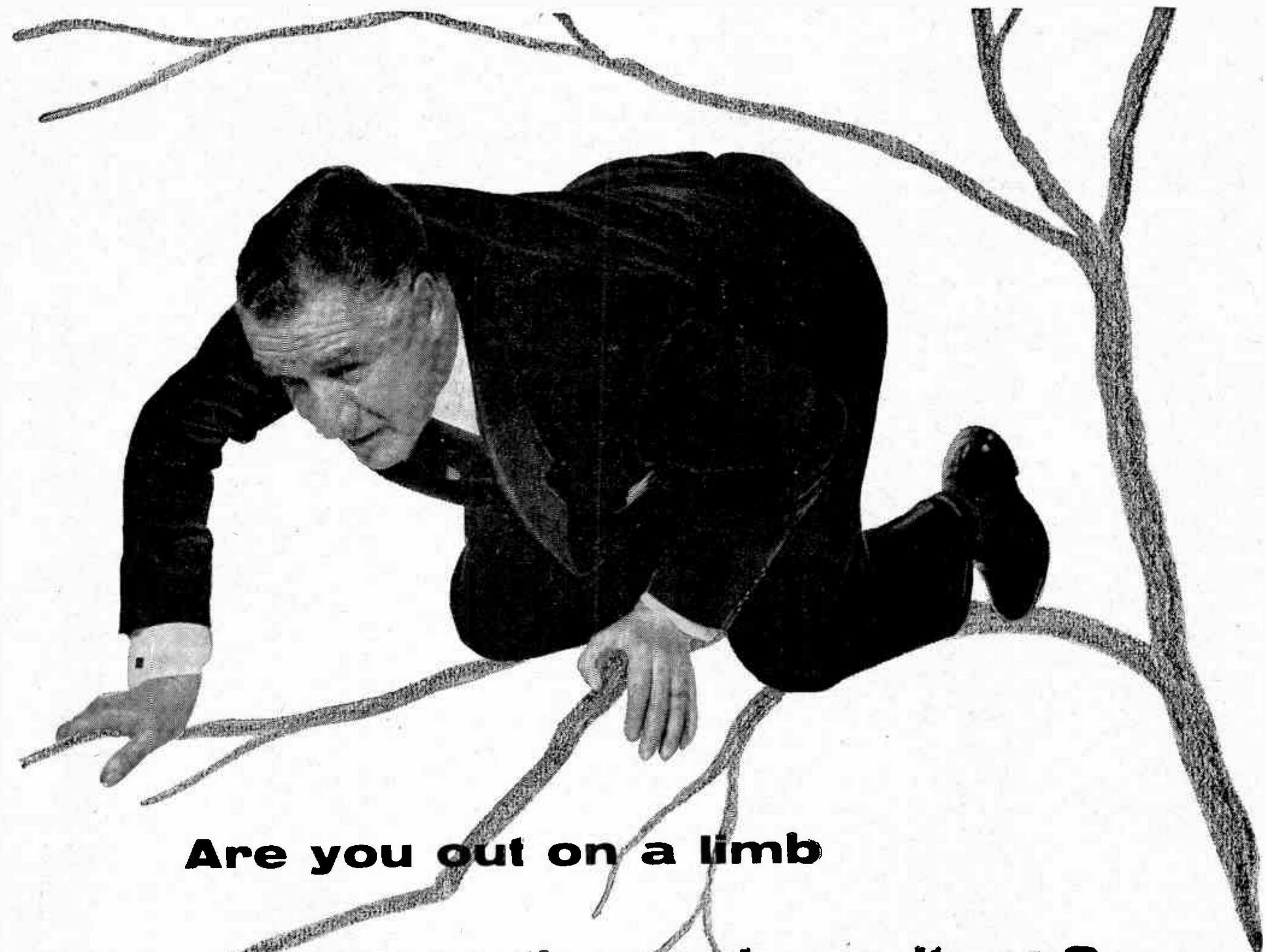

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Technical details available in Leaflet R54.

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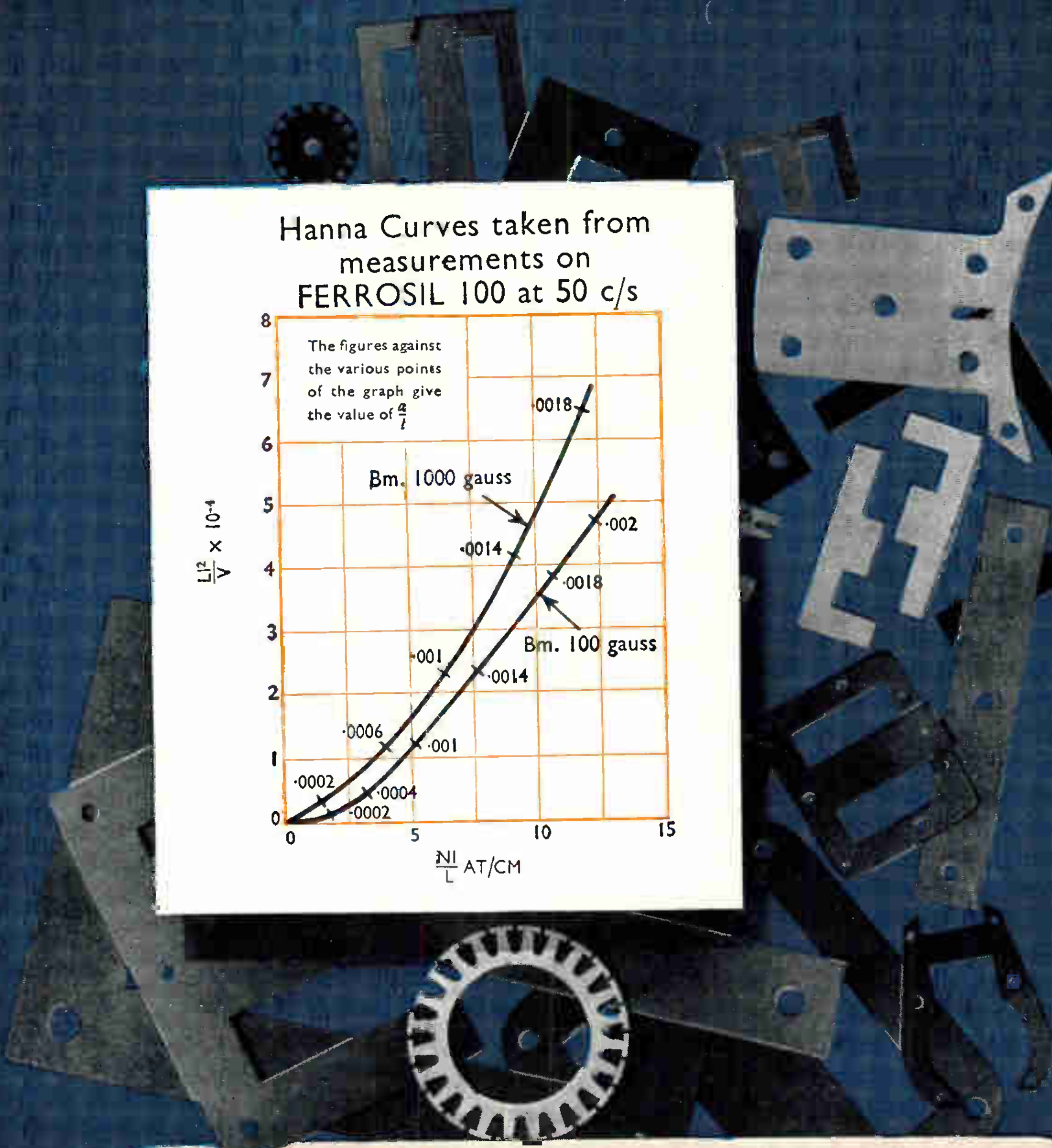
CELESTION

Colaudio

Full technical details available direct from:

Rola Celestion Ltd. THAMES DITTON, SURREY, ENGLAND. Phone: Emberbrook 3402/6



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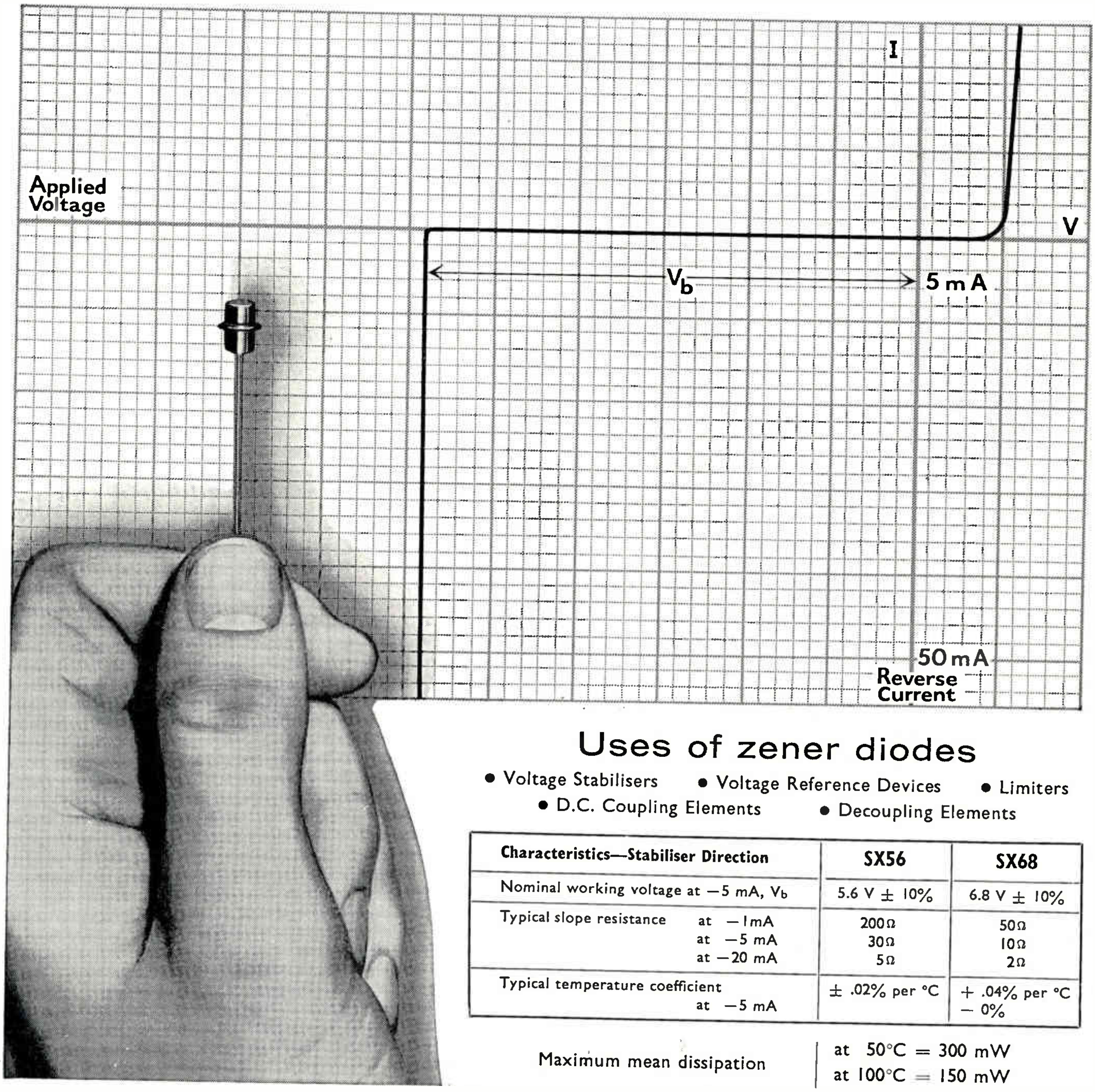
G.E.C.

silicon zener diodes

In production and immediately available

The resistivity of the silicon for these devices has been chosen so that the slope resistance of the reverse characteristic of the diode falls to a very low value when the reverse voltage exceeds a certain value, as shown on the curve.



Uses of zener diodes

- Voltage Stabilisers
- Voltage Reference Devices
- Limiters
- D.C. Coupling Elements
- Decoupling Elements

Characteristics—Stabiliser Direction	SX56	SX68
Nominal working voltage at -5 mA, V_b	5.6 V ± 10%	6.8 V ± 10%
Typical slope resistance at -1 mA	200 Ω	50 Ω
at -5 mA	30 Ω	10 Ω
at -20 mA	5 Ω	2 Ω
Typical temperature coefficient at -5 mA	± .02% per °C	+ .04% per °C − 0%

Maximum mean dissipation | at 50°C = 300 mW
at 100°C = 150 mW

Further information on these devices may be obtained from the G.E.C. Valve and Electronics Department.

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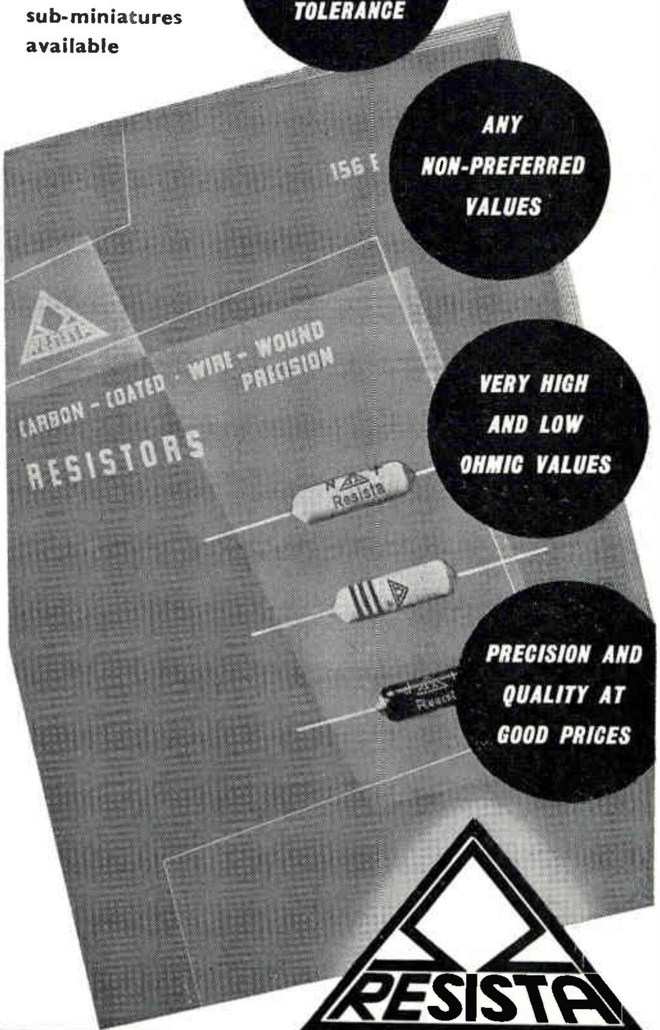

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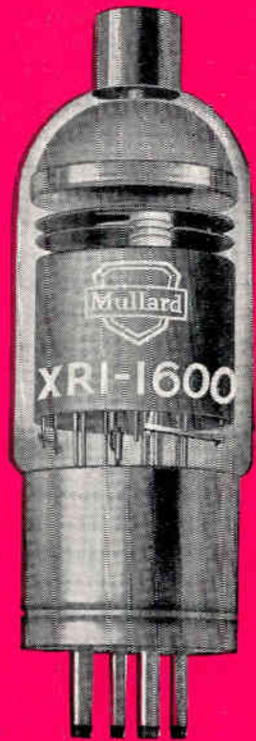

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These thyratrons are designed for the continuous control of currents up to 1.6 amperes.

They are being used to actuate solenoids and electromechanical devices of all kinds as well as for direct electronic switching.

Characteristic advantages of thyratron control are smooth and continuous control of current, negligible control power and high power efficiency.

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PREFERRED RANGE OF SMALL THYRATRONS

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*EN91	2D21	CV797	0.1	1300
EN32	6574	CV2253	0.3	1300
XR1-1600	5796	CV3706	1.6	1500

**Special quality version available.*

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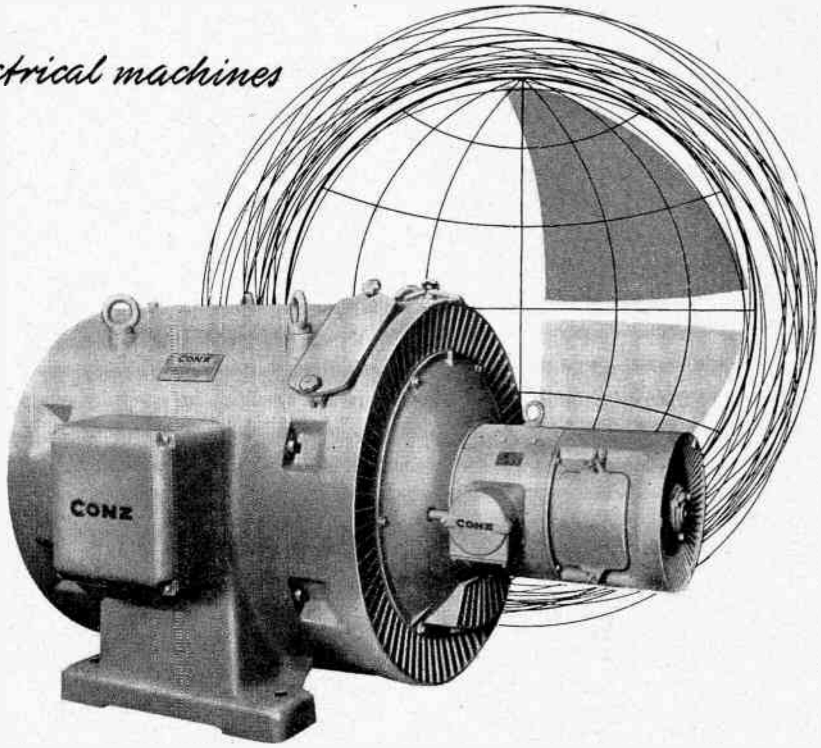

Manufacturing Programme

Threephase Squirrelcage Motors up to 2500 kW
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Converters and Motor-Generator Sets of all kinds
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This work, comprising two volumes, has been specially written for radio and service engineers who wish to acquire a thorough knowledge of television servicing. Both of these volumes have been thoroughly revised and enlarged. In volume I there are new sections on S/C Linearity Coils, Complete C.R.T. Renovation, and Wide-angle Tubes. The section on Flywheel Sync has also been expanded. In addition the section on Instruments has been brought up to date and re-illustrated. The whole of volume II has been revised to bring it up to date and forty-six pages have been added to deal with the new techniques and circuits necessitated by the introduction of alternative programmes. Most of the new material has been added to the chapter on Aerials, which has been enlarged to include Band III aerials and dual-band aerials that receive both sets of programme. Most of the many unconventional and complicated types of aerials and the manner in which they operate are described. There is also a new section on Communal Aerial Systems.

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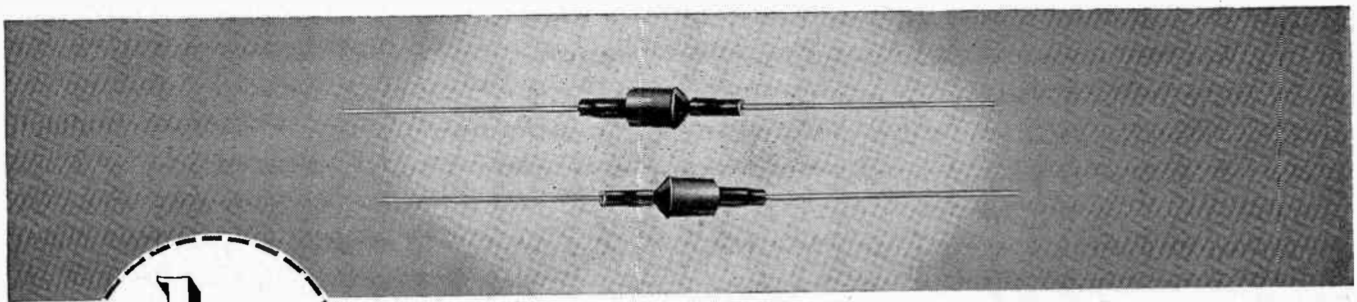

$\frac{1}{2}$ AMP

Mean D.C. at 35°C	½ amp halfwave 1 amp fullwave 1½ amps three phase	**Peak Inverse Voltage in nine grades (−40°C to +100°C)**	RS20A — 50V RS21A — 100V RS22A — 150V RS23A — 200V RS24A — 300V RS25A — 400V RS26A — 500V RS27A — 600V RS28A — 800V
Mean D.C. at 100°C	50% of above		

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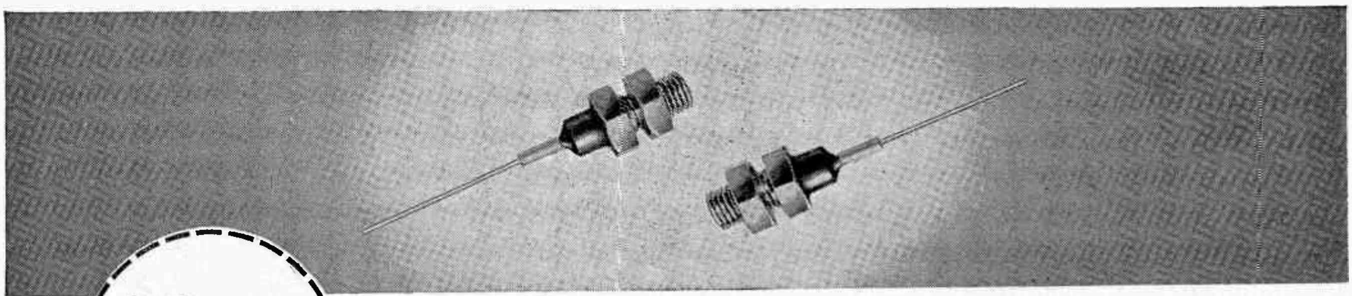

1 AMP

Mean D.C. at 35°C (With 2 sq. in. cooling fin)	1 amp halfwave 2 amps fullwave 2½ amps three phase	**Peak Inverse Voltage in nine grades (−40°C to +100°C)**	RS30A — 50V RS31A — 100V RS32A — 150V RS33A — 200V RS34A — 300V RS35A — 400V RS36A — 500V RS37A — 600V RS38A — 800V
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VACUUM - TUBE VOLTMETER TYPE TF 1300

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MEASURES D.C. up to 300 V

MEASURES OHMS up to 5MΩ

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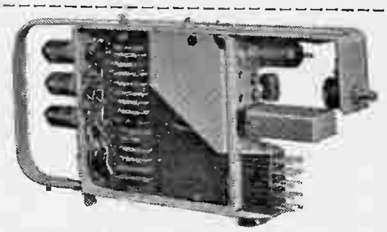

One of the three *printed circuit* drift-corrected plug-in amplifiers used in the Oscillator.

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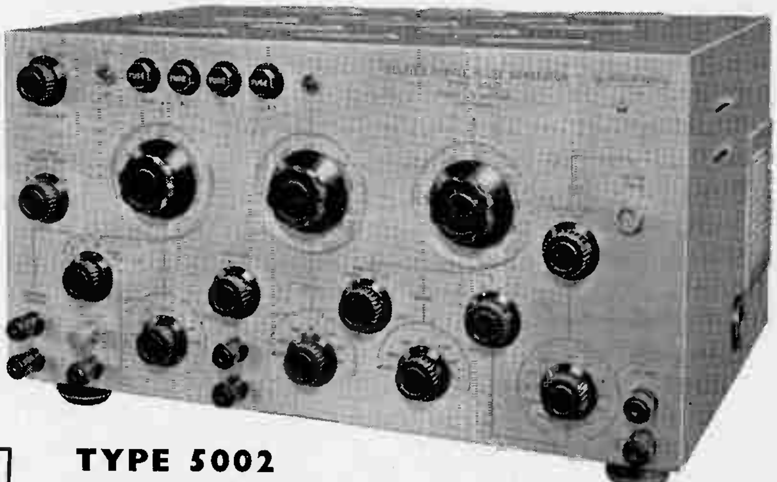

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Rise Time—Optimum	Square Wave Output	Internal Multivibrator
10 mμsecs independent of pulse width.	0·25c/s to 2·5 Mc/s 10 million to 1 range of pulse rate, width and delay calibrated by direct reading dials within 5%.	of new type gives highly stable frequencies continuously variable from 0·1 c/s to 1 Mc/s.

TYPE 5002

Providing for either single pulses or two consecutive pulses spaced by the delay control

External Trigger	Delay from Pre-Pulse	Single Cycle	Main Pulse
from signals of any waveform and polarity at amplitudes down to 0·2 V and frequencies up to 2·5 Mc/s.	to main pulse—jitter free and continuously variable from 0·2 μsec to 2 secs. Pre-pulse output 20 volts positive or negative—0·2 μsec wide.	of a single pulse or a pair of pulses at the width and delay setting—by push button or external contacts.	amplitude 20 mV to 50 V—of either polarity and within 2% accuracy in 10 steps.

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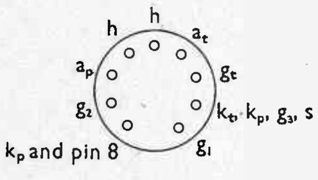
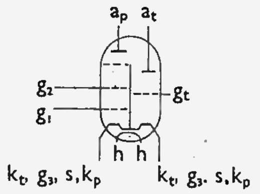
Introducing another outstanding Ediswan Mazda valve, type 30C13

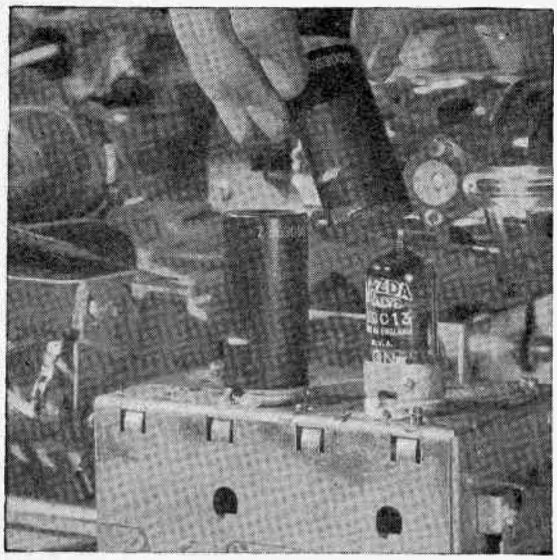
For the information of set designers we are publishing details of individual 0.3 amp heater valves in our 'First Preference' Range for TV circuits. If you are a TV manufacturer we shall be pleased to supply full technical details of our 'First Preference' Range, together with a set of valves for testing, on receipt of your enquiry. The valve dealt with here is the Type 30C13, a VHF Frequency changer for printed circuit tuners.

Heater volts - - - -	9.0
Heater current - - -	0.3 amps
Maximum anode volts	
Pentode - -	250
Triode - -	250
Maximum screen volts -	175

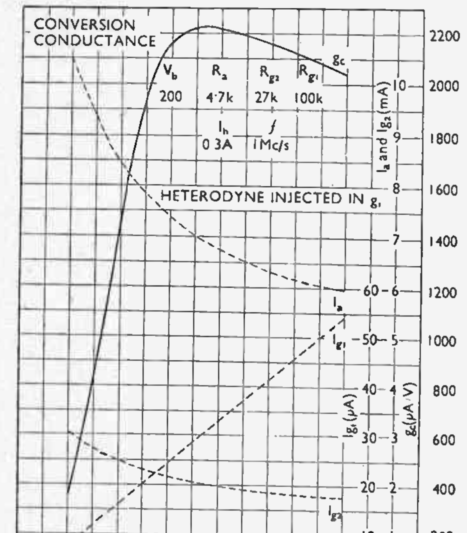


B9A base





Characteristic curves of average Ediswan Mazda Valve type 30C13



TYPICAL OPERATION

As frequency changer with oscillator volts applied to g_1.

PENTODE	
Anode volts	170
Screen volts	150
g_1 Resistance for grid current bias...	100,000 ohms
g_1 Current	42 μA
Conversion conductance	2150 μA/V
Heterodyne peak volts	4.5
Anode current	6.3mA
Screen current	1.9mA
TRIODE	
Anode volts	120
Anode current	6.0 mA (average)

CAPACITANCES

PENTODE		TRIODE	
g_1 to all	6.3 pF	g to E	3.5 pF
ap to all	5.2 pF	a to E	3.3 pF
g_1 to ap	0.016 pF	g to a	1.7 pF

APPLICATION USE OF THE 30C13

The performance of a television tuner on Band III is often limited by the comparatively low input resistance of the triode pentode mixer that is commonly used. One of the principal causes of low input resistance at these frequencies is the cathode lead inductance, part of which is in the valve and part in the external circuit. This inductance causes degeneration and produces resistive damping of the input circuit.

With the use of printed circuit technique the external cathode lead inductance is very low making it more important than ever to reduce to a minimum the cathode lead inductance in the valve. In the 30C13 this has been done by strapping together the pentode cathode and the triode cathode internally with a low inductance connection thus practically halving the cathode lead inductance in the valve base. In addition the disposition of the pins in the base has been chosen so that with the 30C13 the fullest use can be made of the opportunity of providing the best circuit layout for printed circuit techniques when used in conjunction with a cascode valve like the 30L1.

The use of the 30C13 in a printed circuit will result in a gain increase on Band III of 1½ dB compared to the 30C1 used in a wired tuner.

The base connections of the 30C13 provide the following advantages for printed circuit use:

(*a*) The pentode g_1 and cathode are brought out on adjacent pins. This enables the grid trimmer to be placed very close to the g_1 to k capacitance thus minimising errors in alignment that can arise at differing frequencies if the trimmer has series inductance.

(*b*) The g_3 connection is conveniently placed close to the cathode.

(*c*) The heater pins are easily accessible for series connection in a printed circuit board while still allowing easy decoupling to the strapped cathodes.

(*d*) The position of the grid and anode pins of the triode oscillator makes it possible to use short connections to the oscillator coil.

(*e*) The reduction in cathode lead inductance increases the gain on Band III.

These points are illustrated in Fig. 1 which shows part of a printed circuit layout using the 30C13 where the R.F. stage is assumed to be a cascode amplifier using the 30L1.

If the 30C1 were used in place of the 30C13 a much less satisfactory layout would result. This is shown in Fig. 2 where the g_1 circuit has a longer return path to cathode with the added disadvantage that this is shared by the g_2 circuit and for these reasons the 30C1 should not be considered for this application.

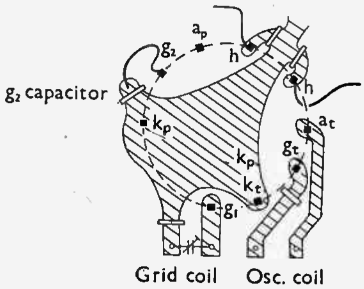


Fig 1 30C13

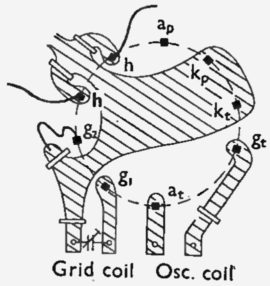


Fig 2 30C1

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ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

CONTENTS

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Permanent Magnets

Design Advisory Service

No. 5

Symbols and Definitions

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.

A_g Cross sectional area of air gap.

A_m Cross sectional area of magnet.

B Magnetic Flux Density; the number of lines of flux per unit area in a section normal to the direction of flux. The c.g.s. unit is the maxwell or line. The M.K.S. unit is the weber.

B_d Value of B on the major hysteresis loop, or demagnetisation curve, corresponding to a given value of $-H$ (denoted H_d), usually at $(BH)_{max}$.

B_g The flux density in the air gap.

B_r Remanence or residual induction; the flux density which remains in the material (a closed or short-circuited specimen) after being magnetised to saturation.

B_{sat} Saturation flux density.

F or M.M.F. Magnetomotive force. That force which produces or tends to produce a magnetic flux.

The c.g.s. unit is the gilbert.

$$1 \text{ gilbert} = \frac{10}{4\pi} \text{ ampere turns.}$$

The M.K.S. unit is the ampere turn (the product of turns and amperes in a magnetising coil).

$$1 \text{ ampere turn} = \frac{4\pi}{10} \text{ gilberts.}$$

H Magnetising force.

The c.g.s. unit is the oersted.

$$\text{Magnetising force} = \frac{\text{M.M.F. (in gilberts)}}{\text{length (in cm.)}}$$

The M.K.S. unit force is the ampere-turn per metre.

$$\text{Magnetising force} = \frac{\text{M.M.F. (ampere turns)}}{\text{length (in metres)}}$$

H_c Coercive force; the magnetising force applied to a fully magnetised magnet to reduce the residual (B_r) to zero.

H_d Value of $-H$ on the major hysteresis loop or demagnetisation curve corresponding to a given value of B (denoted B_d) usually at $(BH)_{max}$.

H_{sat} The value of H corresponding to B_{sat}.

L_g Length of air gap.

L_m Effective magnetic length of magnet.

μ Permeability; the ratio of the magnetic flux density induced in a given medium to that which would be produced in vacuum by the same magnetising force.

Φ Total flux over a cross-section, i.e. the integral of B over the cross-section.

CONVERSION TABLE c.g.s. TO M.K.S. UNITS

		c.g.s.		M.K.S.
Magnetomotive force	F	$\frac{4\pi}{10}$ gilberts	=	1 ampere-turn
Magnetising force	H	$\frac{4\pi}{10^3}$ oersteds	=	1 ampere-turn/metre
Total Flux	Φ	10^8 maxwells	=	1 weber
Flux Density	B	10^4 gauss	=	1 weber/sq. metre

Reprints of advertisements in this series will be gladly sent on application to the address below.

Advertisements that have appeared to date are:—

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VOLUME 35 NUMBER 8 AUGUST 1958

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Education

THE increasing complexity of electronic apparatus is naturally enough making the training of engineers and technicians more and more difficult. The older ones among us have grown up with the development of radio and electronics and have acquired their knowledge of the subject in the easy and natural way. With a basic foundation of physics and electricity and magnetism they have built up their knowledge step by step over many years.

Some people may call this the hard way of finding out for oneself by individual research and hard thinking. We call it the easy way, because it is the interesting way. There is no set course with an examination at the end!

Instead of taking twenty-five to fifty years to acquire his knowledge the student today must do it in a few years' course. There is no escaping this. If progress is to continue each new generation must acquire the technical knowledge of the previous one more quickly.

The way of doing this is by more generalized theory so that instead of having to learn separately a number of apparently unrelated items the student has only one basic thing to learn which includes as special cases all individual items. Unfortunately, it is the individual items which generally hold the interest more than the general, for they usually appear to the student to be more closely related to real life. The general is more hard to grasp and to apply.

That is perhaps why most people find mathematics so difficult. It is so general that it can almost be said that it is necessary only to know mathematics and how to translate physical problems into and out of its language. It does appear that the most important educational problem is that of finding out how to teach mathematics so that the student can become proficient much more readily than at present. The answer probably lies in discovering how to arouse his interest in it.

Two Short Low-Power Ferrite Duplexers

By R. S. Cole, B.Sc., Grad. Inst. P.* and W. N. Honeyman, B.Sc.*

SUMMARY. *Two short ferrite duplexers which are suitable for low-power Doppler navigational systems in X-band are described, and their performances given.*

Until comparatively recently, radar systems have relied for the operation of duplexing on the conventional gas-discharge tubes which are fired by the microwave pulse. It is only in the past few years that the need for alternatives has arisen owing to the advent of c.w. low-power Doppler navigational radars where normal duplexing systems cannot be used. The introduction of ferrites has provided an ideal solution to this problem.

In this article, two types of low-power ferrite duplexer suitable for Doppler navigational systems are described and their performance details given. To distinguish between the two types they will be called the Faraday rotation and the turnstile duplexers although both employ 45° Faraday rotators. Both duplexers are designed to be as short as possible and transmitter-receiver isolations of the order of 25 dB have been obtained over bandwidths of greater than 1·3% in X-band. These performances are quite adequate for low-power systems of about 50 watts mean.

Principles of the Two Duplexing Systems

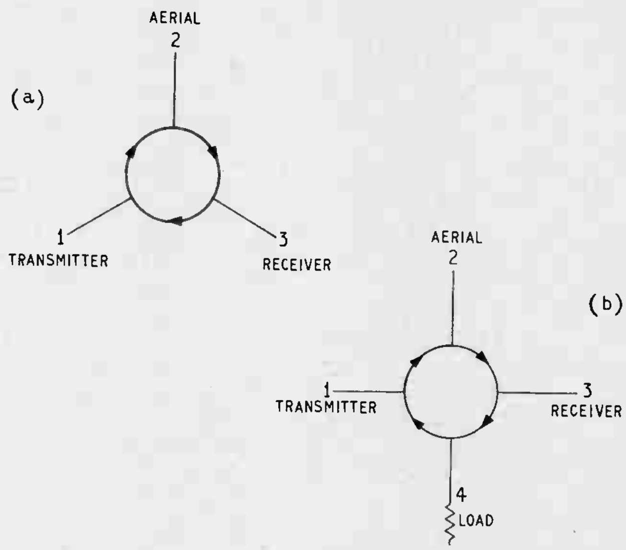
Both types of duplexer are essentially circulators in their action and are shown symbolically in Fig. 1 (a) and (b). The four-port circulator in Fig. 1 (b) shows the principle of the turnstile duplexer in which the fourth arm is terminated in a matched load. It is also possible to use a four-port circulator with the Faraday rotation duplexer and, in this case, the fourth arm would again be terminated in a matched load.

The Faraday rotation-type duplexer described is shown diagrammatically in Fig. 2, where the vector diagram is given to explain the basic mechanism of the device. Essentially, two rectangular–circular transformers, one circular to rectangular T-junction and a 45° Faraday rotator are arranged as shown, where arms one and two are placed at 45° to each other. Transmitted power from the magnetron entering arm 1 sets up a TE_{11} mode in the circular guide with the electric-field vector in the direction shown and is not coupled to arm 3 which is cut off to electric fields in this direction. It then undergoes a 45° rotation in the ferrite, is transformed back into rectangular guide and emerges from arm 2 to the aerial. The received power enters arm 2, and its electric-field vector undergoes a 45° rotation in the same direction as the transmitted vector; this is a property of the Faraday rotator. Consequently, it is cut off in arm 1 and enters arm 3 into the receiver.

The turnstile-type duplexer uses a turnstile junction[1] and a 45° Faraday rotator. The turnstile junction, shown diagrammatically in Fig. 3, consists of two crossed sections of rectangular waveguide with circular waveguide entering the part which is common to both sections. If the junction is correctly matched, power transmitted into arm 1 will divide so that a quarter of the power goes into each of arms 2 and 4. The remaining half sets up a TE_{11} mode in the circular guide (arm 5) with the electric field in the direction A, and no power is coupled into arm 3. Power transmitted in arm 5 with the electric-field vector in the direction B is divided equally but 180° out of phase between arms 2 and 4.

If a 45° Faraday rotator is placed in the circular arm backed by a short circuit then the component of the incident power entering the arm is rotated 45° by the ferrite, reflected at the short circuit and rotated through

Fig. 1. (*a*) *Three-port circulator;* (*b*) *four-port circulator.*



* *Microwave Division, E.M.I. Electronics Ltd., Feltham, Middlesex.*

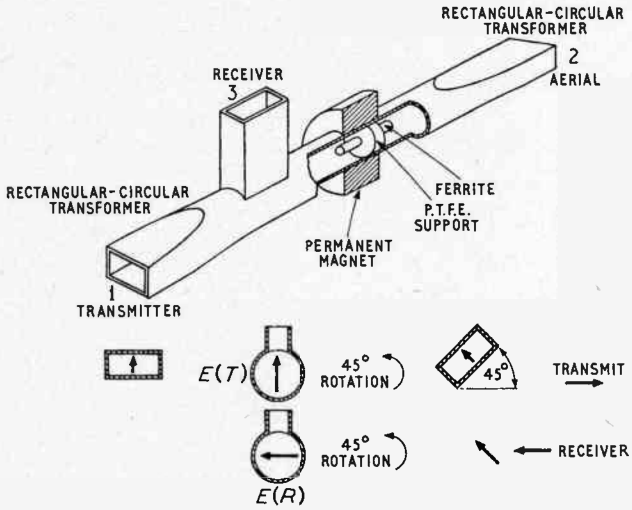
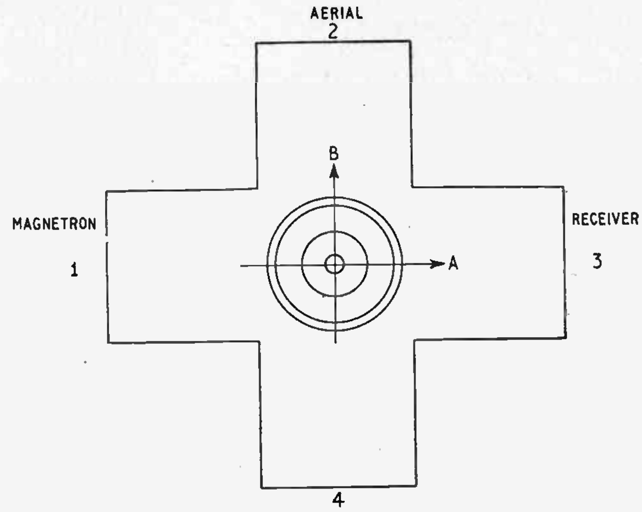


Fig. 2. *Faraday rotation-type duplexer*



a further 45° in the same direction. Thus, power entering with its electric vector in the direction A returns with its electric vector in the direction B and then divides equally between arms 2 and 4. By adjusting the position of the short circuit this reflected voltage in arm 2 can be adjusted to be in phase with the original 50% of the incident voltage (25% of power) in that arm and these then add together while the reflected voltage in arm 4 is out of phase and cancels. Hence, power flows from arm 1 to 2 and since the junction is symmetrical from 2 to 3, etc.

Design of the Faraday Rotator Duplexer

In the construction of this ferrite duplexer, the following components were designed as separate units:

(a) Rectangular–circular transformers;
(b) Circular–rectangular T-junction;
(c) Ferrite assembly.

In order to make a compact duplexer, these component parts could not be separated to allow fringing fields to decay and so the interaction between components was examined in two stages:

(a) Ferrite–T-junction interaction;
(b) Ferrite–transformer interaction.

The v.s.w.r. of all the components must be good, since any reflections from the end of the ferrite or the second transformer travel back into the receiver and reduce the transmitter-receiver isolation. For example, a combined mismatch of 0·8 from these parts gives a reflection 19 dB down on the transmitted power which would mean that the transmitter-receiver isolation would not be better than this value.

Rectangular–Circular Transformers

The diameter of the circular guide was chosen so that only the dominant TE_{11} mode could be transmitted down the air-filled guide. In one particular application, the transmitter and receiver flanges were required to be in line instead of at 45° as shown in Fig. 2, and so the two transformers were designed to incorporate a 22½°

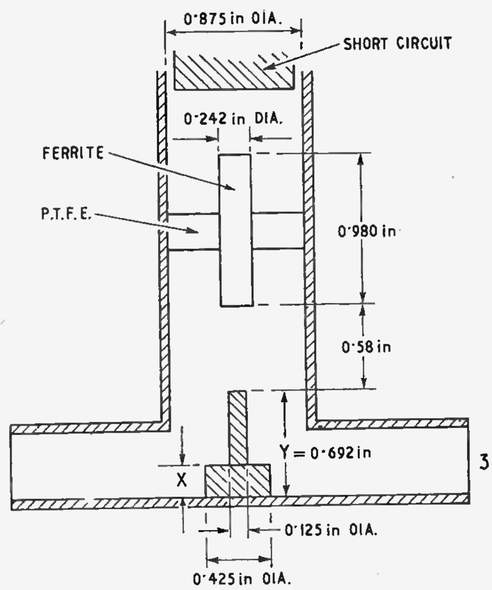


Fig. 3. *Turnstile duplexer showing dimensions of matching devices*

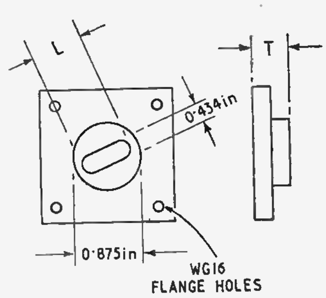


Fig. 4. *Rectangular–circular waveguide transformer*

twist (Fig. 4). The design of this component involved a variation of the two dimensions L and T and it was found that they form a set of orthogonal parameters when plotted on an impedance chart. Hence, a perfect match can be achieved at one frequency and, by designing to the interpolated dimensions, a v.s.w.r. of 0·98 was obtained at the centre frequency. This transformer serves a second purpose in providing a short-circuit to the electric-field vector of the received signal when it arrives back at the first transformer at 90° to its transmit direction. This short-circuit must be positioned to match the received power around the T-junction into arm 3.

Circular–Rectangular T-Junction

The properties required of this junction can be summarized with reference to Fig. 5:

(1) Electric fields in the direction $E(T)$ are required to be transmitted past the rectangular guide with minimum coupling to it.

(2) Electric fields in the direction $E(R)$ are required to couple into the rectangular guide with minimum loss.

A piece of WG.16 (internal dimensions 0·900 in. × 0·400 in.) soldered into the circular guide was satisfactory when two matching posts A and B were used; the post A matched the $E(T)$ field vector past the junction without affecting $E(R)$. To match the $E(R)$ vector through the junction into the rectangular guide, a short circuit must be provided at the best position in the circular guide when, with the post B, a perfect match can be obtained. This short circuit is provided by the first transformer. The coupling of the $E(T)$ field vector into the rectangular guide was found to be 40 dB down on the transmitted signal.

Ferrite Assembly

This is a simple Faraday rotator giving a 45° rotation of the electric field while being a good match so that reflections from it do not enter the receiver. General Ceramics R1 ferrite was chosen because variations of rotation with temperature are small[2], the ferrite being supported in p.t.f.e. which will withstand an operational temperature of 120°C. The length of the ferrite and the thickness of the p.t.f.e. were varied and the magnetic field was adjusted for 45° rotation as each parameter was altered until a match of 0·97 was obtained at the centre frequency and this configuration was then used. On the experimental model an electro-magnet was used to facilitate variation of the magnetic field but, in production, this has been replaced by a permanent magnet. In practice, the magnetic field could be varied up to 10% without materially affecting the performance of the duplexer.

Reaction between Components

In one particular application, the shortest possible duplexer was required and, consequently, large spaces between components could not be tolerated. Before the short duplexer was made, a long one was constructed and the effects of component proximities were investigated separately. The spacing between the ferrite and the T-junction did not affect its performance and it can project into the junction up to a quarter of an inch without any marked deterioration. Any further attempt to save space by this method was limited by a higher aerial–receiver insertion loss. The ferrite-to-transformer spacing could not be reduced below 0·2 in. without seriously affecting the overall performance. This was attributed to interaction between fringing fields of the components affecting their combined performance. It should be possible to overcome this by designing the two components together, but efforts in this direction met with little success and so spacing was adopted.

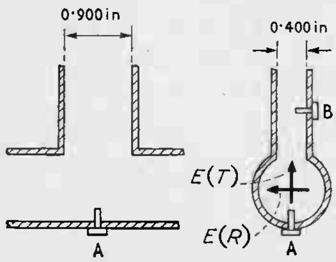


Fig. 5. *Diagram showing positions of stubs and directions of electric vectors*

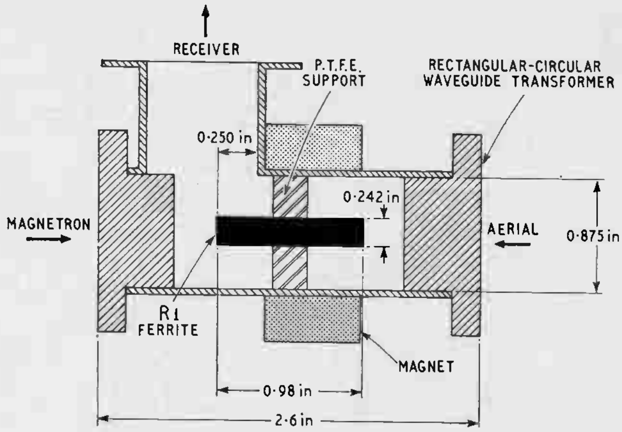


Fig. 6. *Schematic layout of Faraday rotator duplexer*

Performance

The final duplexer which is shown in Fig. 6 is 2·6 in. long and uses a ring magnet giving a field of about 60 oersteds. A magnetron–receiver isolation greater than 25 dB over a bandwidth of 1·8% and 30 dB over 1·1% was obtained. The insertion losses between magnetron and aerial, and aerial and receiver, were less than 0·25 dB as shown graphically in Fig. 7, while the v.s.w.r. at the magnetron port measured with matched loads on the other arms was greater than 0·8 over these frequency bands. The performance of the duplexer was checked over the temperature range of 20°C to 120°C; changes of up to 1 dB in the isolation and 0·05 dB in insertion loss were found.

Design of Turnstile Duplexer

In the construction of this duplexer[3] the following components were designed as separate units:

(a) Turnstile junction;

(b) Ferrite assembly.

In order to keep the device as compact as possible, interaction between the ferrite assembly and matching post of the turnstile junction was examined.

Components

The properties of the turnstile junction described above rely on the correct matching of the junction. This was obtained by means of a symmetrical matching structure with two variable dimensions X and Y as shown in Fig. 3. It was found that a perfect match looking into any one rectangular arm did not necessarily

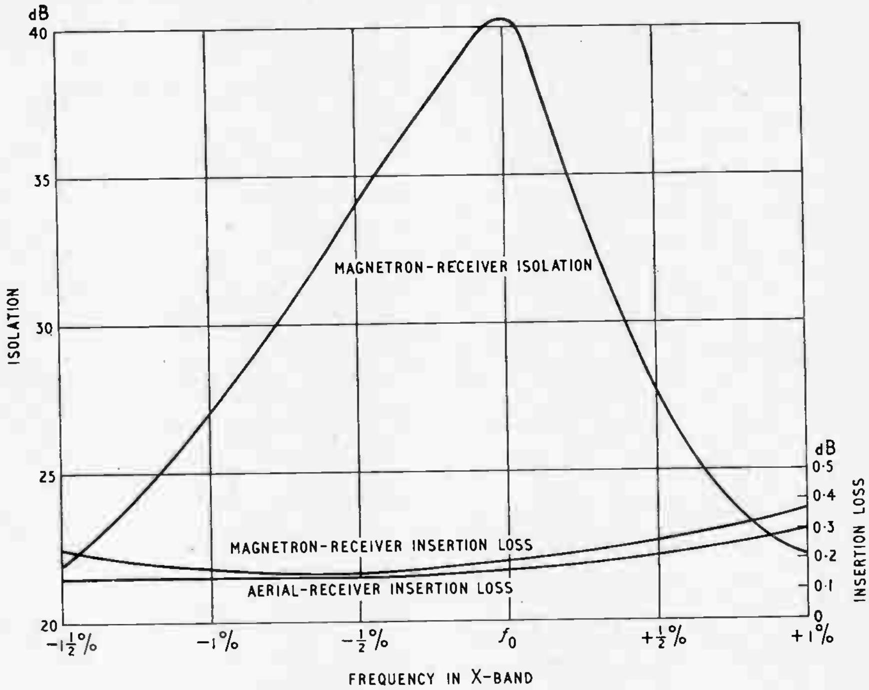


Fig. 7. *Performance of the Faraday rotation duplexer*

mean a perfect match looking into the circular arm, and only one combination of X and Y would ensure a good match in both simultaneously. In practice a v.s.w.r. of 0·97 in any rectangular arm and 0·93 in the circular arm was obtained.

Since this duplexer was designed at about the same frequency as the previous device the same ferrite assembly was used.

Assembly

To set up the duplexer the junction, Faraday rotator and a variable short circuit were assembled as in Fig. 3. The rotation of the ferrite device was adjusted to be 45° by observing a minimum in the cross coupling between arms 1 and 3. The position of the short circuit was then adjusted to give minimum reverse coupling between arms 1 and 4. Under these conditions the magnetron–aerial insertion loss was found to depend on the spacing between the ferrite assembly and the junction matching post. Typical variations in this insertion loss were 0·4 dB for a distance of 0·5 in. between the end of the ferrite and the end of the matching post and 2·1 dB for a spacing of 0·35 in. The bandwidth of this type of duplexer depends upon the length of the ferrite loaded cavity and for this reason it is essential to keep any spacings as short as possible.

Performance

The final duplexer which is shown in Fig. 3 could be made $1\frac{1}{4}$ inches long, with a magnetron–receiver isolation

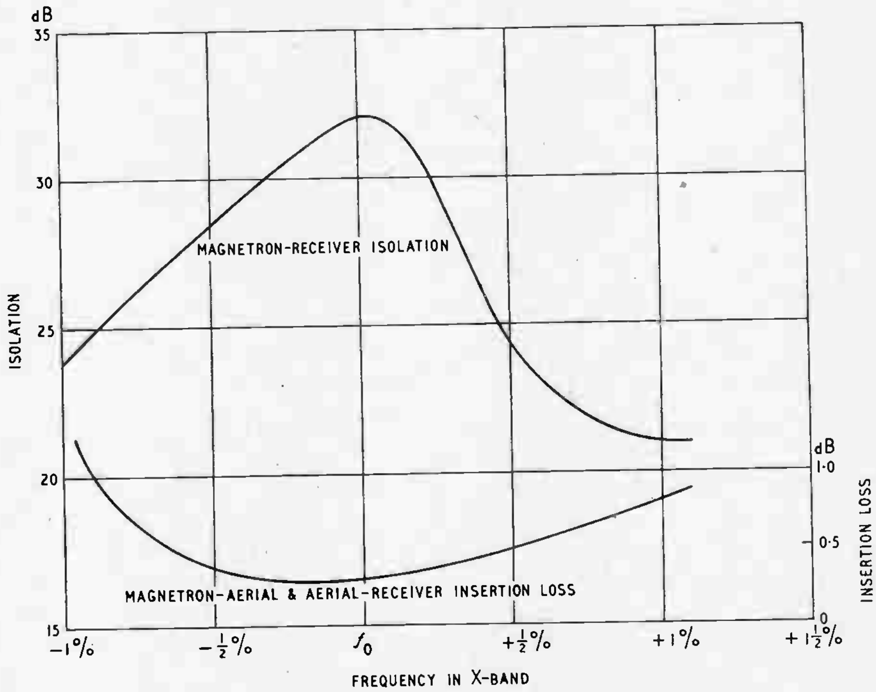
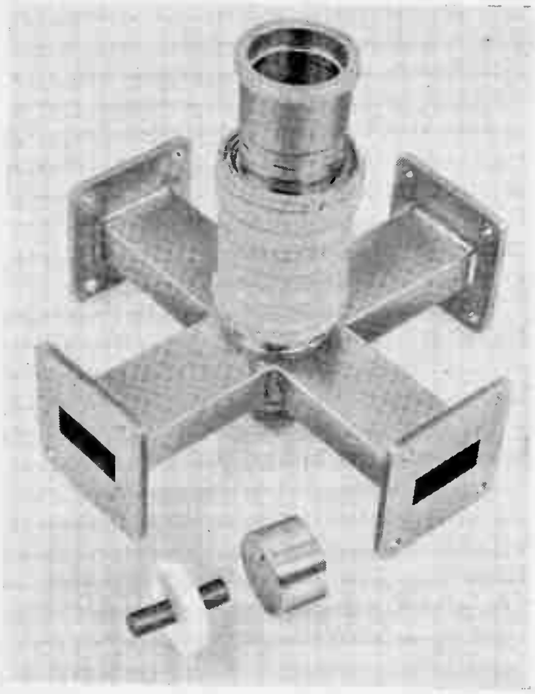


Fig. 8. *Performance of the turnstile duplexer*



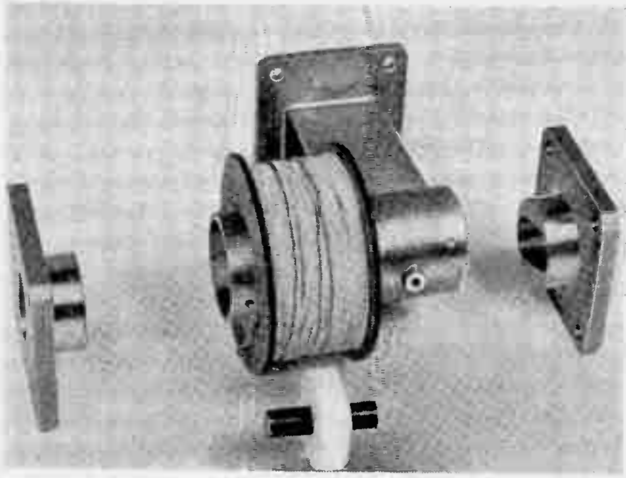


Fig. 9. *The Faraday rotation duplexer*

Fig. 10. (*left*) *The turnstile duplexer*

of greater than 25 dB over a bandwidth of 1·3%. The magnetron–aerial and aerial–receiver insertion losses in this duplexer were kept below 0·45 dB over this band as shown graphically in Fig. 8 while the v.s.w.r. at the magnetron port measured with matched loads in the other arms was greater than 0·8 over the same bandwidth.

Conclusion

A comparison between the performance of the two types of duplexer shows that the Faraday rotation type appears to have a larger bandwidth than the turnstile duplexer, although the latter could no doubt be improved by reducing the length of the circular cavity still further. If a short length between the magnetron and aerial ports is required, then the turnstile duplexer has the advantage that it would only occupy about 1¼ inches, whereas the Faraday rotation type described is 2½ inches long. However, although the turnstile circulator described is only an interim design, it can be seen that both types can be used in low-power narrow-band c.w. Doppler navigational systems.

Acknowledgements

The authors wish to express their thanks to E.M.I. Electronics Ltd., Feltham, Middlesex, for permission to publish this article.

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B.B.C. COLOUR TELEVISION TESTS

The B.B.C. has issued a report on the series of colour television tests carried out by its staff, in co-operation with the radio industry, up to the spring of 1957. It describes the results as promising, though further experimental work will be needed to resolve outstanding problems mainly concerning the design of equipment. The B.B.C. has formed the opinion that the system used (an adaptation of the American N.T.S.C. system) is capable of giving acceptable results. The report forms B.B.C. Engineering Division Monograph No. 18: "The B.B.C. Colour Television Tests: An Appraisal of Results".

Since these tests were completed a further series has been carried out, the results of which are now being analysed.

The field trials on which the B.B.C. now reports extended over a total period of seventeen months and involved the analysis of over a thousand questionnaires, filled in by members of the B.B.C. staff and others who watched the results on special sets.

So far as reception of the colour picture is concerned, the B.B.C.'s verdict is that the colour pictures produced by the N.T.S.C. system as adapted to 405 lines, with the picture sources and display tubes that are at present available, are satisfactory. The problem of registration with the three-tube colour camera requires special attention and certain features of the display tube could with advantage be improved, but the technical performance of the system is adequate for a satisfactory colour television service in the frequency bands at present in use. The tests radiated from Crystal Palace from November 1956 to April 1957 included live studio performances, film and slides. Of the observers, 89% regarded the reception of the live scenes as satisfactory. The 35-mm film was regarded as satisfactory by 100% of the observers, and the 16-mm film by 93%, and the slides by 98%.

The quality of ordinary black and white television, when received on a colour television set, was found to be generally satisfactory.

The quality of the colour transmissions, when received in black and white on ordinary television sets, was also carefully examined; 94% of technical observers found the pictures completely acceptable.

The results of the tests have been communicated to the Television Advisory Committee.

Voltage Standing-Wave Ratio Measurement

THE ATTENUATOR-SUBSTITUTION METHOD

By **E. W. Collings***

SUMMARY. *In certain types of standing-wave measurement, such as the measurement of the response of a cavity resonator in which the nodal position is not required, accurate standing-wave measurements can be made by substituting a short-circuited attenuator for the cavity under test and reproducing the same maxima and minima on an uncalibrated standing-wave detector. Practically all errors of the usual standing-wave measurement can in this way be eliminated. The accuracy of the measurement relies entirely on the accuracy of calibration of the attenuator, and is equal to the attenuator error for settings of up to about 4½ dB; but becomes an increasingly smaller fraction of the calibration error as the attenuator setting increases beyond this value (i.e., as the calibration measurement itself becomes more difficult). The accuracy of the method is discussed, and an example of its application is given.*

The design of standing-wave indicators, and methods of evaluating errors associated with their use, have been considered by several authors[1,2,3]. In an impedance measurement, a standing-wave machine is required to measure both the voltage standing-wave ratio (v.s.w.r.) and the nodal position. Sorrows, Ryan and Ellenwood[3] have stated that errors in these two quantities can arise due to the following causes:

(1) Variation in amplitude and frequency, and the presence of harmonics, in the u.h.f. power source.

(2) Calibration error of the detector system. (The above-mentioned writers have recommended using a calibrated attenuator to determine the probe law.)

(3) Alteration of the standing-wave pattern by the probe itself.

(4) Attenuation in the slotted section through which the moving probe penetrates. This is usually negligible but allowance can be made for it.

(5) Radiation of power by the slot, and reflections from the ends of the slot. These effects are also usually negligible.

(6) Structural defects in the moving parts, which cause the carriage to wobble and the depth of penetration of the probe to vary as the slotted section is traversed. The paper by Sorrows et al showed how the resulting errors in an impedance measurement could be estimated with the aid of an experimentally determined constant.

The chief sources of error in the standing-wave detector itself are therefore:

(1) *Calibration error of the detector system*

This includes errors in experimentally relating detector output to electric field strength in the waveguide or coaxial line (using perhaps a matched termination, a calibrated attenuator, and a u.h.f. power source), together with the loss in accuracy in transferring this information to a calibration chart, and lastly, in an actual experiment, the use of the calibration chart to give the final v.s.w.r.

* *Victoria University of Wellington, New Zealand.*

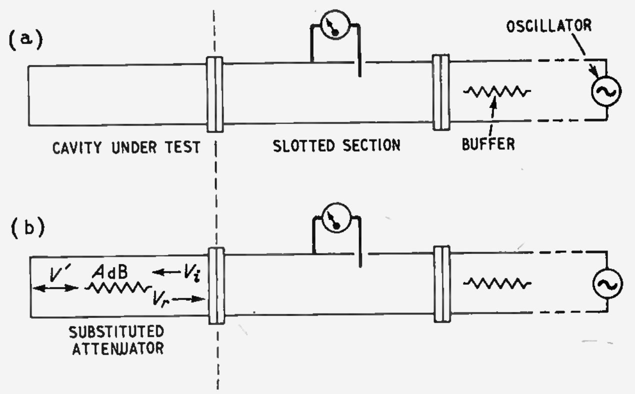

Fig. 1. *Illustrating the attenuator substitution method for measuring v.s.w.r.*

(2) *Change in depth of penetration of the probe as the carriage moves along the slot*

(3) *Reflection of power by the probe*

The last two considerations are inter-related, for error due to change in depth of penetration of the probe, in a carefully-made standing-wave machine, will be negligible at large probe penetrations where probe reflections may have to be considered. According to Altar[2], probe reflections become measurable for (waveguide) penetration greater than 25% ; so that for sufficiently large waveguides it should be possible to choose some intermediate probe penetration that will eliminate both error (2) and error (3). If this is not possible it is better to use a relatively large probe penetration, since error due to probe reflection can be measured and eliminated, whilst that due to change in depth, although the maximum error can be estimated, cannot be allowed for.

TABLE 1
Evaluation of the approximation $A = 8 \cdot 7/S$

S	$8 \cdot 7/S$	$10 \log_{10} (S+1/S-1)$	Error (%)
1·5	5·8	6·99	17
2	4·35	4·77	8·8
3	2·9	3·01	3·7
4	2·18	2·23	2·2
5	1·74	1·76	1·2
6	1·45	1·46	0·7

The Attenuator Substitution Method

In certain kinds of u.h.f. measurements, for example measuring the response of a resonator over a range of frequencies, only the v.s.w.r. needs to be known accurately and the nodal position is not important. In such measurements both calibration error and probe reflection correction may be eliminated by replacing the cavity, after a test, by a short-circuited precision attenuator, and adjusting the attenuator settings so that the same set of maxima and minima are reproduced on the standing-wave machine (Fig. 1). The method is of the *substitution* kind, in which a device setting up a calculable standing-wave replaces the object under test and reproduces the same external conditions.

Calibration error, in the standing-wave detector, is eliminated, since it is used merely to indicate a pair of reference readings for each measurement. Error due to variation of the probe penetration can be eliminated by making the probe of sufficient length ; then even if probe reflections occur, they need not be considered, for they will produce the same effect on the standing-wave pattern, with both the cavity under test and its substitute (the short-circuited attenuator). The effect of probe wobble will be magnified with a deep probe penetration; but it should always be possible to ensure rigidity in a probe carriage assembly, even if high machining accuracy of the slotted top plate cannot be guaranteed.

The error in the v.s.w.r. measurement relies entirely on the accuracy of calibration of the attenuator and is readily assessed. In what follows the accuracy of the method will be discussed, and an example of its application to a particular waveguide standing-wave problem will be given.

TABLE 2
Relationship between other errors in S and A

S	$\Delta S(1\%)$	$\Delta A = \dfrac{8 \cdot 7}{S^2 - 1} \Delta S$	A (dB)	$\Delta A(\%)\left(\dfrac{\text{or} \Delta A/A}{\Delta S/S}\right)$
1·01	0·0101	4·4	23	19
1·03	0·0103	1·5	18·3	8·3
1·05	0·0105	0·92	16·1	5·7
1·08	0·0108	0·55	14·2	3·9
1·1	0·011	0·45	13·2	3·4
1·2	0·012	0·24	10·4	2·3
1·5	0·015	0·105	7·0	1·5
1·9	0·019	0·061	5·1	1·2
2	0·20	0·058	4·8	1·2
3	0·03	0·033	3	1·1
4	0·04	0·023	2·2	1·04

(1) *Standing-wave produced by a short-circuited attenuator*

If V_i is the amplitude of a voltage wave incident on a length of attenuative transmission line short-circuited at the far end (Fig. 1), with V_r the amplitude of the reflected wave, the reflection coefficient is $\Gamma = V_r/V_i$. The incident and reflected waves then produce a standing-wave of ratio S, where

$$\Gamma = \frac{S-1}{S+1} \text{ (e.g., reference 4).}$$

For A dB of attenuation,

$$A = 20 \log_{10} \frac{V_i}{V'} \text{ (Fig. 1)}$$

$$\text{and } A = 20 \log_{10} \frac{V'}{V_r}$$

Therefore

$$2A = 20 \log_{10} \frac{V_i}{V_r}$$

$$\text{and } A = 10 \log_{10} \frac{S+1}{S^2-1} \cdot \qquad \qquad (1)$$

Also by differentiation, the change in v.s.w.r. produced by a small change in attenuation is given by

$$\Delta A = \frac{8 \cdot 7}{S^2 - 1} \Delta S . \qquad \qquad (2)$$

For the purpose of assessing the error in S resulting from a given error in A, equation (2) is too elaborate, and a simple approximate expression for S in terms of A is desirable :

$$A = 10 \log_{10} \frac{S+1}{S-1}$$

$$= 4 \cdot 34 \log_e \left[\frac{1 + (1/S)}{1 - (1/S)}\right]$$

$$= 4 \cdot 34 \times 2 \left[\left(\frac{1}{S}\right) + \frac{1}{3}\left(\frac{1}{S}\right)^3 + \ldots\right], \left|\left(\frac{1}{S}\right)\right| < 1 ,$$

$$= \frac{8 \cdot 7}{S} + \frac{2 \cdot 9}{S^3} + \ldots \quad .$$

The first approximation, $A = 8 \cdot 7/S$, is useful for quickly determining the attenuation required to produce a given standing-wave, especially with large values of S. Table 1 shows the error incurred by the use of this formula to be less than 2% for standing-wave ratios greater than 4.

(2) *Error in v.s.w.r. for a given error in the attenuation value*

(a) The approximation, $A = 8 \cdot 7/S$, which is a convenient form for error calculation, gives: $dA/A = -dS/S$.

The percentage error in v.s.w.r. is thus seen to be equal to that in attenuation, for the range of values of A over which the approximation is reasonably good; i.e., for v.s.w.r. greater than 2 or 3. (See Tables 1 and 2.)

(b) For S less than 2, more and more terms in the series for A must be introduced. The second approximation,

$$A = \frac{8 \cdot 7}{S} + \frac{2 \cdot 9}{S^3}$$

$$= (8 \cdot 7S^2 + 2 \cdot 9)/S^3,$$

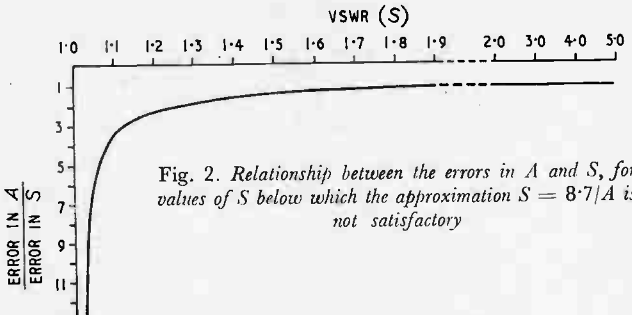

Fig. 2. *Relationship between the errors in A and S, for values of S below which the approximation S = 8·7/A is not satisfactory*

indicates that the error in S becomes some fraction of the error in A. Fig. 2 (from Table 2) substantiates this, and shows that the uncertainty in S becomes an increasingly smaller fraction of the error in A, as S decreases beyond the value 2, and particularly so when S is in the range $1 \cdot 01$ to $1 \cdot 1$.

(a) and (b) reveal two facts about the relationship between the errors:

(1) The error in v.s.w.r. measurement is, to a close approximation, equal to that of the attenuator setting when S is greater than about 2; i.e., when A is less than about $4\frac{1}{2}$dB (Table 2).

(2) As S diminishes below the value 2, the error in v.s.w.r. becomes an increasingly smaller fraction of that in A.

There is therefore no conflict between the accuracy requirements in (1) and (2). The small attenuation needed to produce a large v.s.w.r. can be measured accurately (by a power-ratio method, using a thermistor bridge and the d.c. substitution method); whereas the inaccuracy involved in measuring large power ratios, corresponding to the large values of A needed to produce small standing waves, has a very much diminished influence on the accuracy of the final v.s.w.r.

Thus the form of the relationship between A and S has a compensatory influence on the reduced accuracy with which large attenuations can be measured; the tendency is therefore to keep the percentage error in S reasonably constant over a wide range of values of S.

TABLE 3
Response of lossy resonator

1	2	3	4	5
0	$\frac{45}{1}$			
1	$\frac{41}{2 \cdot 6}$	3·2	2·45	
2	$\frac{36 \cdot 5}{5}$	4·53	3·45	2·35
$2\frac{1}{2}$	$\frac{34 \cdot 5}{7}$	4·83	4·3	2·18
3	$\frac{34}{7 \cdot 5}$	5·01	4·48	2·11
$3\frac{1}{2}$	$\frac{34 \cdot 5}{7 \cdot 5}$	4·95	4·42	2·13
4	$\frac{37}{9}$	4·65	4·05	2·3
5	$\frac{41 \cdot 5}{4 \cdot 5}$	3·77	3·05	

Column 1 sets out the numbers of $\frac{1}{2}$-turns of the test oscillator tuning screw, by means of which the frequency was changed in small increments of approximately 2 Mc/s. Column 2 contains the ratio of maximum to minimum readings of the standing-wave detector; and Column 3 the drum reading of the short-circuited attenuator needed to reproduce these same readings. In column 4 these readings are converted into decibels, and Column 5 sets out the corresponding values of S from a plot of the relationship $A = 10 \log_{10} (S+1)/(S-1)$.

(3) *General applications of the method*

(a) For the setting up of a known standing-wave in order to test the performance of a slotted line standing-wave detector.

(b) With the aid of an uncalibrated slotted line, the short-circuited attenuator can be substituted for the load under test when making an actual standing-wave measurement.

(4) *Limitations of the method*

Before calibrating the attenuator, it must be ascertained that the card itself produces relatively negligible reflections. This is particularly important when very small standing waves are to be produced. Reflection from the card can be reduced by increasing the amount of taper on the leading edge.

Application of the method is restricted to a determination of the magnitude of a standing wave ratio; i.e., a standing-wave machine itself must be used if the nodal position is required, as it is in an impedance measurement.

Application of the Method

In assessing the output noise temperature of a tuned waveguide cavity containing a noise source it was necessary to measure the voltage standing-wave ratio at resonance as accurately as possible. The losses in the particular cavity under investigation by the present author were quite large, owing to the presence of a soft glass bulb (the envelope of an incandescent filament noise lamp) with the result that the resonant v.s.w.r., σ_r, was approximately 2. Because of the form of the relationship between the excess output noise temperature of the cavity $(T_n - T_r)$, σ_r, and the excess noise temperature of the filament [viz., $T_n - T_r = \sigma_r - 1/\sigma_r (T_l - T_r)$*; where T_n, T_l are the noise temperature of the cavity output terminals and the filament, respectively; and where T_r is room temperature] the error in excess output noise temperature was $1\frac{1}{2}$ times that of the standing-wave measurement. It was therefore desired to measure σ_r to an accuracy better than 1%. By using an attenuator ofac curacy $\pm 0 \cdot 06$ dB, σ_r, was measured to an accuracy of $\pm 0 \cdot 8\%$ in the following way:

As the resonant frequency was traversed in small increments of frequency, the maximum and minimum of the standing-wave pattern was recorded at each step. The mounting under test was. then replaced by the short-circuited attenuator. By alternatively adjusting the test oscillator output and the waveguide attenuation, a given pair of maximum and minimum

* c.f. Reference 5.

readings could be reproduced. The attenuator setting was then noted. This was repeated for other standing-wave readings, working outwards both ways from the minimum. The attenuator settings were then converted into decibels from the calibration chart, and the corresponding v.s.w.rs were read from a graph of the relationship $A = 10\log_{10}\ (S+1)/S-1)$. This is illustrated by Table 3, from which a graph was drawn and the minimum v.s.w.r. found[5] to be $2 \cdot 11 + 0 \cdot 02$.

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The Fringe of the Field — *By Quantum*

A TOUCH OF THE SUN

Recently I went off in search of the sun, and had what I should like to think you would agree was a well-earned holiday rather earlier than usual. From the irradiation aspect, it was rather more successful too, and I was very conscious of the unwearied sun from day to day. You will find in Hanbury-Brown and Lovell an excellent account of the sun as a radio source, and details of the main features I shall be mentioning in Special Report No. 4, on "The Distribution of Radio Intensity over the Solar Disc", published in 1954 by the International Scientific Radio Union (in French). A handsome volume ("Selected Lectures in Modern Physics") published by the Nuclear Research Foundation of the University of Sydney in 1958, has a lecture by J. P. Wild which I have used also. But I suddenly realized that the really magnificent thing about all this radio work on the sun is the way in which it fills in the pattern that was already developing about twenty years ago as the culmination of a century of work on common or garden sunshine.

The Pre-Radio Picture of the Sun

What we see as the sun is a luminous ball about $1 \cdot 3 \times 10^6$ km in diameter, of angular diameter a little over 30′, the general optical emission from which is roughly that of a black body with surface temperature 5,700°K, as far as the continuous spectrum goes. The ultra-violet emission suggests a higher temperature than this at or near the surface; and it is known that the internal temperature must be of the order of several millions of degrees at least, perhaps as high as $1 \cdot 4 \times 10^7$ °K. It is tempting to use an earth-analogy, for all its obvious limitations, and to regard what surrounds the visible ball as the sun's 'atmosphere'. If we do this, counterparts to the various atmospheric layers can be named. The 5,700° 'surface' is called the photosphere; above this is the first atmospheric layer, the slightly cooler reversing layer which we can liken to the troposphere; outside this, the chromosphere, where the temperature has risen again to something of the order of 10^4°K, rather corresponding to the top of the stratosphere; and outside it all, extending for several million kilometres at least, the corona which corresponds to the ionosphere (Fig. 1).

You were probably taught, as I was, that the reversing layer in which the Fraunhofer absorption lines characteristic of terrestrial elements appear superposed on the solar spectrum, and the chromosphere which exhibits these lines as bright emission lines when it is seen on its own during an eclipse, are one and the same. The familiar Kirchhoff-Bunsen experiment suggests that they could be; but it is quite certain that they are really distinct. The line spectra show that elements as we know them are there all right, conspicuously hydrogen, calcium, and iron; but their relative abundance in these regions differs from that on the earth. Hydrogen is by far the most abundant.

The light from the corona is so weak that it can only be observed (except at high-altitude stations where the sky-light background is less than at sea level) during eclipses. It shows a line spectrum, and a feeble continuous spectrum. Some of the coronal lines (at one time thought to be due to undiscovered elements!) which could not be attributed to any known or unforbidden emission under conditions then believed likely first led people to think of temperatures about 18,000°K, and ions in a highly ionized state. It is now believed that the temperature of the corona close to the sun is of the order 10^6°K, and that the elements there are pretty well completely ionized; considering the predominance of hydrogen, it is chiefly a plasma of electrons and protons. It seems that the radio emission supports this view. But there are other problems about the corona. The light is plane polarized, suggesting that it is being scattered by small particles, possibly as large as 3μ in diameter; but how can they survive at such temperatures? Scattering by free electrons is another process that has been invoked (not in connection with the polarization, though). That the corona may indeed extend as far as the earth and beyond it is suggested by the zodiacal light, a faint glow observed in the western sky after evening twilight, and in the eastern sky before dawn; and by the *gegenschein*, a faint reflection of the sun in the midnight sky. What are these reflecting particles doing? It would fit in with matters nicely if they were cosmic dust being swept in towards the sun by its gravitational field; but the Doppler effect shows that they are travelling outwards from it. But I think we might leave till the end this vast outer corona-corona, about which there has been some fascinating work done recently as, in any case, all that we shall be thinking about for the moment is the relatively whiff-sized inner part of the corona that is nearest to the sun.

The chromosphere is probably the most interesting region. This erupts. Eruptions, solar flares, or chromospheric prominences, are observed in the neighbourhood of sunspots. The general features of sunspots themselves are well established. These little dark markings, which move across the face of the sun, with eleven (or twenty-two) year cycles of intensity, are huge vortices within which there are strong magnetic fields. Their appearance is associated with radio fade-outs, bursts of solar radio noise, magnetic storms and, less regularly, with enhanced cosmic-ray intensity. But the really direct link between the sunspots and these terrestrial observations is the chromospheric eruption, or flare. The flare itself is a very strong source of hydrogen H_α radiation, and of ultra-violet. Simultaneously with its appearance, radio fade-outs and anomalous phase-changes in radio waves reflected from the ionosphere are reported; these are attributable to the ionizing effect of the ultra-violet emission. Magnetic storms follow a day or two later, the average delay being about twenty-six hours or so; these are attributed to particles which take this time to travel the 93,000,000 miles to us, at about 4,000,000 miles an hour or less, with speeds from about 1,800 to 500 km/sec in metric units. Cosmic-ray activity is observed more rarely, within about fifteen minutes of the eruption. There was a notable outburst in 1956 which is still fresh in the memory; the particles responsible for this, originating at the eruption, must have travelled with speeds between a tenth and a hundredth of that of light.

Taking the earth-analogy literally, it would be hard to see how a body like the sun, with a gravitational pull twenty-eight times as great as the earth's, could have an extensive 'atmosphere'. A merely gravitating gas-kinetic atmosphere in equilibrium would be about 10 km thick. The important additional factor competing with gravitation is radiation pressure, and this is what "keeps it up". But the equilibrium is not entirely stable. E. A. Milne worked out that, for an ion with a vertical velocity greater than 1,600 km/sec, the Doppler effect shifts its absorption line so far to the violet that it absorbs quanta of more than the equilibrium energy, and accelerates outwards clear of the gravitational field.

It would seem, then, that the chromospheric eruptions shoot out a great upsurge of vertically-absconding ions which start off with a velocity great enough to carry them clear of the sun, and continue with this velocity to reach the earth and give the observed magnetic storm a day or two later. If you want a short-term earthy analogy to this, there is the vertical instability of water-saturated air, which draws the energy for its ascent from the latent heat given out as water-vapour condenses—the first stage of the convective-rain sequence; but this doesn't really hold beyond the mere occurrence of instability, of course.

The optical-observation picture of the sun, in particular the expectation of high coronal temperature and extensive ionization there, was brought to this stage during the decade 1930–40. All was ripe for radio observations to continue it when they began. The interesting thing is that on the whole they confirmed these expectations, and also filled in many of the details of the picture.

Emission Processes

Excluding very high-energy electromagnetic radiations of nuclear-change origin, and thinking only of the optical and radio-frequency range, it can be said that all electromagnetic waves are generated by an electron-acceleration process of some kind. Four such processes are known.

(*i*) *Thermal Radiation.* Within a hot gas, collisions between electrons and nuclei provide the sudden acceleration. This is the 'black-body' radiation, for which Planck's distribution law applies. The radiation from the undisturbed sun—that is, over the whole of it when sunspot activity is negligible, and well away from the spots if they are there—is of this kind. Since the distribution of the energy between the different wavelengths of the black-body continuum depends on the temperature, the absolute temperature T can be found from this distribution. This is one way, for example, in which the figure 5,700°K is obtained for the photosphere.

Much higher values of T are found, however, for

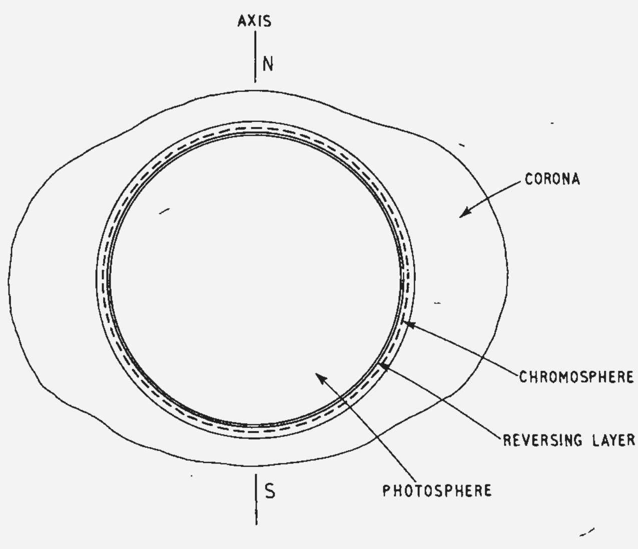


Fig. 1. *The sun as it is observed optically. Reversing layer and chromosphere are not to scale, the depth of the two together probably being of the order 20,000 km. The relative extent of the corona as photographed at eclipses is roughly to scale*

radio wavelengths; the effective temperature computed for wavelength λ increases as λ increases, from about 10^4°K at $\lambda = 1$ cm to 10^6°K at $\lambda = 1$ m. These are not irreconcilable with the photosphere temperature. In the first place, as already mentioned, there are other reasons for believing that the atmosphere round the sun is a good deal hotter than the photosphere. Secondly, they may simply mean, according to Martyn and Ginsburg, that the various radiations originate at different levels in the chromosphere or corona at which such temperatures actually obtain. But it must be remembered that the analogy between the corona and the ionosphere probably holds at least to the extent that there is a good deal of absorption, and also of reflection back to lower levels, of the radiation that reaches it. To put it in another way, just as our own atmosphere cuts down any outgoing radiation, so the sun's atmosphere may cut down the radio emission from the layers below. If this is so, then we would expect that the effective temperatures computed from the black-body laws are not absurdly high, but really more likely to be too low! This point is taken up later.

(*ii*) *Line Radiation.* These are due to quantum transitions between atomic energy levels. Examples which have been discussed in detail in recent "Fringe" articles are the H_α line, and the 21-cm line from atomic hydrogen.

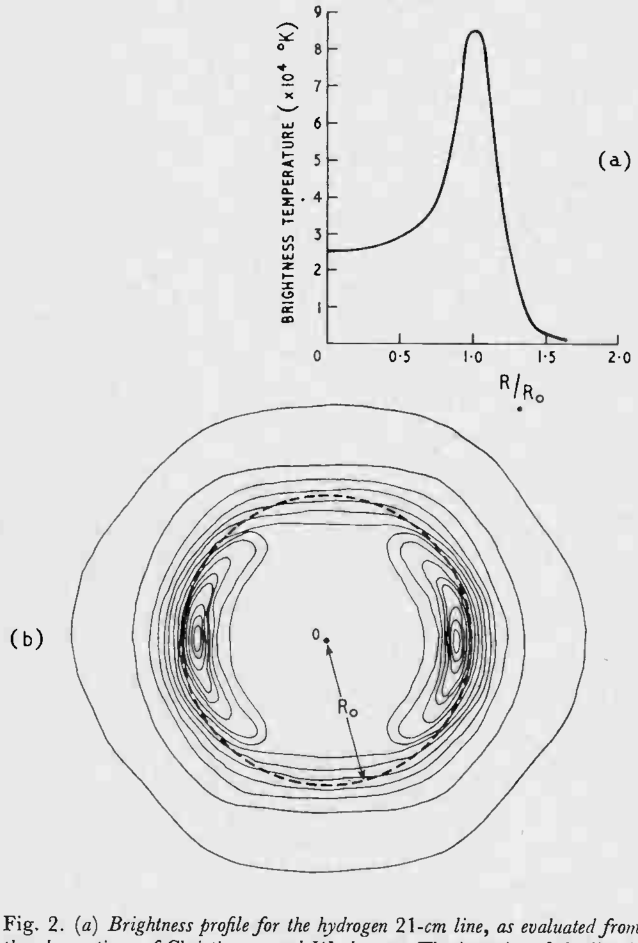


Fig. 2. (*a*) *Brightness profile for the hydrogen* 21-*cm line, as evaluated from the observations of Christiansen and Warburton. The intensity of the line is expressed in terms of 'brightness temperature'*; (*b*) *contours of equal brightness, corresponding to the distribution of* (*a*) *for the* 21-*cm line. Both diagrams refer to the undisturbed sun*

Both these lines are prominent in the sun's emission, particularly in chromospheric eruptions.

(*iii*) *Plasma Oscillations.* These are the natural electron oscillations which occur in a plasma of free electrons and positive ions, as the aftermath of some disturbance of the system which has supplied it with energy. The theory of the plasma resonance and dispersion of an electron gas is given in von Hippel's "Dielectrics and Waves", p. 262. The problem is simplified because the Coulomb forces between the individual *like* charges cancel out symmetrically, and all that has to be considered is the dipole moment per unit volume of the plasma and its simple-harmonic oscillation on being disturbed. You like your analogies, I know; and this is not unlike the acoustic resonator, for which you do not have to bother about the collisions of individual molecules but only consider the pressure fluctuations of the whole. The fundamental frequency is given by

$$f = \sqrt{\frac{e^2 N}{\pi m}},$$

where e is the charge on the electron in absolute e.s.u., m the mass of the electron, and N the electron density. Harmonics of frequency $2f$, $3f$ . . . can also be emitted. The dielectric constant of the plasma for waves of frequency f' is $(1 - f^2/f'^2)$.

The process is thought to be that responsible for the coronal radiation from the disturbed sun, though Hanbury-Brown and Lovell regard this as very speculative.

(*iv*) *Gyro-Radiation.* How the same characters turn up month by month! Yes, radiation of frequency

$$f_g = \frac{He}{2\pi mc}.$$

This does not appear so far to have any direct solar application, but is believed to be the means by which radio sources in the galaxy (where there must be magnetic fields) may be radiating.

A Note on Temperature

So far, temperatures ranging from 5,700°K to 10^6°K have been bandied about, without any too precise statement of their meaning. There are two possible meanings in this context. The first is the 'black-body' or radiation temperature, deduced from the intensity of the radiation we receive; this, when deduced for a particular wavelength, is called the 'brightness temperature' for that wavelength. It can be regarded really as another way of recording the intensity of the radiation; for clearly anything that reduces the radiation—say an absorbing layer between the observer and the source—reduces the 'temperature'. It is in this sense that the name brightness temperature is used in the graphs of Figs. 2 and 3.

But, when speaking of the temperature of the chromosphere at which a certain ultra-violet line can be emitted, or the temperature of the corona at a certain height, we are really thinking of gas-kinetic temperature of the more ordinary kind; the average kinetic energy of translation of an ion or an electron is $3/2\ kT$, where k is Boltzmann's constant. If this is just as irreconcilable as the other meaning was with the kind of temperature that you can feel and reduce with aspirins, the $3/2\ kT$

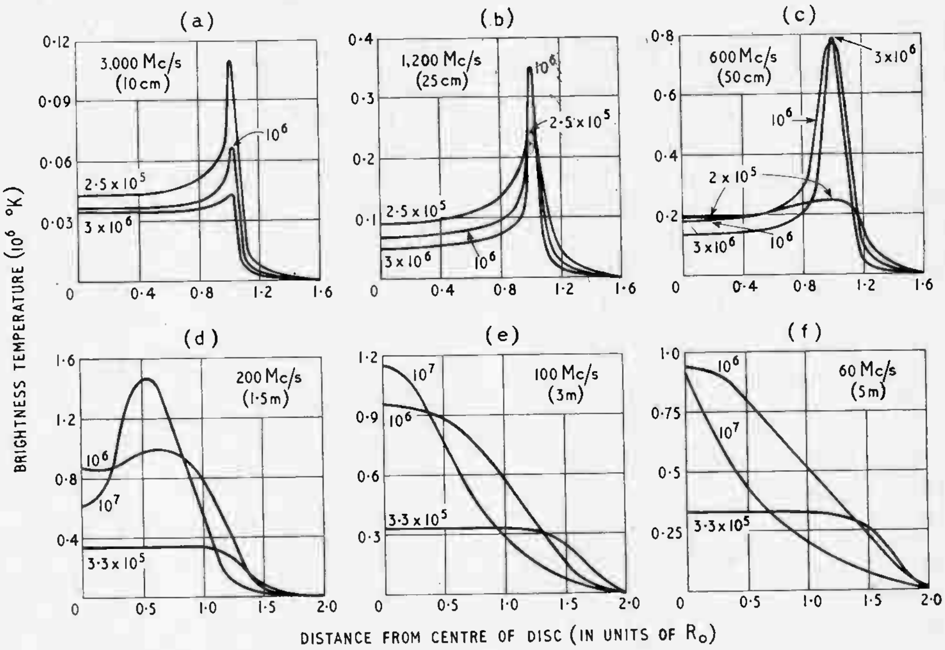


Fig. 3. *Brightness profiles calculated theoretically by Smerd for the undisturbed sun. The brightness is given in terms of 'brightness temperature'. In each figure, three different guesses have been made at the temperature of the corona (that is, the gas-kinetic or electron temperature) as a basis for calculating the three curves. It will be seen that the theory predicts the limb-brightening effect of Fig.* 2 *for the shorter wavelengths.* (*From U.R.S.I. Special Report No.* 4.)

can equally well be converted to ergs or to electron-volts for you. The point is, that it is this figure for T which, together with the electron density N, is used in the calculations of the way in which the various parts of the solar atmosphere behave, and it is this temperature (the 'collision-temperature', or the 'electron-temperature') that is usually understood to be the T referred to. There appear to be two ways of finding T. First, the product $N^2T^{-3/2}$ is proportional to the absorption coefficient for radio waves of any given frequency; when this is found, the next problem is to estimate N from the intensity of the corona's continuous spectrum, assuming it to be electron-scattered light and using Thomson's formula for this process; substituting back gives T. Secondly, the breadths of the coronal emission lines give a value which would be this T if they were solely excited by collisions; as radiation plays some part in the excitation, the temperature actually calculated in this way lies between the radiation temperature and the collision temperature. These matters are discussed fully in an article on "The Physical Condition of the Solar Corona" by C. W. Allen (Reports on Progress in Physics, 1954) which gives tables of estimates of the values of N and T for various coronal altitudes. Hanbury-Brown and Lovell seem to imply that there is uncertainty about the *interpretation* of these coronal temperatures; I think they must simply mean that the figures given by different workers do not concur exactly.

Radio Profiles of the Undisturbed Sun

A single telescope to give the necessary resolving power of 3′ of arc—that is, less than one-tenth of the sun's angular diameter—would, as elementary optical theory shows, require an aperture of 1,000 λ for radiation of wavelength λ. In practice, arrangements which are really descendants of the optical Michelson stellar interferometer are used. For example, W. N. Christiansen at Sydney made an array of sixty-four steerable parabolic reflectors arranged along the arms of a 1,200-foot cross (a 'Mills Cross') for solar observations on the 21-cm line. The earliest attempts at scanning the sun's disc were, in fact, made during eclipse—using the moon to do the scanning; these infrequent observations showed at least that there was something worth investigating!

Fig. 2 gives the results obtained by Christiansen and Warburton for the 21-cm line in the undisturbed sun (a) as a brightness profile, and (b) with contours of equal brightness. The explanation of the enhanced brightness at the two limbs is a rather complicated piece of analysis, worked out by S. F. Smerd and K. C. Westfold in 1949; taking optically reasonable estimates

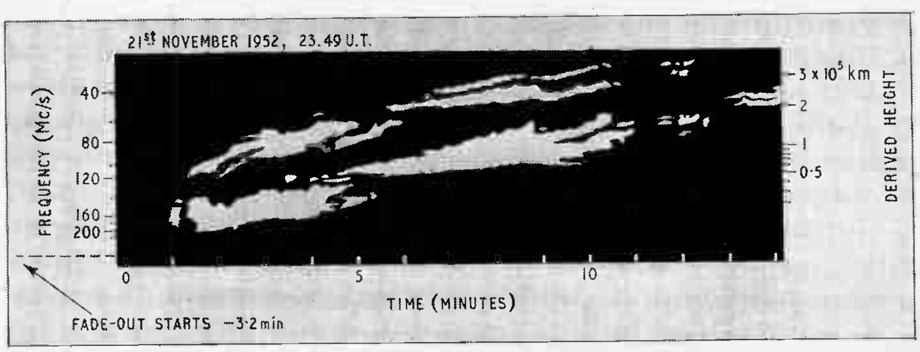


Fig. 4. *Spectrum of a large radio-emission burst from an eruption, and its variation with time. The ordinates on the left (reading downwards) give the frequency, and show fundamental and second harmonic; those on the right, the calculated height. The abscissa is the time-axis.* (*J. P. Wild, J. D. Murray and W. C. Rowe*)

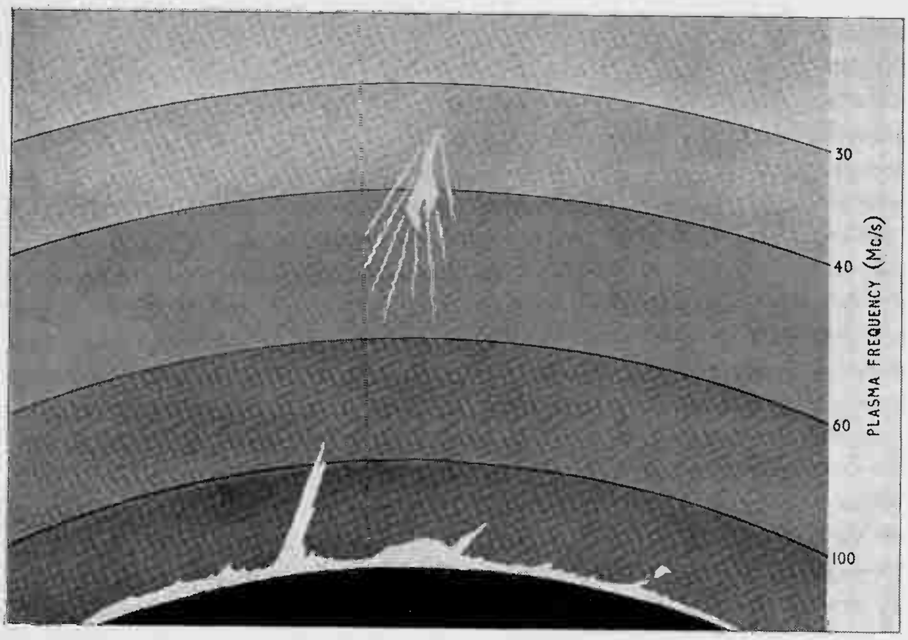


Fig. 5. *Picture of the ejection of a radio-emitting source in the solar atmosphere. The numbers on the right give the plasma frequency at different levels. (J. P. Wild)*

of the electron density N and the temperature T at different levels, they were able to find the refractive index and absorption coefficient at each level for various radio wavelengths, and predicted that a profile of the type 2 (a) should be obtained for λ between 10 cm and 100 cm. For longer wavelengths theoretical profiles shown in Fig. 3 were obtained. It will be noted that the radio sun is a very much more extensive object than the photosphere sun and, in every case, even for the shorter wavelengths, exceeds the bounds of the photographed light-emitting corona.

The Disturbed Sun

While sunspots appear dark (and hence one would suppose cooler) against the photosphere background, the radio emission round them is tremendously enhanced. For the 21-cm line, T at the level of its origin is estimated as about 10^7°K! Bright radio patches, much larger in extent than the spots themselves, accompany them throughout their lifetime. The chromospheric eruptions, or solar flares, are short-lived events which may last a few minutes, or perhaps half an hour; they are accompanied by bursts of radio emission of about 3-m wavelength. Fig. 4 shows the variation of frequency with time for such a burst. It will be seen that there are two main streaks, which can be identified as the fundamental

$$f = \sqrt{\frac{e^2 N}{\pi m}}$$

of a plasma oscillation, and its second harmonic $2f$. Also, the frequency of each decreases as time goes on, with decreasing electron density. Assuming the variation of N with altitude, the height of the travelling radio source is obtained—the 'derived height' is on the right-hand side of the figure. The rate of ascent, at its steepest, appears to be rather more than 1,000 km/sec.

Fig. 5, from an article by J. P. Wild in the University of Sydney publication "Selected Lectures in Modern Physics", gives a picture of the radio-emitting source hurtling out through the corona, as interpreted from these observations. How the whole process originates is still a mystery, but it is very interesting to see how the radio observations link up with the earlier optical speculations on what might be happening during such an eruption. For this seems exactly the kind of thing that Milne predicted from his radiation-pressure instability calculation.

A Little above our Heads

Well, just above the F_2 region of the ionosphere really; and the question is, does a little of the corona really extend that far, and if so what is it like? Here there is some conflict between optics and radio. The zodiacal light suggests that something does, but the polarization cannot so far be reconciled satisfactorily with electron-scattering, and apparently gives a dusty answer. Perhaps conflict is the wrong word. I mean that the spectrum is, so far, associated with larger particles than the electrons which are the chief object of pursuit. For radio echoes from the moon, and observations on audio-frequency ('whistling') atmospherics, give an electron density N of the order of 300 to 1,000 per c.c., and further information is awaited from the work on artificial satellites. Is the outer ionosphere also the outer corona? The magnetohydrodynamics people talk of shock waves propagated from the sun as accompanying magnetic storms, and apparently assume that it is. But this new field of physics, the territory both of Hoyle and of Harwell, is too vast to comment on at short notice.

One suggestion from this new approach must however be mentioned. Prof. Sydney Chapman's account of a preliminary idealized model of the corona is reported in *Nature*, 3rd May 1958. Imagining the complete solar corona to extend to the boundaries of the known solar system, and leaving out of account possible magnetic fields and other complications, he proceeds, taking values of N and T in the corona at $1{\cdot}6\ R_0$, to find what they should be at the earth's distance. He finds N to be between 300 and 4,000 per c.c. (depending on the initially assumed values), and the corresponding T some $2{\cdot}2 \times 10^5\,°K$. It does appear that the electron density fits the observations so far; but of course, this is only a beginning. The interesting feature about Chapman's model is the temperature, with its implications of heat-conduction down through the ionosphere and the revision it may call for in our ideas. It looks as if we may have to think of the earth as travelling through a space which is not just populated with hydrogen atoms coldly exploiting their hyperfine-structure levels, but which is really a rarefied gas at a temperature far above that of the photosphere!

L.F. Random-Signal Generator

By J. L. Douce, Ph.D.* and J. M. Shackleton, B.Sc.*

SUMMARY. *This article describes a simple low-frequency noise generator having a power spectrum which is uniform from zero frequency to about* 15 *c/s. A new technique is employed, utilizing a conventional noise generator followed by a non-linear element. No critical setting-up procedure is required, and the method gives inherent freedom from effects of variations in components and supply voltages.*

In the testing of servomechanisms and real-time-scale analogue computers, a randomly-varying test input signal is often required. The frequency range of interest is usually from zero to a few cycles per second.

The production of a random signal having a uniform power spectrum over this frequency range is difficult, employing conventional techniques. Thus thyratron noise generators and photo-multipliers are found unsatisfactory, since their signal power per unit bandwidth varies appreciably with frequency at low frequencies. Where the primary source of noise gives a small power output, amplifiers must be used and at low frequencies the flicker effect will distort the frequency spectrum of the noise.

Since photo-multipliers and thyratrons give a noise spectrum which is approximately flat over the audio spectrum, a local oscillator and modulator can be used to heterodyne the noise so that the beat-frequencies are centred about zero. This method requires careful adjustment of tuned filters and smoothing networks to give a satisfactory output spectrum[1].

Other methods rejected in view of the complexity involved include tape or film recorders, used for changing the time-scale of available noise, and feeding a random mixture of brass and steel ball-bearings on to magnetic selectors.

The method used in this paper was suggested as a result of the theoretical analysis of a non-linear feedback system subjected to a random disturbance.

Theoretical Basis

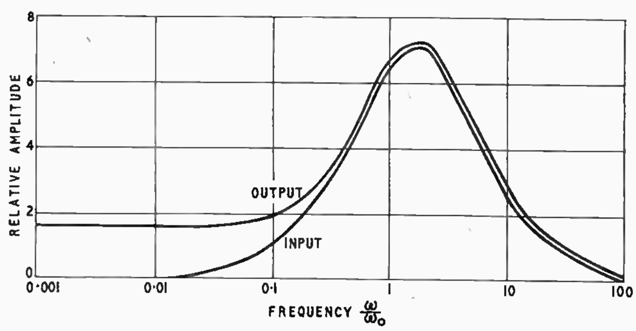
When any signal is applied to a non-linear circuit, the output spectrum will, in general, differ from that of the input. If the input signal consists of several sinusoidal signals, applied simultaneously, the output contains components of the input frequencies, harmonics of these frequencies, and intermodulation terms.

A random signal may be considered as the sum of a large number of small sinusoidal signals, each of differing frequency. Hence, when a random signal of given spectral density is applied to a non-linearity, the output spectrum contains a part having the same spectral density as the input, a part due to harmonics of the input frequencies and a portion due to intermodulation products. Since the input spectrum is continuous, the intermodulation terms contain components down to zero frequency.

For a particular input-power spectrum, it can be shown that the spectral distribution of the distortion terms is uniform over a wide frequency range, extending down to zero frequency[2].

If the signal applied to the non-linearity has negligible low-frequency components, the only low-frequency

Fig. 1. *Input and output spectra of non-linearity*



* *The authors were in the Department of Electrical Engineering at the University of Manchester. Dr. Douce is now at Queen's University, Belfast and Mr. Shackleton is with Avro Ltd.*

components at the output of the device will be due to intermodulation terms. Thus, by passing the output of the non-linearity through a low-pass linear filter, a random signal is obtained having the desired frequency spectrum.

To evaluate the frequency spectrum of a random signal passing through a non-linearity, it is necessary to consider the auto-correlation functions of the input and output signals. The auto-correlation function, $g(\tau)$ is related to the power spectrum by the transformation

$$g(\tau) = \frac{1}{2\pi}\int_{-\infty}^{+\infty} G(\omega)e^{j\omega\tau}d\omega$$

similarly $G(\omega) = \int_{-\infty}^{+\infty} g(\tau)\, e^{-j\omega\tau}d\tau$

For any non-linearity, the auto-correlation function of the output $g_o(\tau)$ signal is related to that of the input $g_i(\tau)$ by the expression

$$g_o(\tau) = a_1^2 g_i(\tau) + a_2^2 g_i^2(\tau) + a_3^2 g_i^3(\tau) + \ldots$$

where a_n depends only on the input power and the form of the non-linear function[3].

For saturation, where the output is limited to $\pm h$, when the r.m.s. noise amplitude is much larger than h, the above expression becomes

$$g_o(\tau) = \frac{2h^2}{\pi}\sin^{-1}\frac{g_i(\tau)}{g_i(0)}$$
$$= \frac{2h^2}{\pi}\frac{g_i(\tau)}{g_i(0)} + \frac{1}{6}\left(\frac{g_i(\tau)}{g_i(0)}\right)^3 + \ldots$$

The output signal is now independent of small variations in input power.

When the input signal consists of a noise extending over a certain band of frequencies; i.e.,

$$G_i(\omega) = \frac{K}{1+\left(\dfrac{\omega}{n\omega_0}\right)^2} - \frac{K}{1+\left(\dfrac{\omega}{\omega_0}\right)^2}$$

The output spectrum is as shown in Fig. 1.

This figure shows that the output spectrum is effectively flat below $\omega = 0 \cdot 1\ \omega_0$. Thus, if a large-amplitude random signal containing frequency components in the range 150–300 c/s is fed into a saturation-type non-linearity, the output spectrum will have constant power per unit bandwidth from 30 c/s down to zero frequency. Also the power in these low-frequency components will be independent of any variation in the input power of the signal applied to the non-linearity, provided that the r.m.s. amplitude of this signal is much larger than the clipping levels.

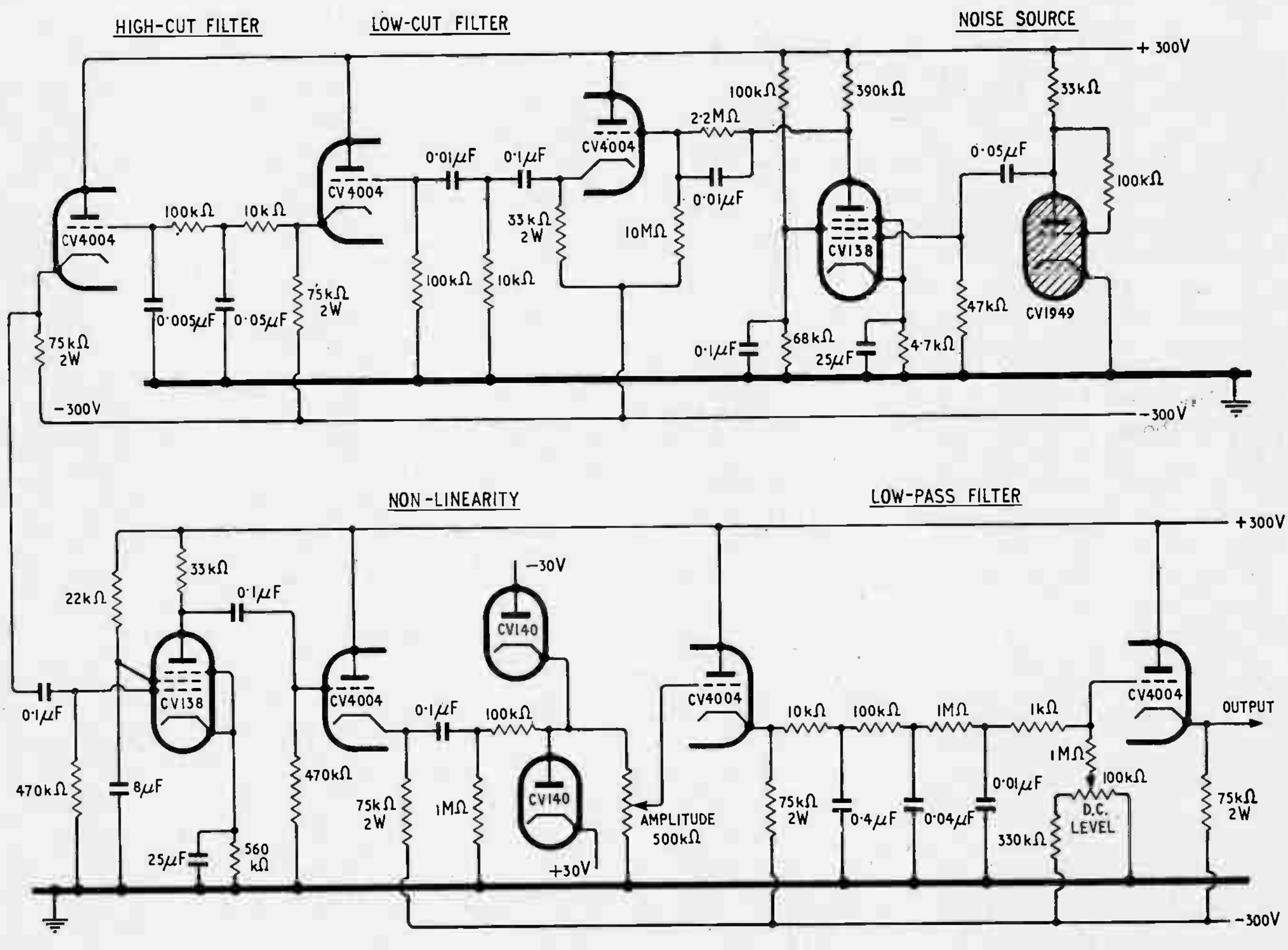
Practical Details

The complete system requires a source of audio-frequency noise, a band-pass filter, a clipping circuit to generate the low-frequency components, and a simple low-pass filter to isolate the desired frequency components.

Fig. 2 gives the practical circuit employed.

A thyratron provides a convenient source of random

Fig. 2. *Circuit diagram*



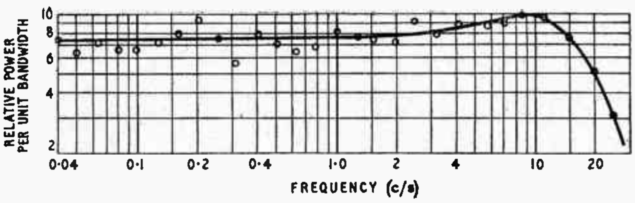


Fig. 3. *Output power spectrum*

noise[4], giving a uniform spectral density over the range 100–1,000 cycles per second. The thyratron is surrounded by a permanent magnet, supplied to increase the noise output. The output signal is of several volts r.m.s. amplitude.

After amplification, the noise is limited in bandwidth by high-pass and low-pass filters. The characteristics of these filters are not critical, and two, three or four stages of filtering may be employed with no effect on the final output spectrum. Wide tolerance components are used throughout.

The filters considerably reduce the r.m.s. amplitude of the noise, and so further amplification is required before the signal is passed into the clipping circuit. From this stage, direct coupling must be employed throughout, to preserve all the very-low-frequency components.

Following the non-linearity, three stages of low-pass filtering effectively remove all frequency components above 30 c/s.

Measurement of the Frequency Spectrum

Accurate determination of low-frequency power spectra presents considerable practical difficulties. A feedback network has been used to simulate a constant-Q circuit at various frequencies, followed by a r.m.s. meter of very long time constant. At the lowest frequencies of interest, $<0{\cdot}5$ c/s, the time required for each reading becomes extremely long, and the accuracy of this method is poor.

The most satisfactory method is to record a sample of noise on a long length (200 ft) of photographic film.

This film is then played back, at various speeds, using a cathode-ray tube curve-follower, and the signal fed simultaneously into four tuned circuits. The range of playback speeds is such that the differing gains of the various filters can be evaluated, and due allowance made.

Measurement shows that the power spectrum is effectively flat from 0·04 c/s to 15 c/s, being 3 dB down at 20 c/s (Fig. 3).

Conclusion

A simple method has been presented for producing a low-frequency random signal. The arrangement is particularly insensitive to factors normally affecting the quality of the output signal, and no critical components are involved. Measurement verifies the predicted spectrum of the noise source.

Acknowledgement

The authors gratefully acknowledge the assistance given by Mr. G. G. Sutton and his colleagues at Radar Research Establishment, Malvern, for their considerable help in the frequency-spectrum analysis.

The method of noise generation described in the paper forms the basis of a commercial instrument manufactured by Servomex Ltd. and is the subject of patent application 2131/58.

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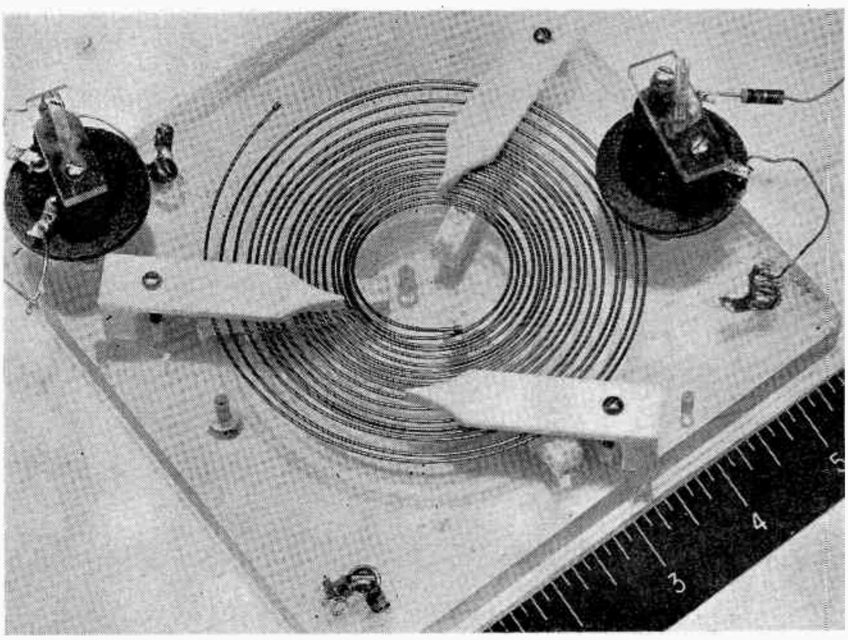
TORSIONAL-WAVE DELAY LINE

A DELAY line capable of providing a long delay in a small space has been developed at Bell Telephone Laboratories. Known as a spiral-coiled wire torsional-wave delay line, it permits the clear resolution of 10-microsecond pulses spaced 20 microseconds apart.

It consists essentially of a length of 0·038-inch diameter wire, made of a ferromagnetic alloy called 'vibralloy', coiled into a flat spiral. Vibralloy has a high mechanical Q and low-temperature coefficient of delay.

Torsional waves are transmitted through the wire at about $2{\cdot}68 \times 10^5$ cm/sec, making the theoretical delay about 3·72 μsec per cm. In one experimental line, a 15-foot length of wire exhibited a delay of 1·65 msec at 840 kc/s.

Ceramic torsional transducers were used to inject and detect the torsional waves. The total insertion loss was 9 dB, of which 6 dB was line loss and 3 dB represented conversion loss in the transducers. The bandwidth was about 100 kc/s.



Close-up view of experimental coiled-wire torsional-wave delay line

Coils for Magnetic Fields

PART 1: THERMAL ASPECTS

By G. M. Clarke, M.A., Ph.D.*

SUMMARY. *The introduction of the foil-wound solenoid for microwave valve applications makes desirable a comparison with the usual wire-wound solenoid. It is found that the internal temperature-rise limitation virtually disappears with foil winding until very high powers are applied. For even higher powers a solenoid wound with coaxial power cable can be used, but the poor filling factor makes this inefficient, and the external heat transfer problem is serious in the power range where it is necessary.*

Electrical and thermal design equations are developed for the three types and also for wire-wound coils cooled through radial or axial partitions. The latter can be made more efficient than aluminium-foil coils if fully impregnated.

The cylindrical wire-wound solenoid has long been a conventional method for producing magnetic fields, either within its bore, or in some external region by the use of pole pieces and yokes. Its two main advantages are:

(a) For a given coil, the resistance can be made to have any value within an enormous range, owing to the availability of insulated copper wire in many gauges.

(b) No special winding equipment is required unless the highest possible efficiency is required, or the coils are sufficiently large to exclude winding on an available lathe.

Against these must be set the disadvantage that round insulated wire, even if layer wound with special equipment, cannot fill the coil section with conductor and so leads to a coil of higher resistance, and thus greater power consumption for a given magnetic field, than is ideally necessary. Secondly, the heat generated within the coil can escape to the surroundings only by crossing the large number of thermal barriers introduced by the insulation of the wire.

The purpose of this paper is to evaluate these difficulties quantitatively and to see how far they may be overcome by winding with different forms of conductor such as metal foil or coaxial power cable.

It is obvious that changing the conductor from wire can only affect internal temperature differences, and so improvement can only be expected when the internal temperature-rise is the limiting factor. This is rarely the case with coils which are of sufficiently low power dissipation to be cooled by natural convection of air or oil or even with forced air, but it is nearly always the case with water-cooled coils such as are frequently necessary for high-power microwave valves.

For a uniform infinitely-long cylindrical solenoid the elementary formula

$$B \text{ (gauss)} = I/0{\cdot}796 \text{ ampere-turns/cm} \quad \ldots \quad (1)$$

gives the field B within the coil, when I is the current in amperes circulating the axis in every unit length. The radius does not enter this formula, which holds for an infinite-length coil of any radial thickness. For finite-length coils the field is reduced from that of Equ. (1) by a form factor which contains both the length and the radial dimensions. However, as we shall be concerned with a comparison between different constructions of coil of the same geometry it will not be necessary to consider this factor. For the same reasons it will not be necessary to take account of factors such as leakage due to any yokes and pole pieces, treatment of which can be found elsewhere[1].

If the section of a solenoid is uniformly packed with a conductor to a fractional filling s then, provided none of the conductors are branched into one another which is the usual case, the current density across the section will remain constant even is the resistivity varies. The total power dissipated in the coil W is then found by integration. Thus

$$W = \int_{r1}^{r2} \int_{-L/2}^{L/2} i^2\rho . 2\,\pi r . dr . dz \quad \ldots \quad (2)$$

where L is the coil length in the z direction, r_2 and r_1 are the outer and inner radii respectively and i is the average current density perpendicular to the axis. If the resistivity ρ is constant as in the case when there is negligible temperature rise in the coil then Equ. (2) can be integrated immediately, and using Equ. (1) gives

$$W = (0{\cdot}796B)^2 L \cdot \left(\frac{\rho\pi}{s}\right) \cdot \left(\frac{R+1}{R-1}\right) \quad \ldots \quad (3)$$

where $R = r_2/r_1$.

This is independent of absolute radial dimensions[2] and sets a lower limit to the power dissipated when the ambient temperature resistivity ρ_0 is used. To obtain

* *Ferranti Valve Laboratory, Edinburgh.*

the correct expression with resistivity temperature coefficient α, the heat flow problem within the coil must be studied. This involves the solution of a Poisson Equation for the temperature θ.

$$\Delta^2\theta + \frac{i^2\rho_0}{k}\cdot\left[1 + \alpha\,(\theta - \theta_0)\right] = 0 \qquad \qquad (4)$$

where k is the effective coil thermal conductivity.

The general solution of Equ. (4) in cylindrical co-ordinates can be obtained formally by separation of variables, but to satisfy the boundary conditions an elaborate computation is required for each shape of coil. Thus we will study the cases corresponding to cooling applied at the ends only, and then on the outer cylinder only. Equ. (4) then becomes one-dimensional with the temperature gradient either entirely axial or entirely radial. An approximation for the two-dimensional case can be made by a combination of these two cases.

Wire-Wound Solenoids

Effective Thermal Conductivity of a Wire-Wound Solenoid

Practical application of the treatment given in the later sections to wire-wound solenoids hinges upon an estimate of the effective average thermal conductivity of the coil.

It is clear that as electrical insulation must be provided between turns, this forms the main thermal barrier. Furthermore if insulated wire only is used then, theoretically, the thermal barrier will be entirely air as only line contacts can be achieved between the turns. Such a coil would be a poor proposition for a solenoid and will not be considered further. If the coil is impregnated with an insulating material then the situation is much improved and we have to consider the effective increase of the insulant thermal conductivity due to the embedded wires.

In a 'scramble-wound' coil the problem is statistical, but little error will result if we consider a regular array and evaluate the thermal conductivity for the same space-filling factor.

Consider two adjacent wires in a square array as shown in Fig. 1(a). As a first approximation, assume that the temperature gradient is zero in the z-direction when calculating the conductivity in the y-direction. As the conductivity of the copper is very large compared with that of the insulant k_i, the effective conductivity k is given by

$$k = k_i\int\frac{dz}{y} \qquad \qquad (5)$$

where $y = f\,(z)$ is the equation of the boundary emphasized in Fig. 1. The factor by which the insulant conductivity is increased by the presence of the wires is thus

$$K = \int_0^a \frac{dz}{d - (a^2 - z^2)^{\frac{1}{2}}} + \int_a^d \frac{dz}{d} \qquad \qquad (6)$$

The first integral can be evaluated numerically but this is hardly justified in view of the straight flow approximation. As we are interested in close spacings between the wires; i.e., $t = d - a$ is small, the denominator above may be approximated by expanding as a power series for small z, and neglecting higher terms.

$$K = \int_0^a \frac{dz}{t + \frac{1}{2}\,(z^2/a)} + t/d \qquad \qquad (7)$$

when it is integrable directly

$$K = \left(\frac{2a}{t}\right)^{\frac{1}{2}}\cdot\tan^{-1}\left(\frac{a}{2t}\right)^{\frac{1}{2}} + t/d \qquad \qquad (8)$$

It is consistent with the small t approximation to ignore the last term and replace the $\tan^{-1}$ by $\pi/2$ to give

$$K = 2\cdot 22(a/t)^{\frac{1}{2}} \qquad \qquad (9)$$

The approximations in the above are equivalent to evaluating the integral of Equ. (5) under a parabola

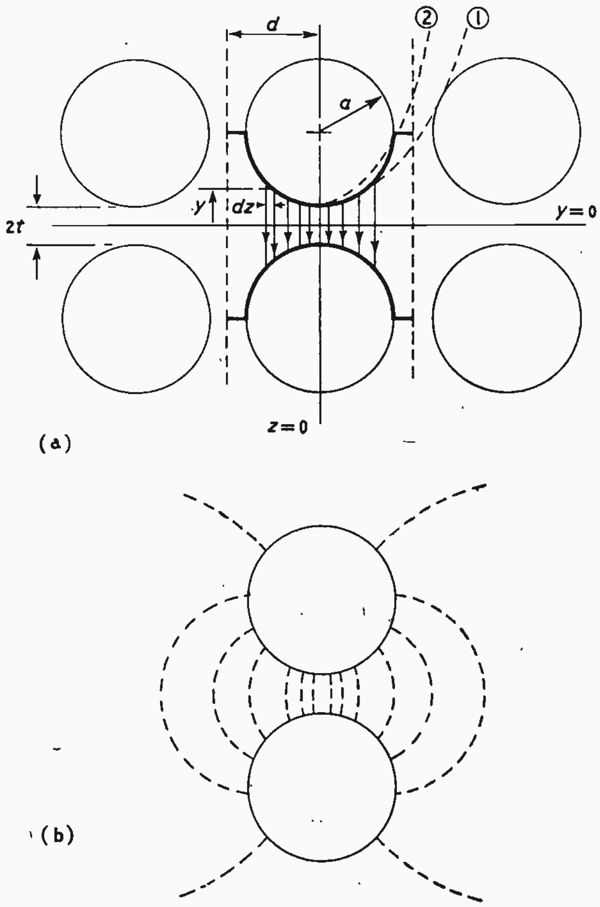


Fig. 1. *Models for heat conductivity of wire-wound solenoids*

[as shown in Fig. 1(a), curve 1] with the limits extending to infinity. As the parabola is everywhere closer to the axis than the circle, Equ. (9) gives a high answer. By replacing the half by unity in the denominator of Equ. (7) the parabola can be made to lie within the circle in the important region (curve 2) and this gives a different constant from that in Equ. (24).

$$K = 1\cdot 57(a/t)^{\frac{1}{2}} \qquad \qquad (10)$$

One cannot conclude that this is a lower limit as the influence of the assumption of straight flow paths is unknown. In fact it probably is so, as can be seen by studying the exact solution for two isolated wires, Fig.

1 (b), which can be obtained by conformal transformation as

$$K = \frac{\pi}{\log_e [1 + (t/a) + \sqrt{(1 + (t/a))^2 - 1}]} \quad .. \quad (11)$$

For small t this tends to the same answer as Equ. (9) and again would be expected to be high owing to the neglect of adjacent wires. Thus it would seem that the constant lies between 1·57 and 2·22, and 2·0 would seem a reasonable value.

Thus we will use

$$K = 2 \cdot 0\, (a/t)^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (12)$$

One could use for t the known insulation thickness of the wire, for the conductors can certainly not approach closer than twice this. Thus the thermal conductivity of an impregnated coil wound with wire having a 10% coating would be increased up to 2 $(10)^{\frac{1}{2}}$ or 6·3 times. However, it is probably preferable in most cases to relate t to the space filling factor which can be deduced from the coil resistance. Thus for the rectangular array of Fig. 1 (a) we have

$$s = \frac{\pi a^2}{4\,(a + t)^2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (13)$$

and using Equ. (12)

$$K = \frac{2}{\sqrt{\left(\frac{\pi}{4s}\right)^2 - 1}} \quad .. \quad .. \quad .. \quad .. \quad (14)$$

This is shown in Fig. 2. Filling factors greater than $\pi/4$ cannot be achieved with cubic close packing, whereas theoretically fillings as high as $\pi/2\sqrt{3}$ or 0·908 could be achieved with hexagonal close packing. In this case the $\pi/4$ factor in Equ. (13) is replaced by $\pi/2\sqrt{3}$, the functional dependence remaining unchanged. Thus Equ. (12) can be substituted, giving

$$K = \frac{2}{\sqrt{\left(\frac{0 \cdot 908}{s}\right)^2 - 1}} \quad .. \quad .. \quad .. \quad .. \quad (15)$$

Although this substitution is dubious as Equ. (12) is based on cubic close packing, the prediction is in the right direction. This can be seen by imagining alternate layers in the rectangular close-packed arrangement to be translated parallel to themselves a distance equal to half the centre-to-centre spacing of the wires. This gives the hexagonal close-packed structure without changing the filling factor and the nearest points of the layers become separated. One would expect this to reduce the conductivity as Equ. (15) indicates. Thus the shaded area of Fig. 2 should cover cases met in practice. The importance of obtaining high packing is obvious but, in practice, this tends to conflict with the equally important requirement of full impregnation.

End Cooling

If the coil is of length L in the z direction and the origin of z is taken at the centre plane of the coil then the temperature distribution obtained by integrating Equ. (4) for an end temperature θ_0 is

$$\theta - \theta_0 = \frac{1}{\alpha}\left[\frac{\cos(\gamma z)}{\cos(\gamma L/2)} - 1\right] \quad .. \quad .. \quad .. \quad (16)$$

where γ is determined by

$$\gamma^2 = i^2 \rho_0 \alpha / k \quad .. \quad .. \quad .. \quad .. \quad .. \quad (17)$$

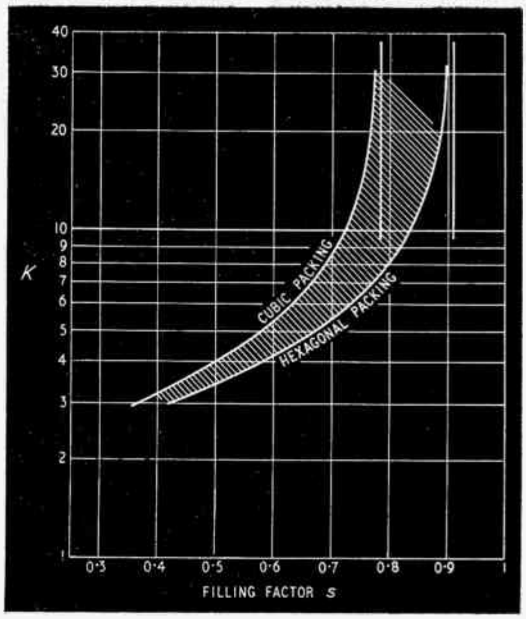


Fig. 2. *Thermal conductivity increase of wire-wound coil as function of filling factor*

The resistivity variation is thus

$$\rho = \rho_0 \frac{\cos(\gamma z)}{\cos(\gamma L/2)} \quad .. \quad .. \quad .. \quad .. \quad (18)$$

Using this in Equ. (2) leads to

$$W_e = (0 \cdot 796B)^2 L \cdot \left(\frac{(\rho_0 \pi)}{s}\right) \cdot \left(\frac{R+1}{R-1}\right) \cdot \left(\frac{\tan\phi}{\phi}\right) \quad .. \quad (19)$$

where ϕ has been written for $\gamma L/2$.

This differs from the expression neglecting temperature coefficient [Equ. (3)] by the introduction of the factor $\tan\phi/\phi$ which tends to unity if the temperature coefficient is small. ϕ is related to the temperature difference $\Delta\theta$ $(= \theta_i - \theta_0)$ between the end of the coil and the centre by

$$\phi = \sec^{-1}(1 + \alpha\,\Delta\theta) \quad .. \quad .. \quad .. \quad .. \quad (20)$$

as can be seen by putting $z = 0$ in Equ. (16)

As yet $\Delta\theta$ is undetermined and a further equation is needed. This is obtained by equating the total power generated W to that leaving the ends of the coil.

$$W_e = 2k \cdot A\left(\frac{d\theta}{dz}\right) L/2 \quad .. \quad .. \quad .. \quad .. \quad (21)$$

where A is the area of each end of the coil, $\pi r_1^2\,(R^2 - 1)$. The temperature gradient here is obtained by differentiation of Equ. (16) and the final result is

$$\boxed{W_e = \frac{8kA}{L}\left(\tfrac{1}{2}\phi \cdot \tan\phi/\alpha\right)} \quad .. \quad .. \quad .. \quad (22)$$

It can be seen from Equ. (20) that for small temperature differences

$$\alpha \cdot \Delta\theta \approx \tfrac{1}{2}\phi^2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (23)$$

and the bracketed term in Equ. (22) tends to $\Delta\,\theta$ which gives the case when temperature coefficient is neglected; i.e.,

$$W_e = \frac{8kA}{L}\Delta\theta \quad .. \quad .. \quad .. \quad .. \quad .. \quad (24)$$

The problem is wholly specified by Equ. (19) and

Equ. (22) and in principle ϕ could be eliminated between them to give the field B as a function of the applied power W. However, the elimination cannot be carried out algebraically; as the temperature difference is of great significance and determines the maximum rating of the coil it is preferable to calculate in terms of ϕ or $\Delta\theta$ using Equ. (20).

Dividing Equ. (19) by Equ. (22) and solving for B gives

$$\boxed{B = 1{\cdot}777 \cdot \frac{(R-1)\cdot D}{L}\left(\frac{sk}{\rho_0}\right)^{\frac{1}{2}} \frac{\phi}{\sqrt{2\alpha}}} \quad .. \quad .. \quad (25)$$

The corresponding formula for α zero is

$$B = 1{\cdot}777\,\frac{(R-1)\,D}{L}\cdot\left(\frac{sk\,\Delta\theta}{\rho_0}\right)^{\frac{1}{2}} \quad .. \quad .. \quad (26)$$

As the temperature coefficient of resistivity of metals is virtually independent of the metal and corresponds to the resistivity increasing linearly from absolute zero, universal curves can be plotted against maximum internal temperatures for various external temperature by putting

$$\alpha = 1/(\theta_0 + 273) \quad .. \quad .. \quad .. \quad .. \quad .. \, (27)$$

This has been used in calculating $(\frac{1}{2}\phi \tan\phi)/\phi$ for use in Equ. (22) and $\phi/\sqrt{2\alpha}$ in Equ. (25) for Fig. 3. The curves labelled (1) correspond to $\alpha = 0$; i.e., the neglect of temperature coefficient of resistivity. It can be seen that compared with the practical case of $\alpha = 1/293$, errors of 10% in dissipation and 20% in field would be introduced by neglecting the temperature coefficient, for internal temperature differences of 400° to 600° C. Normally, internal temperature differences greater than 200° C cannot be tolerated in wire-wound coils and so errors due to neglect of temperature coefficient of resistivity can often be neglected in comparison with uncertainties of the effective thermal conductivity.

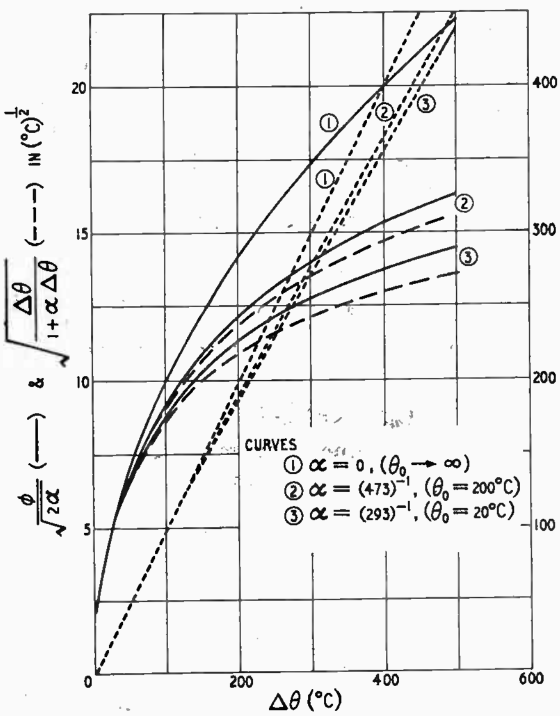


Fig. 3. *Dissipation and field functions against internal temperature difference*

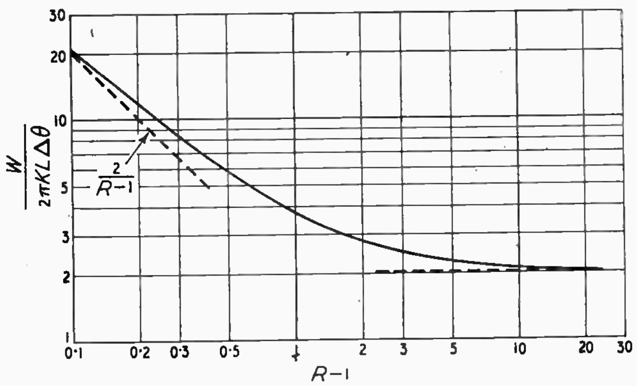


Fig. 4. *Maximum rating of radially-cooled solenoid against radius ratio*

Radial Cooling

The most important practical case is where the outer cylinder is cooled leaving the innermost cylinder at the highest temperature of the system. The exact case can be evaluated in terms of Bessel functions but is laborious, and in view of the small correction for temperature coefficient found in the end-cooled case would not be worthwhile. With constant resistivity the result is

$$W = \frac{2\pi\cdot kL\cdot\Delta\theta}{F(R)} \quad .. \quad .. \quad .. \, (28)$$

$$\text{where } F(R) = \tfrac{1}{2} - \frac{\log_e R}{(R^2-1)} \quad .. \quad .. \quad .. \, (29)$$

If some correction for the temperature coefficient is required, it is reasonable to replace $\Delta\theta$ by the bracketed term in Equ. (22). This procedure is justifiable on the grounds that it becomes exact as R tends to unity, as can be seen by taking the limit of $F(R)$ as R tends to 1 and comparing the result with the linear case of the last section with suitable rotation of co-ordinates. The result of Equ. (28) is plotted against R in Fig. 4. The elimination of W between Equs (29) and (3) gives the field temperature relation,

$$\boxed{B = 1{\cdot}777\left[\left(\frac{R-1}{R+1}\right)\frac{1}{F(R)}\right]^{\frac{1}{2}}\left[\frac{sk\;\Delta\theta}{\rho_0}\right]^{\frac{1}{2}}} \quad .. \, (30)$$

End Cooling versus Radial Cooling

In a wire-wound solenoid where the radial and axial thermal conductivities are the same, the results of the previous sections can be used to compare radial- and end-cooling with a view to determining the boundary between geometries where a change should be made from cooling the ends to cooling the outer cylinder or vice-versa. For the same internal temperature rise $\Delta\theta$ the maximum powers which can be applied in the two cases are given (neglecting the temperature coefficient correction) by Equs (24) and (28). From Equ. (3) the flux is proportional to the square root of the power and thus the ratio of the available fields from the end-cooled case to the radially-cooled case, is

$$\frac{B_e}{B_r} = (D/L)\;[(R^2-1)\cdot F(R)]^{\frac{1}{2}} \quad .. \quad .. \quad .. \, (31)$$

where D is the internal diameter. For a range of L/D this expression is shown in Fig. 5. It can be seen that for what one might call normal radius ratios from 2 to 4, end-cooling is only superior for coils which are significantly shorter than their internal diameter, though this disadvantage disappears if the radius ratio is large enough.

Approximation for Simultaneous Radial- and End-Cooling.

In the end-cooled case the highest temperatures occur in a disc transverse to the axis mid-way along the coil while, in the radially-cooled case, they occur on the innermost cylinder. It has been seen that the correction due to temperature coefficient is small in the axial case, and it follows that it will be negligible for practical purposes with both forms of cooling applied, for the temperature will only be high where the disc and the cylinder overlap, which is a small fraction of the total coil volume.

Thus a crude approximation to the rating of a doubly-cooled solenoid can be made in any given case simply by adding the wattages obtained from a given internal temperature, from Equs (24) and (29). The square of the ordinate on Fig. 5 gives the ratio of the two contributions and thus only the simple end-cooling formula need be used.

Partitioned or Stacked Coils

A method of reducing the internal temperature of wire-wound solenoids is to insert partitions perpendicular to the axis, to conduct heat radially (Fig. 6). Such a coil would normally be constructed by stacking a number of separate coils each with conducting end plates. The partitions will then be twice the thickness of the end plates. To be effective this method must use a reasonable number of partitions, and each individual coil will then be short axially. The system can then be analysed by considering the heat flow to take place axially in the coil and radially in the partitions. Thus a combination of the radial and axial solutions given in earlier sections enables the temperature difference between the outside edge of the fins and highest temperature in the coil to be calculated.

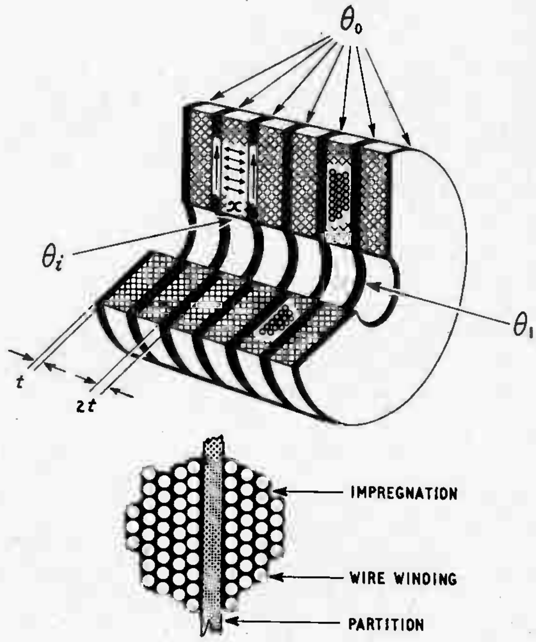


Fig. 6. *Partitioned wire-wound solenoid*

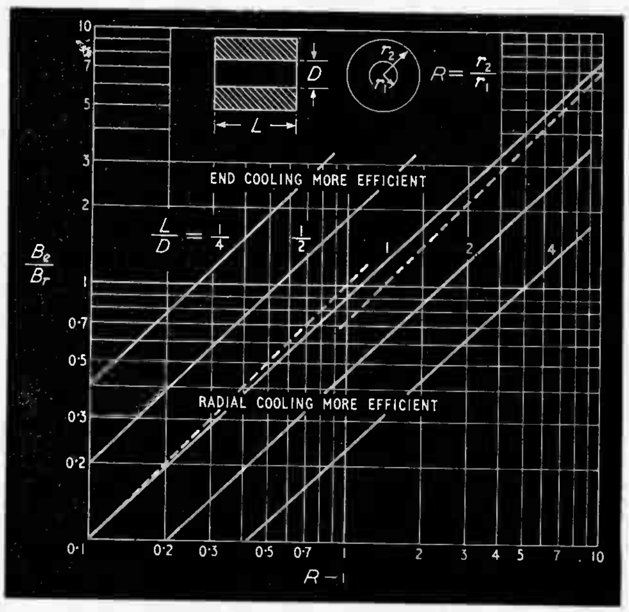


Fig. 5. *Ratio of maximum fields obtainable from end-cooled and radially-cooled solenoids*

If there are n coils the radial temperature difference in the partitions $\theta_1 - \theta_0$ is given by Equ. (28) with the substitutions :—

$W_n/2_n$ for W
t' for L
k_m for k

where W_n is the total power for the partitioned case, t' is the plate thickness at the ends of the separate coils and k_m is the metal-plate thermal conductivity. Thus

$$\theta_1 - \theta_0 = \frac{W_n}{n} \cdot \frac{F(R)}{4\pi k_m t'} \quad \ldots (32)$$

$\theta_i - \theta_1$ is derived from Equ. (24) with the substitutions of W_n/n for W and L/n for L. As these terms occur as a product this temperature difference falls inversely as the square of n. Thus

$$\theta_i - \theta_1 = \frac{W_n}{n^2} \frac{L}{2k\pi\,(R^2-1)\,D^2} \quad \ldots (33)$$

Adding the two temperature differences we obtain

$$\Delta\theta = W_n \left[\frac{F(R)}{n \cdot 4\pi k_m t'} + \frac{L}{n^2 \cdot 2\pi k(R^2-1)D^2}\right] \quad \ldots (34)$$

The wattages for the same temperature difference may be compared with the unpartitioned radially-cooled case of Equ. (28).

$$W_n/W_e = \frac{1}{(k_i K/k_m)\,(L/2nt') + \dfrac{(L/D)^2}{n^2(R^2-1)\,F(R)}} \quad (35)$$

The second term in the denominator is $1/n^2$ times the reciprocal of the square of the ordinate in Fig. 5 and may thus be read off for the required geometry. For $R=3$ and $L/D=1$ it is about unity. The first term in the denominator can be evaluated as a constant, because $2nt'/L$ is the fractional filling of the coil by the metal partitions. This fraction cannot be allowed to be large for two reasons. First it detracts from the filling factor and secondly tends to introduce fluctuations in the axial field owing to the longitudinal disturbance of the current density[3]. If $L/2nt'$ is taken as 10, then the filling factor reduction will not be more than 20% for practical coils and the axial disturbance would be negligible for reasonable n. The conductivity ratio between copper and insulating materials used for impregnation, e.g., epoxy or polyester resins, is about 2,000 to 1 and so taking into account the K factor of Fig. 2 for $s=0{\cdot}7$ which is about 6, Equ. (35) becomes

$$W_n/W_r = \frac{1}{0{\cdot}03 + (1/n^2)} \quad \ldots \quad (36)$$

Thus if a large number of partitions are used, resulting in the temperature drop all occurring in the fins the improvement in rating is 33 to 1. However, a large number of fins is impracticable as the coil becomes very difficult to wind. Furthermore each separate coil must have thick enough plates to make it self-supporting. From Equ. (6) it can be seen that dividing into four gives a rating improvement of 10·8 : 1, into five, 14·3 : 1, and into six, 17·3 : 1. The corresponding field improvements for the same internal temperature will follow the square roots of these numbers.

Interwound Foil Cooling

This is the axial counterpart of the preceding section. Copper foil is interwound between the layers periodically and joined to cooled end-plates. It has the advantage that the coil can be divided into a considerable number of concentric sections (even up to the number of layers) without winding difficulties (See Fig. 7).

Using the method of the last section, except that this time the answer is compared with the end-cooled coil of the same geometry

$$W_n/W_e = \frac{1}{(k_i K/k_m)\cdot(L'/2nt') + \left(\frac{R-1}{4n^2}\right)^2 \cdot (D/L)^2} \quad \ldots \quad (37)$$

The first term in the denominator is the same as that in Equ. (35) and so may be placed equal to 0·03. The second term is likewise similar in being proportional to $1/n^2$ but with a different geometrical dependence. However, for $R=3$ and $L/D=1$ as taken before the equation is the same as Equ. (36). The difference is, however, that n can readily be made large and so the full 33 times improvement in rating and nearly 6 times improvement in field can be approached. Thus this type of coil has much to recommend it, for coils where a high winding resistance is essential. The main disadvantage is the care which must be taken to obtain full impregnation, for the foil impedes the flow of the insulant. This can be overcome by applying it layer by layer during winding, but this is more time-consuming.

Foil-Wound Solenoids

By winding with foil of the full width required for the coil, the thermal conductivity from the winding

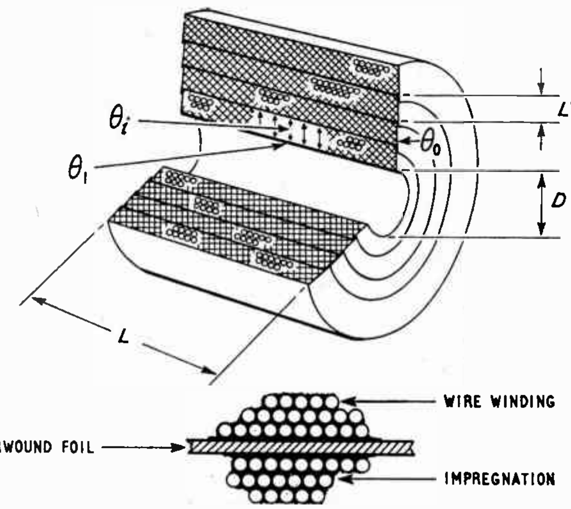


Fig. 7. *Wire-wound solenoid cooled by inter-wound foil*

to the ends is increased by a large factor over the wire-wound case. This proposal was made in connection with aluminium foil[4] for achieving light-weight electromagnets. It was also suggested that holes and slots may be machined in the coil but this requires special after-treatment[5].

The major temperature gradient in such a coil is in the insulating layer which must be provided at the ends of the coil (see Fig. 8). Next to this there is a temperature variation following the usual cosine law* along the foil while the radial gradient is the same as that in the end plates and is normally negligible.

The temperature coefficient of resistivity is not a small correction in foil solenoids as the whole of the metal in the coil is at a high temperature. In cases where the foil drop is small, the correction merely necessitates the use of the high temperature resistivity in the power-field law of Equ. (3).

Thus

$$W = (0{\cdot}796B)^2 L \cdot \frac{\rho_0 \pi}{s}\left(\frac{R+1}{R-1}\right)\cdot(1+\alpha\Delta\theta) \quad \ldots \quad (38)$$

The filling factor s enters into the foil heat flow equation owing to the fact that the cross-section is not fully metal. Thus along the foil

$$\theta_i - \theta_1 = \frac{W\cdot L\cdot}{8k_m A\cdot s} \quad \ldots \quad (39)$$

and across the end barrier of thickness h

$$\theta_1 + \theta_0 = \frac{W\cdot h}{8k_i A\cdot} \quad \ldots \quad (40)$$

Strictly speaking, if the barrier layer thickness h is comparable with the inter-foil insulation thickness, some factor between $1/s$ and unity should appear in the right-hand side of Equ. (40) but this will be ignored. Thus in terms of the total temperature difference

* This is not quite exact as the current can redistribute laterally in a foil solenoid. The rise in temperature and resistivity at the centre increases the current density (and thus the field) towards the ends. Thus the foil drop *must* be kept small for uniform fields. Under these conditions the cosine variation of resistivity gives a good estimate of the current uniformity even though the non-uniformity is neglected in deriving the temperature variation. The percentage non-uniformity of current or field for aluminium windings will be greater than in copper due to the poorer thermal conductivity of the metal. In fact, the percentage current non-uniformity is just $\alpha\Delta\theta$ times the second term of the denominator of Equ. (41).

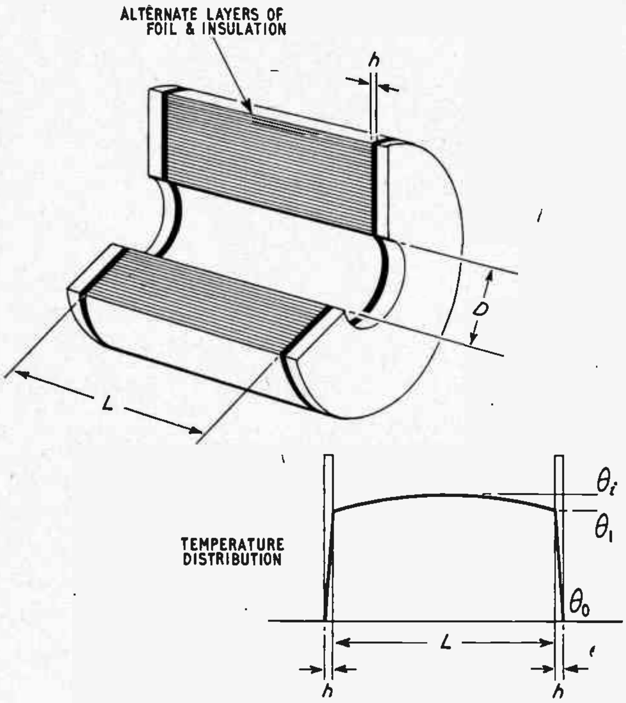


Fig. 8. *Foil-wound solenoid*

$$\boxed{W = \frac{8k_t A}{h} \cdot \frac{\Delta\theta}{\left[1 + \left(\frac{k_t}{k_m}\right)\left(\frac{L}{h}\right)\frac{1}{s}\right]}} \quad \ldots \quad \ldots (41)$$

The temperature variation along the foil is significant if the second term in the bracket of Equ. (41) is significant compared with unity. For this to be so, L/h must be comparable with k_m/k_t.

As used previously k_m/k_t is about 2000 if the metal is copper and 1000 if it is aluminium. Thus if 0·5 mm is the thickness of the end layer and $s=0{\cdot}8$, the foil drop is greater than the end drop if the coil is longer than 125 cm for copper and 62·5 cm for aluminium. These are large coils by normal standards and so the end layer drop is normally the most important.

Using Equ. (41) a simple comparison can be made with the same geometry of end-cooled wire-wound solenoid, described by Equ. (24).

$$W_f/W_e = \frac{\Delta\theta_f}{\Delta\theta_w} \cdot \frac{(L/h) \cdot (1/K)}{\left[1 + \left(\frac{k_t}{k_m}\right)\left(\frac{L}{h}\right)\frac{1}{s}\right]} \quad \ldots \quad \ldots (42)$$

where $\Delta\theta_f$ and $\Delta\theta_w$ are the maximum internal temperature differences which can be tolerated in the foil and wire-wound cases. Foil solenoids are often wound with a polyester film insulant which can be obtained in small thicknesses and melts at about 240°C. In this case $\Delta\theta_f/\Delta\theta_w$ would be about unity. If an anodized film were used then thermally a considerable improvement would be possible.

However, taking $\Delta\theta_f/\Delta\theta_w$ as unity, $K=6$ for the wire, $s=0{\cdot}8$ for the foil, and $L/h=200$, e.g., a 10 cm long coil with 0·5 mm end layer, we have

$$\begin{aligned} W_f/W_e &= 29{\cdot}6 \text{ for copper foil,} \\ &= 26{\cdot}7 \text{ for aluminium foil,} \quad \ldots \quad \ldots (43) \end{aligned}$$

the slight difference being due to the larger foil-temperature drop in the aluminium case (amounting to 20% of the total drop). These figures are a little *less* than can be achieved by infinite interwinding of foil in a wire-wound coil using 10% of winding space for cooling. As nearly all the winding space is used for cooling in the foil solenoid one would expect a *ten times improvement* over a 10% interwound coil for an indefinitely thin layer at the end of the foil. Thus it can be seen that a considerable thermal capacity is thrown away by the necessary end layer. However an insulating layer less than 0·5 mm thick is by no means difficult to prepare on the ends of foil solenoids. Layers as much as ten times thinner are readily prepared by initially machining the end of the coil, etching away the ends of the foil a few mils below the insulant, and then filling the surface groove with a resin. Taken to the extreme, this filling can be loaded with alumina powder which is fifty times a better thermal conductor than the usual resin, but the examples given later show that even with the 0·5 mm layer used above, the internal temperature rise is beginning to be unimportant in comparison with the temperature difference needed externally to obtain adequate heat transfer from the end surface.

The field advantage will not be the square root of Equ. (43) as, first, the coils will normally have different filling factors, secondly, the temperature coefficient correction will reduce the field to a greater extent in the foil case, and thirdly the metals may not be the same. The expression is

$$\frac{B_f}{B_w} = \sqrt{\frac{\Delta\theta_f}{\Delta\theta_w} \cdot \frac{(L/h) \,.\, (s_f/s_w)}{(1 + \alpha\Delta\theta_f)K} \, (\rho_{0w}/\rho_{0f})} \quad \ldots (44)$$

For the values used above together with $s=0{\cdot}8$ for the foil-wound and 0·6 for the copper-wire solenoid, the field ratios are 4·7 for copper foil and 3·5 for aluminium foil at low temperatures, the ratios falling slowly with increasing temperatures.

The field temperature equation is obtained by eliminating W between Equ. (38) and Equ. (41),

$$\boxed{B = 1{\cdot}777 \, \frac{(R-1)D}{L}\left(\frac{L}{h}\right)^{\frac{1}{2}}\left(\frac{sk_t}{\rho_0}\right)^{\frac{1}{2}}\left(\frac{\Delta\theta}{1+\alpha\Delta\theta}\right)^{\frac{1}{2}} \frac{1}{\left[1 + \left(\frac{k_t}{k_m}\right)\left(\frac{L}{sh}\right)\right]^{\frac{1}{2}}}} \quad \ldots (45)$$

Solenoids Wound with Coaxial Line

If a solenoid is wound with bare coaxial cable (Fig. 9) instead of wire, then the temperature gradient is almost entirely between the inner conductor and the outer for the outer conductors can be made as one thermally by soldering, or for very high power coils by circulating water between the outer conductors. The main disadvantage is that the space factor is considerably reduced. If b is the outer radius of the coaxial and a the inner radius then s in Equ. (3) must be replaced by $s(a/b)^2$. Furthermore, as in the case of the foil solenoid, the whole of the electrically-conducting metal is at the high temperature and so a factor $(1+\alpha\Delta\theta)$ must appear in Equ. (3). One apparent advantage is that power cable has a gypsum filling allowing temperatures up to the melting point of copper to be sustained. Making the modifications to Equ. (3).

$$\boxed{W=(0\cdot796B)^2\cdot L\cdot\left(\frac{\rho_0\pi}{s}\right)\cdot\left(\frac{R+1}{R-1}\right)\cdot(b/a)^2\cdot(1+\alpha\Delta\theta)} \quad \text{.. .. (46)}$$

The temperature difference between the inner and outer of the coaxial is everywhere the same and proportional to the dissipation per unit length in the inner conductor.

$$\frac{W}{\frac{1}{2}N(R+1)\pi D} = \frac{2\pi k_t\Delta\theta}{\log_e(b/a)} \quad \text{.. .. .. (47)}$$

where N is the number of turns. The number of turns is equal to the filling factor times the sectional area of the coil divided by the area of the coaxial.

$$N = \frac{sL\cdot(R-1)D}{2\pi b^2} \quad \text{.. .. .. .. (48)}$$

where s in the usual filling factor for wires.

Thus inserting this into Equ. (45).

$$\boxed{W = 8\pi kL\Delta\theta\left(\frac{(R^2-1)\cdot(D/b)^2\cdot s}{\log_e(b/a)}\right)} \quad \text{.. (49)}$$

Once again the elimination of $\Delta\theta$ between the field equation Equ. (46) and the thermal equation Equ. (49) is laborious and it is better to work in terms of the temperature difference $\Delta\theta$. Thus, eliminating W,

$$\boxed{B = 1\cdot777\left[\frac{2}{(b/a)\sqrt{\log_e(b/a)}}\right]\left(\frac{(R-1)D}{b}\right)\cdot\sqrt{\frac{s^2k_t\cdot\Delta\theta}{\rho_0(1+\alpha\Delta\theta)}}} \quad \text{.. (50)}$$

The function $[\Delta\theta/(1+\Delta\theta)]^{\frac{1}{2}}$ is shown in Fig. 3 for three different outside temperatures θ_0.

The most important parameters for obtaining the largest field for a given temperature are b/a and b. The nearer b/a can be made to unity the better is the coupling of the dissipation to the outer conductor. Unfortunately, standard power cables are not designed for this application and have b/a not much less than $2\cdot0$. From Equs (47) and (48) it can be seen that if cables could be obtained with b/a equal to $1\cdot2$ then a $3\cdot3$ times improvement in field could be obtained for only $3\cdot8$ times the power input.

Equ. (50) shows how for a given b/a the flux limit can be raised inversely as the outer diameter of the coaxial. This is due to the falling power density per unit length of cable, which is not accompanied by any change in coupling of the power to the outer conductor, the latter depending only on the radius ratio b/a. Commercial cables have a lower limit in the outer diameters of about $0\cdot25$ in.

Comparing Equ. (48) with the wire-wound end-cooled case of Equ. (26) it can be seen that many factors are common, leaving

$$B_c/B_w = \frac{2}{(b/a)\sqrt{\log_e(b/a)}}\cdot(L/b)\cdot\sqrt{\frac{\Delta\theta_c\ 1}{1+\alpha\Delta\theta_c\Delta\theta_w}} \quad \text{.. .. (51)}$$

Although $\Delta\theta$ in the coaxial case may be some twice that for an impregnated coil, the last factor is actually

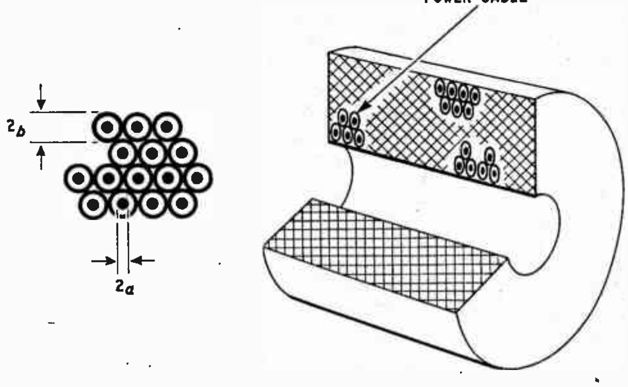


Fig. 9. *Solenoid wound with coaxial power cable*

less than unity for maximum temperature operation owing to the adverse effect of the temperature coefficient. For example, with $\Delta\theta_c = 400°C$. $\Delta\theta_w = 200°$ C and the surrounding at $\theta_0 = 20°$ C so that $\alpha = 1/293$, the factor is $0\cdot92$. Nevertheless a considerable advantage is gained from the other factors; e.g., $L/b = 30$ (corresponding to $0\cdot25$ in. diameter cable and a 10-cm long coil) $b/a = 2$,

$$B_c/B_{we} = 33 \quad \text{.. .. .. (53)}$$

This is seven times better than copper foil using a half-millimetre end layer! It is achieved at the expense of efficiency due to the poor filling factor but enables fields to be obtained which would otherwise not be possible. Further, unlike foil solenoids the axial length is unlimited, either from the heat flow or the winding aspect.

Conclusions and Practical Example

For each of the coil types discussed, an electrical equation has been given relating the field B to the power, followed by a thermal equation relating the power to the internal temperature difference. As the temperature affects the resistivity, it modifies the electrical equation and the field cannot then be obtained directly from it. Thus the power has been eliminated between the electrical and thermal equations to give a field-temperature equation. This, together with the thermal equation specifies the performance of any given coil. They have been framed throughout for easy reference.

Using them, together with Fig. 2 for wire-wound types and Equs (35) and (37) for partitioned or interwound types, a performance chart can be readily calculated for any shape of coil. Fig. 10 is such a chart calculated for

$R = 3$
$L/D = 2$
$L = 10$ cm
$k_m = 4\cdot2$ watts cm/°C for copper
$k_m = 2\cdot1$ watts cm/°C for aluminium
$k_t = 2\cdot1\ 10^{-3}$ watts cm/°C in all cases.
$s = 0\cdot8$ for foil
$= 0\cdot6$ for wire.
$= 0\cdot9$ for coaxial.
$\alpha = 1/293$ (i.e., cooled surfaces at ambient temperature.

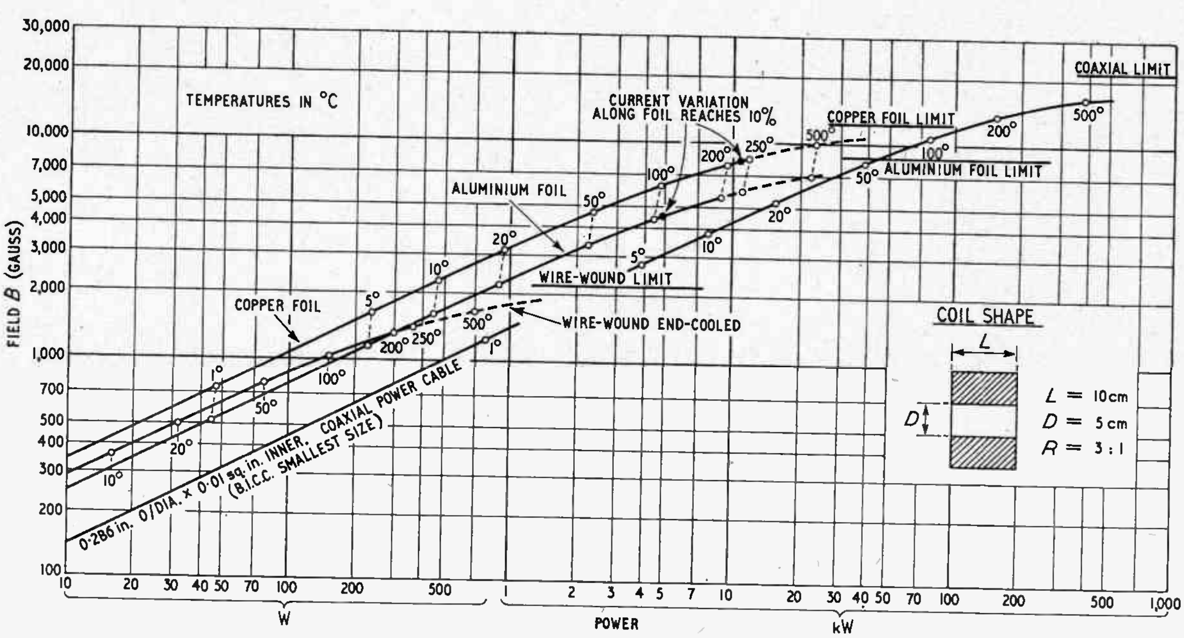


Fig. 10. *Performance chart of a solenoid with various windings*

The performance chart is a field–power graph with the internal temperature differences written along the curves. It is immediately evident that the thermal coupling of the dissipated power to the outside of the coil is so great in the case of coaxial wound coils that tremendous power inputs can be applied without inducing significant temperature differences. For the example shown an applied power of 100 kW would only produce about 120°C temperature rise of the inner conductor of the cable. That this is to be expected can be seen by considering the unwound length of the coil which is 3,430 cm (109 turns).

The power density is thus only 29 watts per cm. To apply such high total power would need an extremely efficient *external* cooling facility. Thus one can never realize the advantage of the refractory nature of the filling.

Fig. 10 also shows that the same shape of coil wound with foil is also extremely well thermally coupled, even with the 0·5-mm end layer assumed. Because of its greater filling factor, more flux can be generated than with the coaxial type for the same power, and temperature differences approaching the melting point of polyester film (250°C) are not set up until 10 kW input are used (giving 8 kilogauss with copper and 5·8 kilogauss with aluminium). However, up to half of this temperature difference might have to be allowed to develop outside the coil with such power, and only some 100°C could then be tolerated internally. This gives a maximum power rating of above 5 kW (1·2 and 4·5 kilogauss respectively). The corresponding wire-wound solenoid, without partitions, could not be driven above 150W and 1,000 gauss without exceeding 100°C internal temperature difference. It can be seen from the field temperature equations, that all coils have an upper limit of field when the temperature is large compared with $1/\alpha$; and these limits are shown on Fig. 10. These can only be approached if temperatures of the order of 500°C can be tolerated.

By comparing temperatures on Fig. 10 of foil and wire at the same power input it can be seen that the foil rating is about thirty times better. However, as discussed above in the section on interwound coil cooling, this difference can be made up by interwinding copper foil taking up 10% of the winding space. In this case, one would need for example 40 layers of 0·005-in. foil. The resulting plot has not been included in Fig. 10 for the sake of clarity but it will follow the copper-foil curve multiplied in field by a factor of 0·54/0·8 or 0·82, the square root of the filling factors. This leaves it better by 10% than the aluminium-foil coil which is 0·75 (the square root of the resistivity ratio) times the copper foil case.

A thinner end layer would make the aluminium foil coil more efficient only at powers exceeding 10 kW where the interwound coil would start to deteriorate in field due to greater temperature rise but, in this case, there would be a difference in current density between the centre and ends of the foil exceeding 25%.

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Conditions for Minimum Variation in a Function—1

There is often a requirement that some quantity associated with a circuit shall be maintained as nearly constant as possible over a given range of frequencies. In filter design, for example, we would like to have the modulus of the transfer function constant at all frequencies within the pass-band, so that the ratio of output voltage to input voltage is constant, and we would also like to have infinite attenuation outside the pass-band. We would also like to have phase varying linearly with frequency within the pass-band. These three requirements are incompatible, and some kind of compromise is in practice usually required. For some applications, constancy of the modulus of the transfer constant is the prime requirement, while for others, it is linearity of phase-change with frequency that matters most.

In many cases, the transfer function will be a polynomial, in the Heaviside operator p, or the reciprocal of such a polynomial. For the purposes of this article, we can replace p by $j\omega$, and the square of the modulus of the transfer function will then be a polynomial in ω^2 at frequency $\omega/2\pi$.

If the modulus is to be kept as constant as possible, its square must also be kept as constant as possible, so our problem is essentially: how do we keep a polynomial 'as constant as possible' over a given range and what is the most effective way to formulate specifically the requirement that the polynomial shall be 'as constant as possible'? We shall, therefore, confine our attention to polynomials in this article. In next month's article, however, we shall extend the ideas used here to the case where linearity of phase with frequency is the prime requirement. The rate of change of phase with frequency is usually a rational function of ω^2 and not a polynomial, but it is still possible by purely algebraic techniques of a reasonably straightforward nature to choose the parameters available so that this rational function shall have approximately the form required.

How then are we to keep the polynomial

$$x^n+a_{n-1}x^{n-1}+a_{n-2}x^{n-2}+\ldots+a_2x^2+a_1x+a_0 \qquad (1)$$

as constant as possible in the range $0<x<1$, say? There is no loss of generality in taking the coefficient of x^n to be unity or the range of x to be from 0 to 1, since we could if we wished replace x by kx where k is at our disposal, and if (1) is multiplied or divided by a constant, any variations in the polynomial (1) in the range $0<x<1$ would be multiplied or divided by the same constant. For detailed work we shall usually take n to be not greater than 4. A higher value introduces unnecessary complications, while for a lower value of n the inevitable variation of the polynomial (1) is considerable.

Now it may happen that for our particular application constancy for low values of x is what matters most, and a variation when x is near 1 is less important. If this is the case, clearly what we must do is to make

$$a_1 = a_2 = a_3 = \ldots = a_{n-1} = 0 \qquad (2)$$

for then the polynomial (1) will only differ from a_0 appreciably when x is very near 1 if n is large. The polynomial will then vary from a_0 to (a_0+1) in the range $0<x<1$, and almost all the variation will take place when x is near 1.

For many applications it will not be the coefficients a_1, a_2, etc. which are given directly in terms of circuit parameters, but these parameters will be more directly related to the algebraic (linear and quadratic) factors of (1). It is therefore worth noting the well-known result that if a_0 is positive and equal to r^n and Equ. (2) applies, the polynomial (1) can always be immediately factorized. These factors are given for n=3, 4, 5, 6, in such a way that they can easily be inferred for higher values of n on the rare occasions when such values occur:—

$$n=3;\ x^3+r^3 = (x+r)\left(x^2-2rx\cos\frac{\pi}{3}+r^2\right) \qquad (3)$$

$$n=4;\ x^4+r^4 = \left(x^2-2rx\cos\frac{\pi}{4}+r^2\right)\left(x^2-2rx\cos\frac{3\pi}{4}+r^2\right) \qquad (4)$$

$$n=5;\ x^5+r^5 = (x+r)\left(x^2-2rx\cos\frac{\pi}{5}+r^2\right)\left(x^2-2rx\cos\frac{3\pi}{5}+r^2\right) \qquad (5)$$

$$n=6;\ x^6+r^6 = \left(x^2-2rx\cos\frac{\pi}{6}+r^2\right)\left(x^2-2rx\cos\frac{3\pi}{6}+r^2\right)\left(x^2-2rx\cos\frac{5\pi}{6}+r^2\right) \qquad (6)$$

In Equ. (3) the second factor could be written (x^2-rx+r^2); in Equ. (6) the middle factor is (x^2+r^2), but these factors have been included as written to make the general result more easily seen. There is always a factor $(x+r)$ initially when n is odd, but this factor is absent when n is even.

If we now regard the polynomial (1) as the square of the modulus M of a polynomial transfer function, so that x is ω^2, it is interesting to note the well-known result that if M^2 has factors

$$\omega^2+k^2,\quad \omega^4-2a^2\omega^2\cos\theta+a^4 \qquad (7)$$

then the corresponding factors of the transfer function itself are

$$p + k, \; p^2 + 2ap \sin \tfrac{1}{2}\theta + a^2 \quad .. \quad .. \quad .. \quad (8)$$

so that, from Equs. (3)—(6) with x replaced by ω^2 we can determine transfer functions associated with 'maximally flat' filters. The proper procedure for realizing a network whose transfer function is known is not considered here ; this is outside the province of a mathematician.

We have now found simple conditions [Equ. (2)] for a polynomial to have 'maximally flat' variation in the range $0 \leqslant x \leqslant 1$, but these conditions will be unsuitable if our requirement is that there shall be as little variation as possible in the polynomial (1) over the whole of that range. Suppose first that the polynomial (1) is only a quadratic

$$a_0 + a_1 x + x^2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Now let X be the variable part of the quadratic (9), so that

$$X = a_1 x + x^2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (10)$$

Then, if $a_1 \geqslant 0$, X increases steadily from zero when $x = 0$ to $(1 + a_1)$ when $x = 1$, and its variation is thus $(1+a_1)$ in this case. If however, a_1 is between 0 and -1, X becomes negative as x increases, reaches a minimum value $-\frac{1}{4}a_1^2$ when x has the positive value $-\frac{1}{2}a_1$, returns to zero when x has the positive value $-a_1$, and finally reaches the positive value $1+a_1$ when $x = 1$. The total variation of X in the range 0 to 1 is thus from $-\frac{1}{4}a_1^2$ (at $x = -\frac{1}{2}a_1$) to $(1 + a_1)$ at $x = 1$, that is

$$(1 + a_1) + \tfrac{1}{4}a_1^2 = (1 + \tfrac{1}{2}a_1)^2 \quad .. \quad .. \quad .. \quad (11)$$

Again, if a_1 is between -1 and -2, X will be always negative, and will have a minimum value $-\frac{1}{4}a_1^2$ when $x = -\frac{1}{2}a_1$, so the variation in X is $\frac{1}{4}a_1^2$.

Finally, if $a_1 < -2$, X continually decreases, and its final negative value is $(1+a_1)$, which gives a range of variation of $-a_1 - 1$ (a positive number greater than 1). From this it is seen that the least total variation occurs when $a_1 = -1$, and has the value $\frac{1}{4}$, whereas in the 'maximally flat' case ($a_1 = 0$) the variation is four times as great. From Equ. (10), the condition $a_1 = -1$ implies

$$X = x(x - 1) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (12)$$

It is convenient to write

$$x = \tfrac{1}{2}(1 - \cos\theta) = \sin^2(\theta/2) \quad .. \quad .. \quad .. \quad (13)$$

so that θ is zero when x is zero, and π when x is 1. Substituting into (12) gives that

$$X = -\sin^2\frac{\theta}{2}\cos^2\frac{\theta}{2} = -\tfrac{1}{4}\sin^2\theta = -\tfrac{1}{8}(1 - \cos 2\theta) \quad .. \quad .. \quad (14)$$

In deriving Equs. (12) and (14) we made no prior assumption about the nature of X; Equ. (12), however, shows that X is unchanged if x is replaced by $(1 - x)$, or that there is (even) symmetry about $x = \frac{1}{2}$. We shall make the assumption that if the polynomial (1) is of higher degree than a quadratic, then it must have symmetry in order to have the least possible variation; the algebra is thus very greatly simplified. When the polynomial (1) is cubic, it is the variation of

$$Y = a_1 x + a_2 x^2 + x^3 \quad .. \quad .. \quad .. \quad .. \quad (15)$$

which we have to consider, and our assumption means that we must work with Y in the form

$$Y = A(x - \tfrac{1}{2}) + (x - \tfrac{1}{2})^3 + (\tfrac{1}{2}A + \tfrac{1}{8}) \quad .. \quad (16)$$

since this last expression has the required symmetry (odd because Y is of odd degree) about $x = \frac{1}{2}$, and has its constant term adjusted so that Y is zero when x is zero as is required in Equ. (15).

Now if A is positive or zero, Y steadily increases with x, so the range of variation of Y is

$$R_1 = Y(1) - Y(0) = A + \tfrac{1}{4} \quad .. \quad .. \quad .. \quad (17)$$

If however A is negative, Y is decreasing when $x = \frac{1}{2}$. For small negative values of A, this decrease will continue until Y falls to a minimum $2\,(-A/3)^{3/2}$ below its value $(\frac{1}{2}A + \frac{1}{8})$ for $x = \frac{1}{2}$; this minimum occurs for $x = 0 \cdot 5 + (-A/3)^{\frac{1}{2}}$. Y thereafter increases with x. Y also, by symmetry, has a maximum value $2\,(-A/3)^{3/2}$ above its value $(\frac{1}{2}A + \frac{1}{8})$ for $x = \frac{1}{2}$; this maximum occurs when $x = 0 \cdot 5 - (-A/3)^{\frac{1}{2}}$. By considering whether the maximum and minimum values occur in the range $0 < x < 1$, and whether they do or do not exceed the values of Y when x is 0 and 1, as in the case of the quadratic already discussed, we find that the least total variation occurs when A is $-3/16$. Substituting from Equ. (13) for this value of A, we obtain

$$Y\,(\text{least variation}) = 1/32\,[1 - \cos 3\theta] \quad .. \quad (18)$$

so that the least total variation of Y in the range $0 < x < 1$ or $0 < \theta < \pi$ is $0 \cdot 0625$, and the values of a_1, a_2 when this least variation occurs are found, by comparing Equs. (15) and (16), to be $0 \cdot 5625$ and $-1 \cdot 5$.

Proceeding similarly when the polynomial (1) is a quartic, we have to consider the variation of

$$Z = a_1 x + a_2 x^2 + a_3 x^3 + x^4 \quad .. \quad .. \quad .. \quad (19)$$

which, by our assumption of (even) symmetry about $x = 0 \cdot 5$, can also be written, in a manner analogous to Equ. (16), in the form

$$Z = (x - \tfrac{1}{2})^4 - B\,(x - \tfrac{1}{2})^2 + (\tfrac{1}{4}B - \tfrac{1}{16}) \quad .. \quad (20)$$

For a small total variation of Z in the range $0 < x < 1$ in Equ. (20), it is easily found that B must be positive. Z then necessarily has a maximum value $(\frac{1}{4}B - \frac{1}{16})$ at $x = \frac{1}{2}$, but if B is small, Z exceeds this value for values of x near 0 and 1. It turns out that the least total variation of Z occurs when $B = \frac{1}{4}$; substitution into Equ. (20) for this value of B from Equ. (13) gives

$$Z = -[1 - \cos 4\theta]/128 \quad .. \quad .. \quad .. \quad (21)$$

and by comparing Equs. (19) and (20), we find

$$a_1 = -0 \cdot 25, \; a_2 = 1 \cdot 25, \; a_3 = -2 \quad .. \quad .. \quad (22)$$

We are now in a position to infer the general law for the least-varying polynomial of any degree : if Z_n is the varying part of the least-varying polynomial (1) of degree n, we infer from Equs. (14), (18) and (21) that

$$Z_n = \frac{(-1)^{n+1}}{2^{2n-1}}\,[1 - \cos n\theta] \quad .. \quad .. \quad .. \quad (23)$$

This result (23) is equivalent to the results obtained by Tchebycheff, which gave rise to the polynomials that bear his name. The preceding argument has not been intended to prove Tchebycheff's results afresh, but merely to bring out their meaning and significance. For Equ. (23) is readily understandable; it implies that if we require a polynomial of degree n to vary as little as possible in the range $0 < x < 1$, the least variation obtainable is $1/2^{2n-2}$, and the polynomial in question will have the maximum possible number $(n-1)$ of turning-values in the range $0 < x < 1$. All the maxima will be equal, and all the minima will be equal, and the extreme values for $x = 0$ ($\theta = 0$) and $x = 1$ ($\theta = \pi$) will be equal to the maximum or minimum values. Maxima or minima occur when $\theta = k\pi/n$ (k integer) or $x = \frac{1}{2}[1 - \cos(k\pi/n)]$.

Equ. (23) is thus the most convenient form for expressing the result; we must remember that $\cos n\theta$ is a polynomial of degree n in $\cos\theta$, which from Equ. (13) is $(1 - 2x)$. The Tchebycheff polynomial of degree n is obtained from Equ. (23) by substituting y for $\cos\theta$ in the polynomial for $\cos n\theta$, and therefore has symmetry. Substituting $(1 - 2x)$ for $\cos\theta$ would enable us to determine the coefficients a_1, a_2, etc., of the polynomial (1) so that it should have the least possible variation, but the values of the coefficients by themselves would not make that symmetry obvious.

If we now compare the variation obtained by the 'maximally-flat' approach [Equ. (2)] and the minimum total variation or Tchebycheff approach [Equ. (23)] we see that the total variation is far greater under the maximally-flat approach, especially when n is large. In the general case it may be worth considering replacing x by $y + \frac{1}{2}$ in the polynomial (1) and making as many coefficients of powers of y zero as possible. The polynomial (1) would thus become

$$b_0 + y^n \quad \ldots \quad (24)$$

b_0 being the value of the polynomial (1) when $x = \frac{1}{2}$. The variation of y is from $-\frac{1}{2}$ to $+\frac{1}{2}$, so the variation of the expression (24) is $1/2^n$ when n is even and $1/2^{n-1}$ when n is odd. But if x is ω^2 in the polynomial (1), we shall have to replace ω by $z + 1/\sqrt{2}$ in the polynomial (1), and this will give us a polynomial of degree $2n$ in z instead of n, in which both even and odd powers of z will appear. This polynomial in z will therefore only be free from powers of z below z^n when we impose the maximally-flat conditions, and the reduction in variation achieved by changing the origin will be correspondingly less than in the case (24). No corresponding difficulty occurs in connection with the Tchebycheff approach.

We have thus obtained two approaches to the problem of obtaining a polynomial which shall vary as little as possible in a given range. The maximally-flat approach is advantageous when variations at one end of the range are important, but the variation over the whole range is much less under the Tchebycheff system.

Linear Frequency Discriminator for Sub-Carrier Frequencies

By P. Kundu*

SUMMARY. *The discriminator described is suitable for use with low-frequency f.m. signals having large deviations, such as are found in f.m. sub-carrier systems. A heptode valve is used, with out-of-phase signals on its two control grids. An RC phase-splitting network is provided for obtaining the correct phase relationship between the two voltages. The conditions for maximum linearity are derived.*

Transmission of information using the f.m. sub-carrier technique is finding wide applications in the field of telephone communication using frequency division multiplexing, in facsimile transmission, and in telemetry. In order that the information may be reproduced faithfully the discriminator must have a linear characteristic at a low frequency with a wide deviation, since the sub-carrier technique, unlike conventional f.m. systems, often uses carrier frequencies as low as few hundred c/s, with deviations as high as 20% of the mean carrier. While a linear demodulating system is essential to avoid signal distortion, stability is a pre-requisite in telemetry applications for the maintenance of calibration.

Of existing methods,[1,2,3] which consist basically of the LC phase discriminator and the pulse-counter type, the former has been widely used for narrow-band applications in the r.f. range, while the latter is finding applications in wide-band systems in the l.f. range. The application of the former in the l.f. range is limited owing to the difficulty of designing coils having a prescribed Q, non-linearity of iron cores, and the adverse effect on Q of the temperature variations, which can give rise to instability of operation. Also, this method requires critical adjustments for optimum performance. The author has not come across any published literature dealing with its suitability as regards linearity and stability of operation over the range of frequency deviation of interest. On the other hand, the pulse-averaging type[3], although it is quite often used for f.m. sub-carrier demodulation, calls for quite involved circuitry for forming the pulses properly and also for maintaining the pulse shape uniformly to avoid non-linearities with increasing carrier frequencies.

Resistance-capacitance frequency selective circuits have also been used by stagger-tuning two parallel-T networks in the feedback path of high gain amplifiers[4]

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to simulate LC discriminator characteristics. The use of RC parallel-T networks, although it eliminates some of the limitations of LC circuits, requires quite a few precision high-stability elements and a number of adjustments to align the networks for the desired operating condition. Moreover, the demodulated output obtained is proportional to the logarithm of frequency and a corrective network must be used to obtain a response which is a linear function of frequency, thereby reducing the sensitivity of the system.

A method based on the principle of phase discrimination is described here, which employs an all-pass RC phase-shifting network with a multigrid demodulator valve. This method makes it possible to obtain an essentially linear output/frequency characteristic over a frequency deviation greater than is actually required for most of the applications mentioned above. While the method avoids the complex circuitry and the number of adjustments needed with parallel-T networks, it eliminates the disadvantages of inductance-capacitance circuits. The discriminator performance compares favourably with existing systems, including pulse-averaging types, as regards linearity, stability and frequency response, while it uses a simpler technique, less components, and is easy to set up.

Theory of Operation

The principle of demodulation utilizes the fact that when a time-delay is imposed on a frequency-modulated signal by passing it through a delay network, the relative phase difference between the delayed and the original signal is a function of the instantaneous frequency. When these two signals are mixed in a valve, the output current will contain a component that is a function of frequency.

In a heptode mixer, a suitable choice of parameters results in a linear relationship between the signal-grid to anode mutual conductance g_m and the oscillator-grid voltage. Variation of g_m due to signal-grid voltage becomes negligible when the input to the grid is very small.

Let the f.m. signal be

$$e_c = E_1 \sin(\omega_c t + m_f \sin pt) \quad \ldots \quad (1)$$

where $\omega_c = 2\pi f_c$, f_c = carrier frequency, m_f = modulation index, $p = 2\pi f_m$, f_m = modulation frequency. The delayed signal is then given by

$$e_d = E_2 \sin[\omega_c t + m_f \sin pt + \phi(\omega)] \quad \ldots \quad (2)$$

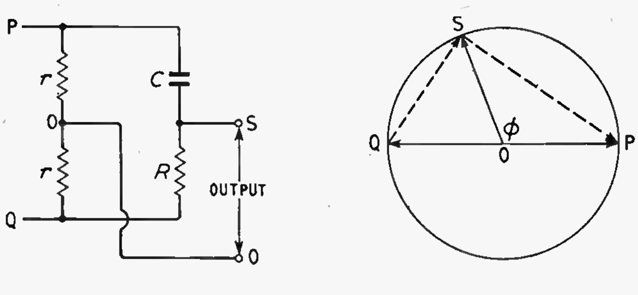


Fig. 1. (*a*) *RC phase-shifting network*, (*b*) *vector diagram of circuit voltages*

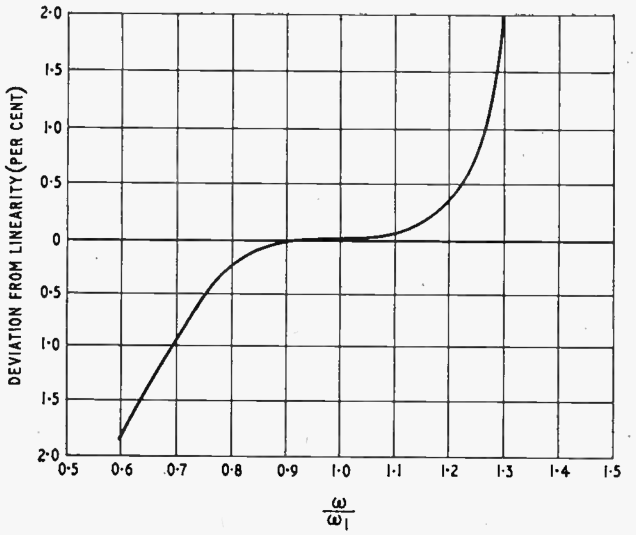


Fig. 2. *Curve showing deviation from linearity against normalized frequency*

Where $\phi(\omega)$ is the phase-shift introduced to the modulated signal and is a function of the instantaneous frequency. When the signals given by equations (1) and (2) are applied to the oscillator and signal grids of the multigrid mixer, the output current is given by

$$i = \frac{KE_1E_2}{2}\left[\cos\phi(\omega) - \cos\{2\omega_c t + 2m_f \sin pt + \phi(\omega)\}\right] \quad \ldots \quad (3)$$

When the high-frequency component is removed by a filter then the output current is given by:

$$i_o = \frac{KE_1E_2}{2} \cos \phi(\omega) \quad \ldots \quad (4)$$

Since $\phi(\omega)$ is a function of the frequency of the modulated signal the output is thus a function of the modulating signal. Ideally, this phase variation should be such that the output varies linearly with the frequency deviation of the modulated signal.

Condition for Linearity

The phase-shifting network used here consists of a simple resistance-capacitance series network connected across a balanced low-impedance source as shown in Fig. 1(a). The output voltage is then taken between earth and the junction of the resistance and capacitance, the vector diagram of the respective voltages being shown in Fig. 1(b). As the frequency of the input signal varies, the magnitude of the output voltage OS remains constant while its phase with respect to the voltage OP is given as:

$$\phi = 2 \tan^{-1} 1/\omega CR \quad \ldots \quad (5)$$

But, as the plate current of the mixer is proportional to $\cos\phi$, the output may be written as

$$e_o = E_0 \cos\left(2 \tan^{-1}\frac{\omega_0}{\omega}\right) \quad \ldots \quad (6)$$

Where ω_0 is given by $\omega_0 = 1/CR$ which means that at this frequency the voltages across the resistor and capacitor are equal and the output of the phase shifter is 90° out of phase with the input. At this frequency the output of the discriminator as given by Equ. (6) is zero, and a balanced output is obtained about this point for small frequency deviations of the input signal. But as the requirement for wide-band cases makes it most desirable to provide a linear change in voltage for a linear change in frequency, it is necessary to determine the optimum operating point satisfying the condition for minimum distortion. This can be accomplished mathematically by equating the second derivative of the output voltage with respect to the input frequency ω to zero to locate the point of inflection.

From Equ. (4),

$$\frac{de_o}{d\omega} = 2\,E_o \sin\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\frac{\omega_0}{\omega^2+\omega_0^2}$$

$$\therefore \frac{d^2e_0}{d\omega^2} = -\sin\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\frac{4\,\omega\omega_0}{(\omega^2+\omega_0^2)^2} - \cos\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\frac{4\omega_0^2}{(\omega^2+\omega_0^2)^2}$$

At the point of inflection $\frac{d^2e_0}{d\omega^2} = 0$

$$\therefore \sin\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\omega + \cos\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\omega_0 = 0$$

$$\text{or, } \sqrt{\omega^2+\omega_0^2}\,\sin\left[2\tan^{-1}\frac{\omega_0}{\omega} + \tan^{-1}\frac{\omega_0}{\omega}\right] = 0$$

whence $3\tan^{-1}\frac{\omega_0}{\omega} = n\pi.$

For $n = 1$, $\tan^{-1}\frac{\omega_0}{\omega} = \frac{\pi}{3}$.

$$\text{i.e., } \omega = \frac{\omega_0}{\sqrt{3}} \quad \ldots \quad (7)$$

$$= \omega_1 \text{ (say)}$$

This means that if the frequency of the f.m. carrier is ω then the *CR* value of the phase-shifter must be adjusted to a value so that $\omega_0 = \sqrt{3}\,\omega$ in order to get maximum linearity in the discriminator characteristics.

The degree of non-linearity about this point of inflection may be determined by the amount of departure of the actual response characteristic from an ideal straight line having the same slope as at this point.

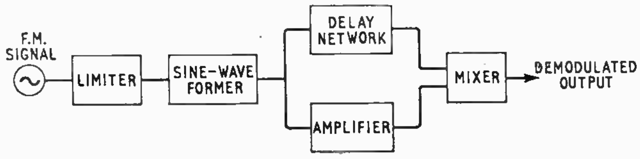


Fig. 3. *Block diagram of a complete discriminator system*

The slope of the response at $\omega = \omega_0/\sqrt{3}$ is:

$$\left(\frac{de_0}{d\omega}\right)_{\omega=\omega_0/\sqrt{3}} = \sin\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)\frac{2\omega_0}{\omega^2+\omega_0^2}$$

$$= \frac{3\sqrt{3}}{4} \quad \ldots \quad (8a)$$

The equation for a straight line passing through this point having the same slope is given by

$$y - y_1 = m\,(x - x_1)$$

$$\text{Or } y = \frac{3\sqrt{3}}{4\omega_0}\left(\omega - \frac{\omega_0}{\sqrt{3}}\right) + \cos\,(120^\circ)$$

$$\therefore y = \frac{3\sqrt{3}}{4}\frac{\omega}{\omega_0} - \frac{5}{4} \quad \ldots \quad (8b)$$

$\therefore$ The percentage of non-linearity is given by

$$\frac{\left(\frac{3\sqrt{3}}{4}\frac{\omega}{\omega_0} - \frac{5}{4}\right) \sim \cos\left(2\tan^{-1}\frac{\omega_0}{\omega}\right)}{\left(\frac{3\sqrt{3}}{4}\frac{\omega}{\omega_0} - \frac{5}{4}\right)} \quad \ldots \quad (9)$$

From this equation the non-linearity in the response may be estimated to determine the maximum frequency deviation over which the system may be used. A curve is drawn in Fig. 2 to show the percentage deviation from linearity of the demodulator output with the frequency deviation about the point of inflection of the response curve determined theoretically. The maximum

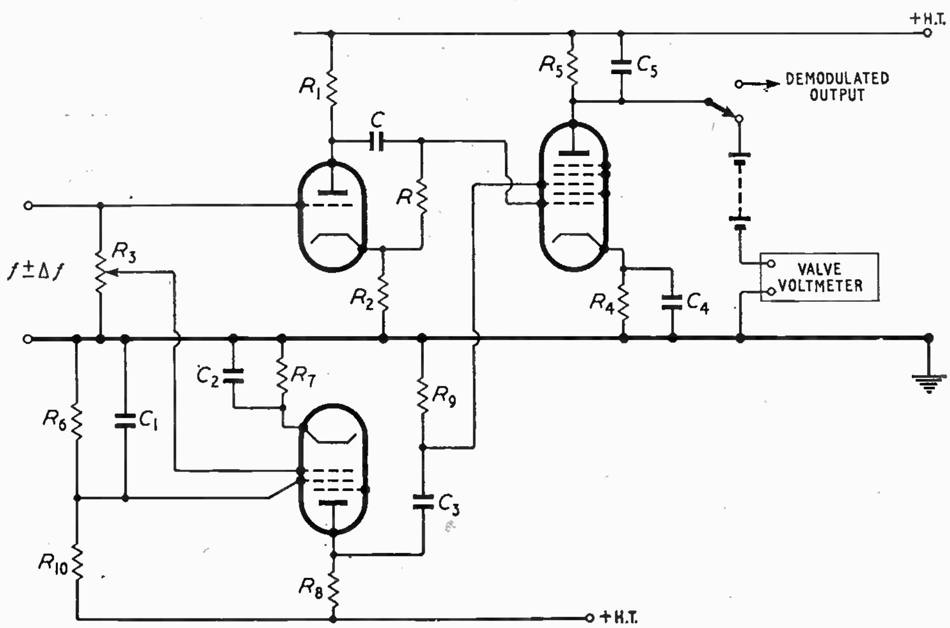


Fig. 4. *Circuit diagram of discriminator*

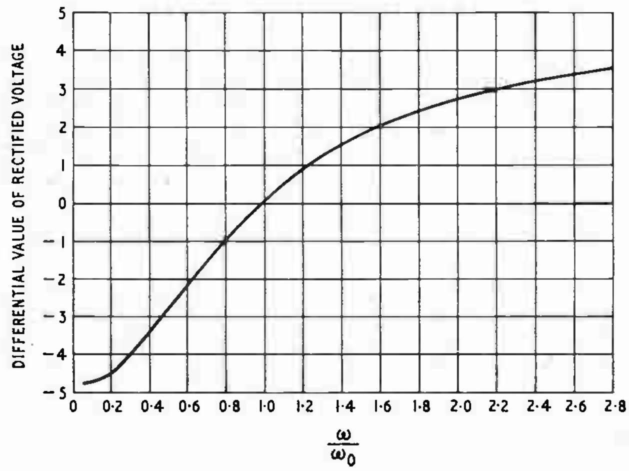


Fig. 5. *Measured discriminator characteristic*

non-linearity remains within 0·4% for a frequency deviation of 20%.

Circuit Details

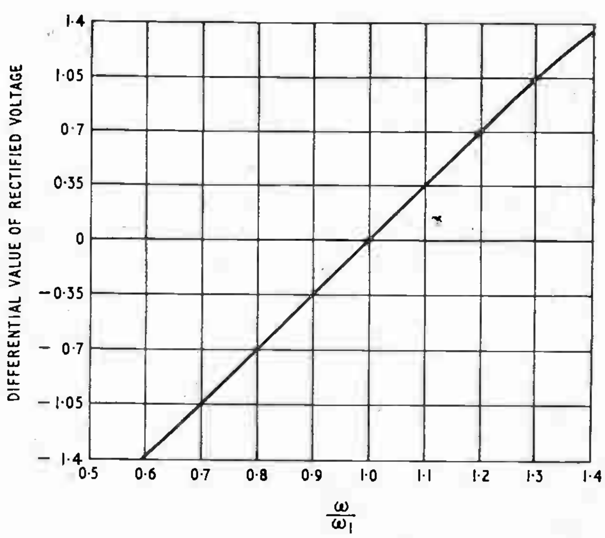
In the block diagram given in Fig. 3 the different parts of the system are shown to illustrate the principle of operation, while Fig. 4 shows the essential parts of the discriminator circuit schematically. The component parts of this circuit consist mainly of a half-lattice *RC* phase-shifting section fed from a balanced phase-inverter stage, an amplifier having a constant gain and negligible phase shift at all frequencies of importance and a heptode mixer stage. The basic design procedure for each of the above stages is quite well defined. The phase-shifting stage is a high-g_m h.f. triode having matched anode and cathode load resistors so that the *RC* network may be fed from a balanced low-impedance source and the output is uniform over a wide frequency range. The amplifier uses an r.f. pentode of large gain–bandwidth product so that the requisite gain with negligible phase-shift is obtained with a resistive load over the desired frequency range. The a.c. signal from the phase-shifter is applied to the signal grid of the 6L7 mixer along with the d.c. voltage developed across its cathode, which is about 5 V. The voltage at the cathode of the mixer is 11·5 V, which makes the signal grid voltage −6·5 V. Further, the design of the circuit is such that the signal and oscillator grid are fed with 0·4 V and 4 V respectively for any input. The applied electrode voltages are not critical, but should preferably be adjusted so as to obtain a linear relation between g_m and e_{g3} together with maximum conversion conductance.

The output of the multigrid demodulator, besides being a function of the phase difference between the signals at the two grids, is proportional to the product of their amplitudes, which makes the detected output proportional to the square of the amplitude of the signal voltages. In order to avoid adverse effects due to envelope variation of the input signal, the discriminator must be preceded by a well-designed limiter and sine-wave former circuit as is the usual practice. It may, however, be mentioned that the presence of a small fraction of harmonics will have a comparatively small effect owing to the square-law action. The detected audio-frequency output voltage is developed across the mixer output load resistance R_5 while the signal frequency and its harmonics are eliminated by C_5. The value of C_5R_5 may be chosen to provide any necessary de-emphasis.

Experimental Results

Suppose the discriminator shown in Fig. 4 is required to be set up for receiving a signal of a particular frequency ω_1. The procedure is similar to the alignment of conventional *LC* discriminators, but the adjustment is much less critical and may be carried out easily with simple equipment. First the value of *CR* must be selected so as to ensure a quadrature phase relation between the input and output of the phase-shifter stage at a frequency $\omega_0 = \sqrt{3}\omega_1$. With an unmodulated oscillator set up at ω_0, the resistance *R* is adjusted until no change is observed at the output voltage indicated by the voltmeter or, in the anode current of the mixer as indicated by a milliammeter, when the signal-grid voltage is switched on and off. With this setting of *R*, observations were made of the voltage variation at the output with input frequency with an input frequency of $\omega_2 = \omega_0$ as well as $\omega_1 = \omega_0/\sqrt{3}$. The linearity of the output and the uniformity of performance were studied using various sub-carrier frequencies. A typical response characteristic is shown in the Fig. 5, and from the nature of the curve it may be seen that the maximum linearity lies about the point predicted theoretically; i.e., if the circuit is aligned at a frequency $\omega_0 = 1/CR$, then the maximum linearity is obtained for an input signal frequency $\omega_1 = \omega_0/\sqrt{3}$. It may, however, be noted that at $\omega_2 = \omega_0$ a balanced output having zero response for carrier may be obtained but the linearity requirement is not satisfied, unless it is used for extremely narrow-band operation or the response is corrected by means of a suitable network. In order to study the characteristic more closely under the maximum linearity condition given by Equ. (7), a response curve

Fig. 6. *Discriminator response over the useful frequency range*



is drawn about the frequency $\omega_1 = \omega_0/\sqrt{3}$. Fig. 6 shows a typical response curve showing the discriminator performance under the condition for maximum linearity. In telemetry applications the stability of the system is also an essential factor for the maintenance of the accuracy of calibration. After allowing time for warming-up of the instrument, the zero shift was found to remain within $\pm 0{\cdot}02$ V for a full-scale reading of $1{\cdot}5$ V. The drift may be due to the frequency drift of the signal generator used, which is specified as $0{\cdot}5\%$. It can be seen from the slope of the response characteristics that a drift of about 35 mV is caused by this amount of instability in the oscillator. The zero drift of the discriminator has, however, been found to be negligible when no input signal is applied. The linearity of response obtained experimentally closely follows that given by the theoretical curve of Fig. 2.

Conclusion

The measured performance of the discriminator agrees closely with theory. It provides an extremely linear characteristic over a wide bandwidth with good stability and sensitivity and thus satisfies the basic requirements of a low-frequency discriminator needed for various applications using f.m. sub-carrier techniques. The overall performance of this system compares favourably with that of existing types, while it is simpler and needs only a few straightforward adjustments.

Acknowledgement

The author records his sincere thanks to Prof. H. Rakshit, D.Sc., for his kind encouragement and for very valuable discussions during the progress of the work. He also acknowledges gratefully some helpful suggestions from Mr. A. K. Chowdbury.

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[4] Paul T. Stine, "Parallel-T Discriminator Design Technique", *Proc. Nat. Elec. Conference*, 1950, Vol. IX, pp. 26-39.

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*The Telegraph Construction and Maintenance Co. Ltd., Mercury House, Theobalds Road, London, W.C.*1.

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

Temperature-Compensated Crystal Oscillators

NEW METHOD OF STABILIZATION

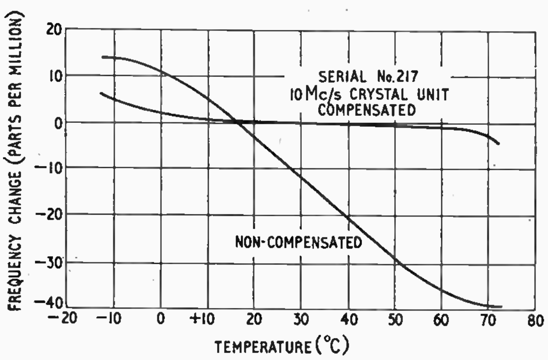
A novel method of keeping the frequency of a crystal oscillator constant despite temperature changes has been developed by the Automatic Telephone and Electric Co. A passive network containing a thermistor, the latter being enclosed in the same envelope as the crystal, and preferably in thermal contact with it, replaces the usual crystal oven.

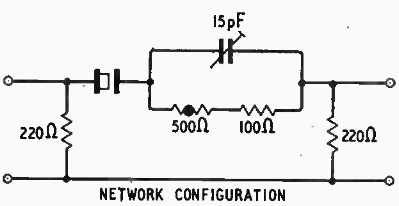
The system makes use of the fact that the frequency of a crystal oscillator depends to some extent on the nature of the electrical load placed across the crystal. The frequency can be stablized by making the load reactance vary in an appropriate manner with temperature. Suitable temperature-dependent reactances do not exist, but the same effect can be obtained by means of a fixed reactance and a bead-type thermistor. The effect of such a network can be seen from Fig. 1, a typical network and the resulting frequency-temperature characteristic being shown in Fig. 2.

The reactive part of the compensating network is either a capacitor or an inductor, depending on the sign of the temperature coefficient of frequency. Where the sign changes with temperature, as with BT-cut crystals, a combination network such as that shown in Fig. 3 may be used.

The size, weight, and power consumption of an oscillator compensated by these means are less than those of an oscillator using a crystal oven. A typical unit is shown in Fig. 4. The absence of contacts should result in better reliability.

Fig. 1. *Frequency drift of a 10-Mc/s crystal oscillator with and without compensation*





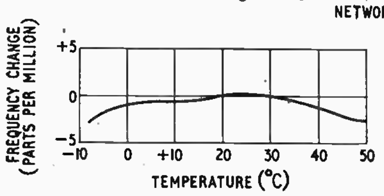


Fig. 2. *Frequency drift of a 37-Mc/s oscillator using the network shown*

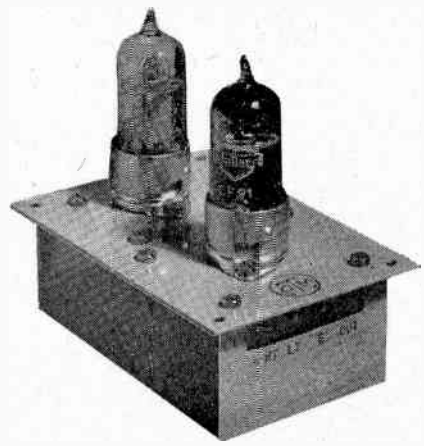


Fig. 4. *Thermally-compensated crystal oscillator* (*Courtesy Automatic Telephone and Electric Co., Ltd.*)

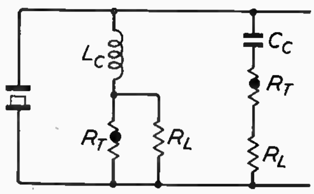


Fig. 3. *Combination network for BT-cut crystals L_C, C_C are the compensating reactances, and resistors R_L define the limits of compensation*

Design Considerations

The rate of change of crystal-oscillator frequency for a change in crystal-load capacitance is given by:—

$$df = \frac{-C_1 dC_2 \times 10^6}{2(C_0 + C_2)^2} \text{ parts per million} \qquad (1)$$

where C_0 is the crystal equivalent shunt capacitance,
C_1 is the crystal equivalent series capacitance
C_2 is the crystal load shunt capacitance.

For any one type of crystal element and a fixed load capacitance, Equ. (1) becomes a constant, K, taking dC_2 as 1 pF. If a CR network is used instead of a pure load capacitance C_2, then

$$df \approx -\frac{K\omega C}{2\pi}\left[\sin\left(\tan^{-1}\frac{1}{\omega CR}\right)\right]^2 \text{ c/s} \qquad (2)$$

The part of (2) within the brackets is equal to $\sin^2(\tan^{-1} Q)$ where Q is that of the CR network. When the resistor in the network is a thermistor, the curve of $\sin^2(\tan^{-1} Q)$ changes with temperature in the opposite manner to the change in frequency with temperature of an AT-cut crystal. Temperature compensation can then be achieved by making $Q = 1$ at the centre of the working temperature range. Similar considerations apply when an RL network is used to obtain compensation.

The technique described above has been patented.

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.

Tunnel Effect

SIR,—In various places in physical electronics, as, e.g., in the theory of strong field emission, one encounters the 'tunnel effect' in which electrons penetrate potential barriers although, from simple classical theory, they never would, at least as long as they are considered as charged points.

The following way of looking at the matter, in which note is made of the fact that the electron may actually be thought of as distributed over a certain region, may be of some interest.

It is worth pointing out that, once it is admitted that the 'particle' is a 'distributed' rather than a 'point' object, this result is not really so strange. Consider, for instance, a one-dimensional problem for simplicity. Let a rod of length L have an immobile electrical charge Q uniformly distributed along it, and let its length be in the x direction, in which direction it is moving with a velocity corresponding to a kinetic energy E. The potential energy is zero, except in one region of length $a < L$, where there is an electrostatic potential V_0/Q volts corresponding to a potential energy $V_0 > 0$ for the rod *if it were all inside that region.* However, only a length a will be inside the region at once ; accordingly, the rod will penetrate the potential barrier if $E > V_0a/L$. Formally, one might say that the probability P of penetrating the barrier is given by

$$P = \frac{1}{1 + f\left(\dfrac{V_0 a}{EL}\right)} \qquad \ldots \quad (1)$$

the function f being defined by $f(u) = 0$ for $0 \leqslant u < 1$ and $f(u)$ is infinite for $u \geqslant 1$.

In wave mechanics, if a particle with an energy E and mass m and, corresponding to this, a wavelength $\lambda = h/(2\ m\ E)^{1/2}$ is incident upon a barrier of potential energy $V_0 > E$, the probability of penetration may be written as[1]

$$P = \frac{1}{1 + \left(\pi \dfrac{V_0 a}{E\lambda}\right)^2 \dfrac{\sinh^2\left\{\left(\dfrac{2\pi a}{\lambda}\right)\left(\dfrac{V_0}{E} - 1\right)^{1/2}\right\}}{\left\{\left(\dfrac{2\pi a}{\lambda}\right)\left(\dfrac{V_0}{E} - 1\right)^{1/2}\right\}^2}} \qquad \ldots \quad (2)$$

This resembles Equ. (1) to the extent that near $V_0 a = E\lambda$ there is a rapid transition from P near unity to P very small. Thus the wave-mechanical behaviour of an object in penetrating a potential barrier has a certain similarity to that of an object of finite extent penetrating a potential barrier in the classical theory. This furnishes another illustration of the fact that part of the strange behaviour of 'particles' obeying wave mechanics is due to the fact that they should really be considered as 'distributed' rather than 'point' objects.

Pacific Semiconductors, Inc., H. L. ARMSTRONG.
Culver City,
California, U.S.A.
11th June 1958.

REFERENCE

[1] L. I. Schiff, "Quantum Mechanics" (McGraw-Hill Book Company, Inc., New York 1949), 1st edition, Chapter 5, Section 17. Equ. (2) here is the same as Equ. (17.7) in the reference, but changed to contain λ explicitly.

Parallel Four-Terminal Networks

SIR,—In his article (page 207 et seq) in your June issue, Mr. F. E. Rogers propounds a theorem which has been widely used in network analysis and synthesis, and for which the basic theory is to be found in "Communication Networks," Vol. II, by E. A. Guillemin (John Wiley, 1935). In Chapter IV of that book Guillemin discusses the properties of interconnected four-terminal networks and, among other things, shows that networks connected in parallel may be defined by a set of short-circuit admittance parameters each one of which is the sum of the corresponding admittances of the individual structures. Guillemin is careful to point out, however, that this is only true for restricted classes of networks and is invalid, for instance, for the network shown in Fig. 1. The theorem discussed by Mr. Rogers thus breaks down in this and similar cases.

Mr. Rogers suggests that his theorem may have value in network synthesis; and this is indeed true because it has been used

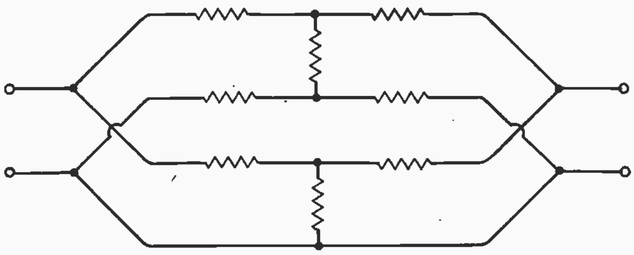


Fig. 1

often since it was first applied to such problems by E. A. Guillemin in a paper on the " Synthesis of R-C Networks", *Journal of Mathematics & Physics*, April 1949.

Electrical Engineering Dept., J. T. ALLANSON.
The University of Birmingham.
11th June 1958.

SIR,—The objective in my article was to set forth and illustrate in simple terms a network property that should be widely known amongst electrical engineers and students in general, yet which does not appear to be thus known; and it has been set in words, because it is believed to be of sufficient importance to warrant such an interpretation. At no point has originality been claimed for the actual relationship derived, for it is so simple that it must have been independently arrived at by many people on many occasions.

Of course, the background theory is contained in Guillemin's classic work, as indeed it is contained merely in the definitions of transfer and output admittance together with Kirchhoff's long established laws. It was not known, however, that the relationship had been used in an explicit form as the basis of a procedure for network synthesis, which has hitherto not been within my closest experience. I am therefore indebted to Mr. Allanson for drawing attention to Guillemin's paper. It is unfortunate, however, that a network property of such simplicity and general electrical interest should be overshadowed by a forest of poles and zeros, in a journal neither very accessible to, nor widely studied by, the average electrical engineer: it is not available in either the I.E.E. or the Patent Office Library, for example. Moreover, Guillemin's interpretation conforms to his special purpose, and the relationship is derived from a nodal equation for the particular case of a 1-ohm termination. He thus obtains (Equ. 5, p. 21, in the paper referred to), for a single network initially,

$$Y_{12} = I_2/E_1 = -E_2/E_1 = \frac{y_{12}}{1 + y_{22}}$$

Here Y_{12} denotes a 'transfer function' that is identified simultaneously with an admittance and a voltage ratio. While numerically correct for the case of 1-ohm termination, and appropriate to Guillemin's immediate purpose, this equation does not bear dimensional examination; and for this reason it does not facilitate recognition of a property of general interest.

It is taken for granted in my article that networks to be connected in parallel will be of types that it is legitimate to parallel; i.e., networks free from structural interaction. The case cited by Mr. Allanson in his Fig. 1 must therefore be regarded as a pointless hypothetical one for, notwithstanding that it is taken from Guillemin's book, it depicts two networks that are not valid for paralleling; the connection of an unbalanced-type network in parallel with a balanced-type network can only result in partial destruction of the latter, by the short-circuiting of a terminal-pair. Some issues of a similar kind are in fact considered in my own book, in relation to the practical interpretation of equivalence between networks (p. 209 et seq, for example). The class of network in mind is, I think, quite clearly conveyed in the introductory section of my article, and by the configurations chosen for illustration. The statement of transfer voltage-ratio in the manner of a theorem might however be improved by including the word *independent*: "... for any number of *independent* four-terminal networks, operating ..."

I would like to take this opportunity of correcting two minor errors which have crept into my article. In Equ. (2) the term $Y_{tn} Y_l$ in the denominator should be $Y_{tn} + Y_l$ and in the numerator of Equ. (21) V_{s2}/Z_{sn} should be V_{s2}/Z_{s2}.

The Polytechnic,
London.
18th June 1958.

F. E. ROGERS

Parallel-T RC Selective Amplifiers

SIR,—With reference to this article by Messrs. Ward and Landshoff (April 1958), the statement is made that "on measurement it was found that the circuit is extremely predictable, errors in the resonant frequency always being less than 1/3%".

The impression is thus conveyed that the authors regard their presented theory as being completely verified merely because the resonant frequency error is less than 1/3%. If, however, the selectivity curve is plotted, not to a decibel scale, but to a linear scale on the ordinate axis, large discrepancies are likely to be found between predicted and experimental results.

This is one measurement which will point out the importance of the loss of the capacitors of the parallel-T network. A full analysis, together with experimental verification, of the effects of this capacitor loss has been presented[1] for the case of the symmetrical form of the parallel-T in which it has been shown that this capacitor loss may be taken up by suitable reduction to the shunt arm resistor. It is therefore suggested that this cause, and not that of shunt component inaccuracies, is that of importance. This is then in contradiction to the authors' statement that "as may be predicted, errors due to inaccuracies in the shunt components were found to have a slightly more serious effect", as even for accurately determined components (of the order of 0·05%) the shunt resistor still differs from its value inducted by a theoretical analysis which neglects capacitor loss.

The value of the statement that "design calculations were made using five-figure logarithms" is thus debatable.

Department of Electrical Engineering,
University of the Witwatersrand, Johannesburg.
23rd May 1958.

K. POSEL

REFERENCE

[1] K. Posel, "The Unbalanced Symmetrical Parallel-T Network taking Account of Capacitor Loss", *Transactions of the South African Institute of Electrical Engineers*, July 1957.

SIR,—I would like to thank Mr. K. Posel for drawing our attention to his paper, which had unfortunately escaped our notice.

It is difficult to understand how, by plotting our selectivity curve to a linear scale in the y axis, discrepancies between measured and calculated amplifier performance will suddenly appear; surely the scale to which a graph is plotted cannot make any difference to the results it conveys provided it is corrctly interpreted. In any case, we did not rely on a plot of the selectivity curve to find the resonant frequency, which was measured by applying the input and output of the selective amplifier to the direct-coupled X and Y deflection amplifiers of an oscilloscope and adjusting the input frequency to give 180° phase shift. a_0 was measured by comparing the input and output amplitudes at this frequency and the value of Q was found by observing the two frequencies at which the output had fallen to $1/\sqrt{2}$ of its value at resonance. These last two measurements correspond with the use of the linear ordinate scale which Mr. Posel advocates.

In his paper, Mr. Posel treats the symmetrical form of the filter in which $K = 1$, $R_a = 0$, $R_0 = 0$. Therefore, from our Equ. (11a) $n = 2$ and from Equ. (10) $\omega_0 = 1/CR$. He finds that, owing to capacitor losses, with this value of n, the transmission of the network does not reach zero at any frequency, unless the shunt-arm resistor, R_1, satisfies the equation $R_1 = R/2 - 2R^2/R_s$, in which case zero output occurs at $\omega = \omega_\infty = \frac{1}{CR}\left(1 + \frac{2R}{R_s}\right)^{\frac{1}{2}}$, where R_s is the loss resistance of either of the series capacitors, C. If $R/2 = R_1(1 + d)$, the network transmission goes through a minimum at $\omega = \omega_m = \omega_\infty \left[1 - \frac{d}{4}\left(1 - \frac{3R}{R_s}\right)\right]$ making $a_0 = \frac{A}{1 + Ad/8}$.

Mr. Posel's equations can be rewritten in terms of the power factor ($\cos\phi$) of C. From the above $d = \frac{4R}{R_s(1-4R/R_s\lambda)}$. Substituting this into the equation for ω_m, eliminating ω_∞, simplifying and neglecting powers of R/R_s higher than the first, gives $\omega_m = \omega_0$. Now $\cos\phi = 1/\omega_m C R_s$, hence $R_s = 1/\omega_m C \cos\phi = R/\cos\phi$. Therefore $R_1 = \frac{1}{2}R(1-4\cos\phi)$ and $A/a_0 = 1 + \frac{1}{2}A\cos\phi$.

Since stability of the component values is of the utmost importance only high-stability capacitors would normally be used in the construction of the parallel-T network of a selective amplifier. These are available in mica or polystyrene film dielectrics in values up to 4 μF, with power factors from 0·0001 to 0·0005 for values up to 0·1 μF, rising to about 0·001 for higher values.

If we assume a selective amplifier $A = 300$, with a symmetrical network, in which $\cos\phi = 0{\cdot}0005$, we have the following errors: $\omega_m = \omega_0$; i.e., no error in centre frequency; $A/a_0 = 1{\cdot}075$ or 7·5% error; $R_1 = (1{\cdot}002)\ R/2$, an error of 0·2% in the value of the shunt resistor.

Even if we neglect the most important consideration, that of resonant frequency accuracy, the fact that an error of 0·2% in the value of the shunt arm resistor alone (or in our treatment 0·1% in each of the shunt components) can cause a change of 7·5% in gain, underlines the necessity of using 5-figure logarithms in the design of the network. Being well aware, from Equ. (14), of how rapidly a_0 changes with n, we neither expected, nor claimed, a high degree of coincidence between predicted and measured values of a_0 and Q. It was noticed that experimental values of a_0 were more frequently on the low side of A than on the high; although no special significance was attached to this at the time, it was most probably due to the effect described in Mr. Posel's paper.

Rank Cintel Ltd.,
Lower Sydenham, London, S.E.26.
24th June 1958.

J. J. WARD

Manufacture of Silicon Transistors

SIR,—In your June issue there is an article on the manufacture of silicon transistors by Dr. Kendall of Texas Instruments Limited. In the first paragraph on page 206 there is a statement: "One serious drawback to the alloyed silicon transistors made to date is their mechanical instability when subjected to temperature cycling."

Table 2, on the same page, indicates only one other manufacturer besides ourselves of this type of transistor. This other company's products are, in the text of the article, explained not to suffer from the defect in question. The direct implication is that the Mullard silicon alloyed transistor types OC 200 and OC 201 do, whereas in our life tests this has not been the case.

In addition I should like, on the basis of considerable experience of the mass production of silicon alloy transistors, to make the following comments:

The alloying temperature for aluminium–silicon as used in production gives a few per cent difference in solubility as compared with indium–germanium. We do not find the temperature of alloying a major difficulty, as stated by Dr. Kendall. It may in any case be overcome by the use of aluminium–silicon alloys for the emitter and collector; these alloys will have smaller penetrations into the silicon.

An aspect of technology not mentioned by Dr. Kendall is the method of making contact to the aluminium. Aluminium is, of course, not easy to solder to and the method of connection may alter the technique from equilibrium alloying to non-equilibrium alloying using, for example, aluminium wires. This fact is important as it influences the whole design of an alloy transistor.

Referring to the doubt Dr. Kendall throws on the thermal stability of aluminium silicon alloys which are used both in alloy transistors proper and also in the Texas Instruments diffusion power transistor described by Dr. Kendall, we ourselves do not in fact find any mechanical instability with temperature of our silicon alloy transistors as compared with similar dimension germanium alloy transistors. The OC 200 and OC 201 have been temperature cycled over a considerable range with no adverse effects: it is therefore not then correct to suggest that lack of temperature stability is a drawback to their use.

The Mullard Radio Valve Co., Ltd. M. SMOLLETT
Southampton, Hampshire.
10th July 1958.

New Books

Passive Network Synthesis

By JAMES E. STORER, Ph.D. Pp. 319. McGraw-Hill Book Co. Ltd., 95 Farringdon St., London, E.C.4. Price 64s.

The development of techniques for the synthesis of networks has accelerated during recent years; but to those not actively engaged on advanced network design problems, comprehension of the salient principles and significant approaches has been impeded by the complexity and diversity of published papers during this evolutionary period.

Publication during the past year or two of a number of text-books devoted to synthesis occurs at a time that is opportune for a perspective view of the subject, and these books are therefore very welcome.

Dr. Storer's book is distinctive. Among the other works in mind, it is the smallest; yet through a concise style, it succeeds in conveying and illustrating at least the principles of most approaches now accepted as important. To the extent that it is a relatively small book, its primary function is perhaps that of a concise survey; but this alone represents a valuable contribution to the vital process of sifting and consolidating existing knowledge.

The book comprises 31 short chapters, distributed in four parts entitled: Impedance Synthesis; Network Synthesis using Image Parameters; Modern Realisation Methods for Two-terminal-pair Networks; and Rational Fraction Approximations. Following an introductory chapter in which the problems and criteria of analysis and synthesis are clearly compared, the conditions for realizability of impedance functions, fundamental to all synthesis procedures, are briefly explained. Foster's reactance theorem, its extensions and Cauer networks, are then developed over several chapters. Amongst these is one of special interest for its concern with simplification of procedure by means of "The Foster Preamble". Other approaches to the synthesis of impedances considered in this section of the book are those of Brune, Darlington, and Bott-Duffin.

In part two, four chapters are devoted to an outline of image parameter theory, *m*-derived filters, and frequency transformations. Notable features are the inclusion of extensive closely-graded attenuation tables for *m*-derived filters, and the lucid treatment of frequency transformations. Further chapters relate to *m*-derived and maximally-flat delay lines, and the lattice network, with particular reference to equivalences and constant-resistance networks.

Part three embraces discussion of four-terminal network types, the forms in which external parameters may be specified, and realizability. The synthesis procedures that follow cover Darlington's transfer-function approach to four-terminal reactance networks; the lattice structure, with special interest to be derived from its manipulation into unbalanced forms, and exploitation of the well-known theorems relating to impedances common to the four arms of a lattice; LC ladder networks by the Cauer-Guillemin method; RC networks in general, and Guillemin's important voltage-transfer-function approach in particular; the use of active networks (gyrators), and the Scott-Blanchard iterative process, which has the appeal of easy comprehension as well as practicality.

In the fourth and final section of the book, an introductory discussion of the approximation problem is followed by some specific methods, including noteworthy attention to the important distribution analogies of poles, zeros and attenuation in the electric circuit, with lines of opposite charge and potential in the electric field. The closing chapter is devoted to an outline of time-domain synthesis.

Knowledge of fairly advanced mathematics, particularly complex variable theory that is inseparable from present synthesis approaches, is assumed; but understanding is enhanced by good numerical illustrations in the text. It is a pity, however, that the problems terminating each chapter, in conformity with a bad American practice, have no answers. Occasionally the exposition is unbalanced: approximation methods, for example, seem to warrant greater weight.

The author's interpretation of duality is thought to illustrate a current looseness of definition. While a network that is inverse to another must be its dual, a pair of networks that are duals are not necessarily inverse: duality does not demand a quantitative relationship, but a parallelism. For this reason, the term 'duality constant', used in relation to quantitatively inverse networks on p. 130, is considered undesirable.

The book is highly recommended, particularly to those entering upon a study of synthesis and desiring an initial comprehensive view of its salient mathematical approaches. The inclusion of a chapter on so modern a topic as synthesis with active networks (gyrators) is a further measure of its value.

F. E. R.

Electronic Semiconductors

By E. SPENKE. Pp. 402. McGraw-Hill Publishing Co. Ltd., 95 Farringdon St, London, E.C.4. Price 85s. 6d.

The English translation of Spenke's "Elektronische Halbleiter" has made a welcome appearance and will serve to propagate this work even more widely. The book stands among the best that have been written as an introduction to semiconductor physics and it fulfils its aim of bridging the gap between the elementary and advanced texts.

The initial chapters introduce a simple quantum mechanical picture of conduction processes in semiconductors, together with a description of imperfections in crystals, their interactions and the mass action laws governing them. The concept of the 'hole' is elaborated, together with some of the approximations involved in its use. The ideas developed in the first part are then used to elucidate the mode of action of transistors and diodes. One might make some objection to the continued use of the term "Type 'A'" transistor which is an historical expression to which no real significance now attaches. The last part of the book is devoted to a more thorough treatment of the quantum mechanics and statistics of semiconductors, together with interesting chapters on lifetime and boundary layers.

The translation is generally competent but in a few places German expressions have been translated too literally; for example, 'poisoning' with its pejorative associations is used instead of compensation doping or mixed conduction. This book can be wholeheartedly recommended to the beginner and also to the specialist for its interesting and lucid exposition of semiconductor physics.

M.S.

Television in Science and Industry

By V. K. ZWORYKIN, E. G. RAMBERG and L. E. FLORY. Pp. 300. John Wiley & Sons Inc., available in the U.K. from Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 80s.

The authors, who believe that "The possibility of translating vision to distant and inaccessible points by electrical means can . . . make a material contribution to all phases of human activity", have written a book showing how closed-circuit television can be used for purposes other than entertainment. They begin with a short history of the evolution of television, followed by a review of fields of application of closed-circuit systems, a long description of actual apparatus, an account of progress to date, and a forecast of future applications. Mathematics are excluded except for an appendix in which flying-spot and 'Vidicon' television microscopes using ultra-violet radiation are compared. The book will be of interest to those who wish to use closed-circuit systems. G.S.

Feedback Control Systems

By Otto J. M. Smith. Pp. 694 + xviii. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.1. Price £5 1s. 6d.

This book is intended for the reader who has already some knowledge of feedback control systems and who is familiar with Fourier and Laplace transforms. The book is divided into four sections. Parts I and II are Linear Analysis and Linear Synthesis respectively, Parts III and IV are Non-Linear Analysis and Non-Linear Synthesis respectively.

The first two chapters scan the various methods of linear analysis. They indicate the use of the Laplace complex-frequency S plane if the original data is in analytic form and the use of the L plane, (i.e., a plot of log-gain versus phase) if the original data is in measured form.

From chapter three onwards great stress is laid on design in the S plane including root-locus methods. On page 1 of this 694-page book there occurs the rather uncompromising statement, "Throughout this book, it will be assumed that the reader has and uses a Spirule or a potential-plane analogue". The prospective reader need not however be dismayed as one can construct root-locus plots using a disc of tracing paper and a pair of compasses.

The following chapters cover the analysis and synthesis of multi-loop systems, systems with random signals and systems with distributed parameters.

The latter half of the book is devoted to a comprehensive treatment of non-linear systems including relay systems and multilevel-decision systems. Common non-linearities treated are dead-zone, saturation, hysteresis, spring reload, mechanical stop and friction effects. Predictor control of saturating systems is covered in some detail. The book ends with a chapter on carrier systems.

This book gives a clear and comprehensive account of the modern analytical and design techniques which are rapidly becoming indispensable to the control systems engineer. The treatment is theoretical in character and thus the design of system components is not covered. An Appendix gives most of the compensation networks but only two outline circuits appear in the figures.

There is one slightly surprising omission from the contents; no information is given on the derivation of analytical transfer functions from measured data. A.T.M.

Bibliografia Marconiana

By Giovanni Di Benedetto. Pp. 243. Consiglio Nazionale delle Ricerche, Piazzale delle Scienze No. 7, Roma, Italy. Price L 2000.

The main part of this work consists of two bibliographies, the first of Marconi's own works and the second of writings about Marconi. There is also a useful chronological table detailing the principal events in Marconi's life and a list of honours conferred upon him; a name index is also included.

Research for Industry. 1956-57

Pp. 159. Published for D.S.I.R. by Her Majesty's Stationery Office, Kingsway, London, W.C.2. Price 7s. 6d.

A summary of work done by industrial research associations aided by the British Government.

Leitfaden zur Berechnung von Schallvorgängen

By Dr. H. Stenzel. Second edition revised by Dr. O. Brosze. Pp. 168. Springer-Verlag, Reichpietschufer 20, Berlin W. 35, Germany. Price D.M. 31.50.

Pulse and Time-Base Generators

By D. A. Levell, M.Sc., A.M.I.E.E. Pp. 175. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 25s.

Deals mainly with valve circuits; there is a short final chapter on transistors. A preliminary analysis of linear circuits is followed by chapters on non-linear circuits (largely non-mathematical), pulse modulators, trigger circuits and electromagnetic time-bases.

Faisceaux Hertziens et Systèmes de Modulation

By L. J. Libois. Pp. 509. Editions Chiron, 40 rue de Seine, Paris 6e, France.

The major part of the book is devoted to a discussion of the various modulation systems suitable for use in microwave links for multi-channel telephony and for television.

Les Ondes Centimétriques

By G. Raoult. Pp. 401. Masson et Cie, 120 Boulevard Saint-Germain, Paris 6e, France. Price Fr. 7300.

Textbook on microwaves for degree students.

I.E.E. COUNCIL

The constitution of the I.E.E. Council for the next Session is: S. E. Goodall, M.Sc.(Eng.), F.Q.M.C.; The Past Presidents; Vice-Presidents: Sir Willis Jackson, D.Sc., D.Phil., Dr.Sc.Tech., F.R.S., G. S. C. Lucas, O.B.E., Sir Hamish D. MacLaren, K.B.E., C.B., D.F.C., LL.D., B.Sc., C. T. Melling, C.B.E., M.Sc.Tech., A. H. Mumford, O.B.E., B.Sc.(Eng.); Honorary Treasurer: E. Leete, with Chairmen and Past-Chairmen of Section and Local Centres.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for June* 1958

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

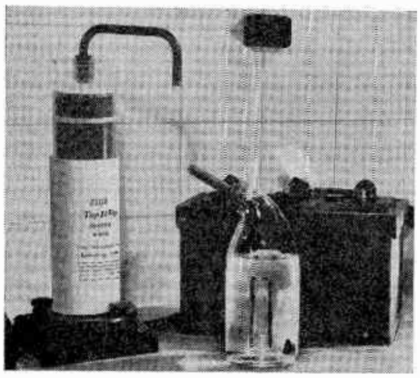
* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

New Products

Distilled Water on Tap

The Elga Top-It-Up is said to provide distilled water instantly, at tap speed, for battery service.

Ordinary tap water is fed in through the base and passes through a column of ion



exchange resins. These remove impurities and pure distilled water flows from the outlet pipe (see photograph).

The cost of producing distilled water by deionisation is claimed to be much lower than the cost in carboys.

Deionisation (Elga) Ltd.,
*Railway Place, London, S.W.*19.

Radar Tubes

Two new radar tubes are announced by G.E.C.

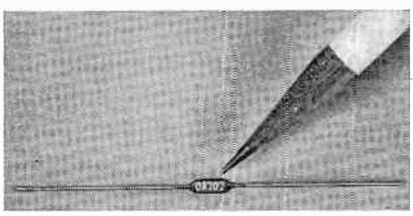
Type 1252 WTM is a 5-in. high-definition tube intended for airborne radar uses where equipment weight is important. It uses low-voltage electrostatic focusing and a bleeder chain can be used to supply intermediate electrode voltages. The tube has a yellow-green aluminized long-persistence screen.

2269 YMM is a high-precision 9-in. flat-faced radar tube with an aluminized orange long-persistence screen, designed for airborne and other light-weight radar. The spot size under normal working conditions is said to be about ¼ mm and the glass-work is manufactured to close tolerances, allowing service replacements without disturbance to surrounding components.

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

Semiconductor Devices

A number of silicon and germanium devices have been announced by Mullard. These include a silicon rectifier, silicon zener diodes, silicon transistors, a gold-bonded germanium diode and sub-miniature germanium transistors. Some details from the makers' advance specifications are given below.



Silicon Junction Diode OA202 (Illustrated)

Maximum inverse voltage (peak or d.c.)	150 V
Maximum forward current:	
Peak	100 mA
Mean	30 mA
Forward voltage drop at 30 mA:	
at 25°C	0·9 V
at 100°C	0·8 V
Reverse Current at −150 V:	
at 25°C	0·05 μA
at 100°C	5 μA
Maximum ambient temperature	125°C

Silicon Zener Diodes OAZ200 to 207

These have nominal voltages of 4·7, 5·1, 5·6, 6·2, 6·8, 7·5, 8·2 and 9·1 V, and the maximum rated reverse current is 40 mA. For the 6·2-V unit (OAZ203) typical slope resistances are 160 Ω at 1 mA and 6 Ω at 5 mA, and a typical change in zener voltage with temperature is 2·4 mV/°C. The maximum permissible junction temperature is 150°C.

Silicon Transistors OC200, OC201

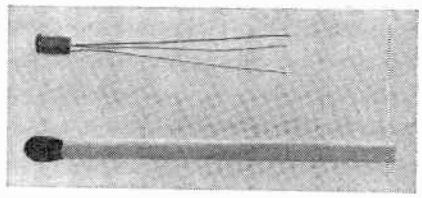
These are low-power, medium-frequency p-n-p alloyed types. Some data were given in the June issue, p. 206 These have low collector bottoming voltages (typically about 100–130 mV) and low noise figures (6 dB at 1 kc/s for the OC201).

Gold-Bonded Germanium Diode OA5

Maximum inverse voltage:	
at 25°C	100 V
at 75°C	50 V
Maximum forward current:	
Peak	350 mA
Mean—at 25°C	115 mA
at 75°C	35 mA
Pulse (1 μsec)	1 A
Average forward voltage drop at 300 mA, 25°C	0·9 V
Average reverse current at −50 V: at 25°C	2·5 μA
at 60°C	32 μA

Sub-Miniature Germanium Transistors OC57, 58 and 59

These are p-n-p types of very small physical size (see photograph). Typical common-emitter current gains for the three types respectively are 50, 70 and 100. The common-emitter collector leakage current (I'_{co}) is 100 μA maximum at 25°C for all



types, the minimum cut-off frequency in the common-emitter circuit is 10 kc/s, and the maximum noise figure 10 dB. The maximum rated collector dissipation at temperatures up to 40°C is 10 mW.

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

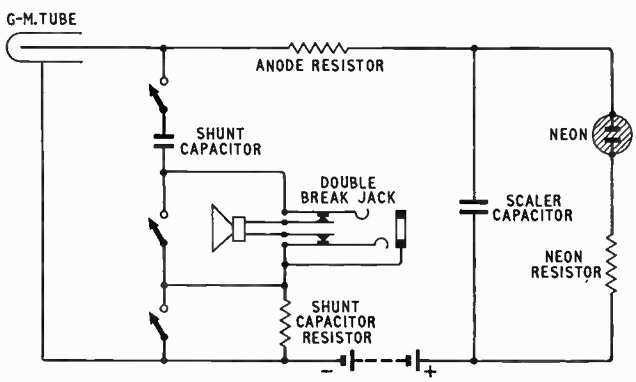
Inexpensive Geiger Counter

The Utility Geiger Counter is described as a self-contained, light-weight portable instrument employing a sensitive low-voltage halogen-quenched Geiger tube and counter providing three ranges of visual response by flashing neon, together with a loudspeaker count of each single ionizing event at low radiation levels.

It is manufactured under licence from the United Kingdom Atomic Energy Authority and employs a patented circuit evolved at the British Atomic Weapons Research Establishment (see diagram).

Typical background counts over chalk areas are said to be 40–50 per minute while, over West of England granite, away from metalliferous areas, counts range from 118–180 per minute, according to locality.

The high-tension power is derived from a battery pack of internationally-available 30-V batteries. These are assembled in a plastic case which is moulded from high-impact polystyrene. The current consumption under normal conditions is so low



(a millionth or so of an amp) that it is unnecessary to provide an on-off switch.

Thumb-operated controls permit single-handed operation on the medium and low ranges, but an extra long shoulder strap is provided for convenience when field-surveying or prospecting for thorium or uranium ores.

A pulse output socket provides for muting the internal speaker and substituting a miniature crystal earpiece, or for attachment to an oscilloscope or recording scaler.

Specification: Size 9 in. by 6 in. by 3 in.; weight, 3 lb.; background count, 90/min.; operating potential, 400–450 V; current consumption, 0·001 to 10 μA; response exceeds 30,000 counts/min.; output pulses approximately 1, 15 and 50 V. Tube, effective length 11·8 cm; plateau, minimum length 80 V, slope 5% per 100 V; dead time, 100 μsec at 6,000 c/min.; rise time, 5–10 μsec at 6,000 c/min.; gamma efficiency, 1%; temperature range, −55°C to +60°C; stainless-iron electrode.

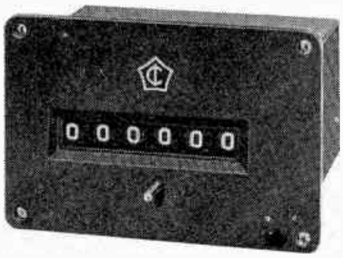
Radiation Monitors Ltd.,
52 Tottenham Court Road, London, W.1.

High-Speed Counter with Quick Re-Set

This 6-figure unit (Series 100) is described as a low-priced accurate instrument with a very high counting rate, low consumption and a very long mechanical life, with provision for instantaneous re-set. The rated counting speed is 20 per second.

The number wheels and associated pinions are made of a low-friction nylon and the assembly is actuated by a balancing escapement system set in jewelled bearings. This movement is operated by two small electro-magnets spaced around a common armature.

Instantaneous re-set is obtained by the deflection of a short lever and facilities are provided for electrical re-set.

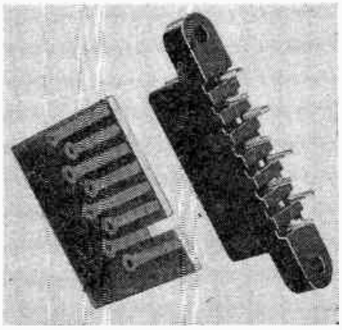


The counter is available for either base (proud) or flush-panel mounting and versions are available for a.c. mains or d.c. low-voltage supplies.

Counting Instruments Ltd.,
5 Elstree Way, Boreham Wood, Herts.

Printed Circuit Components

Recently-introduced printed-circuit connectors include 8, 12 and 18-pole units based on a 0·15-in. module and an 8-pole unit (illustrated) based on a 0·1-in. module. Points from the makers' specification for the latter unit are given below:



- Contact resistance, $4·7 \times 10^{-3}\,\Omega$ (typically $5–15 \times 10^{-3}\,\Omega$ after 3,000 insertions).
- Current rating for < 30°C temperature rise, 3–4 A.
- Breakdown voltage between pins 2·5 kV (at sea level); 450 V (at 68,000 ft).

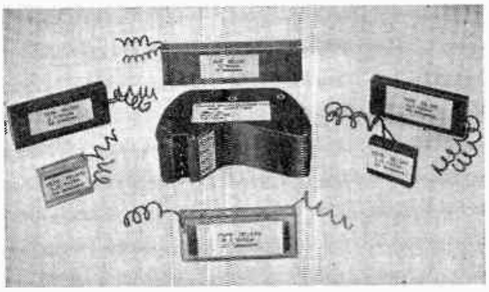
Belling & Lee Ltd.,
Great Cambridge Road, Enfield, Middx.

Potted Silver-Zinc Accumulators

A new cell type HT125 has been produced to meet the need for a small light rechargeable battery.

Strips of from 1 to 10 cells are arranged in any desired form before potting in resin. The user can therefore be offered a battery which is hermetically sealed and which will fit into the odd space which so often is all that is available for the power supply.

The HT125 cell has a nominal capacity of 0·125 ampere hour, being suitable for discharge over a range of currents from 10 to 500 mA. A single cell weighs ⅓ oz. and measures 1⅛ in. × 1¼ in. × 9/32 in., but the size-to-weight ratio improves as the number of cells in a battery increases, since the thickness of potting resin remains sensibly constant. A 15-V nominal unit weighs 2⅓ oz. and measures 4¾ in. × 1¼ in. × 7/16 in.

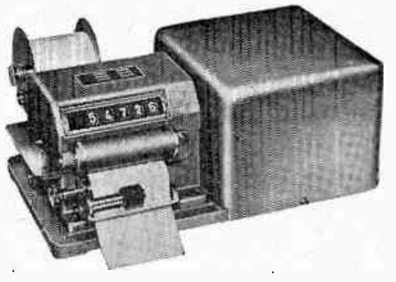


Venner Accumulators Ltd.
Kingston By-Pass, New Malden, Surrey.

Electrical Printing Counter

A new electro-mechanical printing counter is available in various models to suit different applications. These include the counting of electrical impulses or closures of a switch up to a speed of 5 per second, the recording of elapsed time in units of 0·01 minute or in seconds and the counting of revolutions of a shaft up to 50 per second. The count is printed in numbers of up to six digits on a paper roll 2⅜ inches wide or on cards.

An automatically-changed serial number or hand-adjusted code, consisting of two figures or letters can be printed in addition to the counting digits. All characters of the code and digits are printed simultaneously.



The printing counter is described as a compact instrument occupying a volume less than 11 inches in length, 6 in. wide and 4½ in. high, weighing approximately 11 lb, and is suitable for building into automatic control panels and measuring equipment. It can be supplied with fully-automatic action or with a hand-operated print lever; the printing, paper shifting and re-setting operations being performed automatically in a total time of one second by means of a torque motor energized by an electrical impulse.

Radiatron,
7 Sheen Park, Richmond, Surrey.

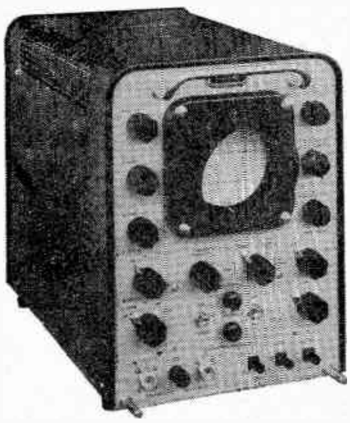
Oscilloscope

This new Solartron oscilloscope, type CD 814, has Y amplifiers with a constant bandwidth of 0·9 c/s–8 Mc/s (−3 dB) at all gain settings. The rise time is given as 40 mμsec, and the maximum sensitivity as 40 mV/cm. Sensitivity is variable in steps by means of an input attenuator and continuously over a 10 to 1 range by means of a gain control. A 50-c/s square-wave calibrating waveform is available at voltages of 100 mV, 1 V, 10 V and 100 V, all ± 5%. The cathode-ray tube is a 4EP1 with a 3½-inch screen and a post-deflection accelerator.

The sweep repetition rate is quoted as 10 c/s–185 kc/s, giving sweep speeds of 0·2 μsec/cm to 0·01 sec/cm, and trace expansion of × 10 is provided. There is an internal television sync separator. The X-amplifier response is 1·5 c/s–1 Mc/s (−3 dB) and gives an X sensitivity of 500 mV/cm.

Time markers of 0·1, 1, 10 and 100 μsec are provided.

Solartron Electronic Group Ltd.,
Thames Ditton, Surrey.



Abstracts and References

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

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

ACOUSTICS AND AUDIO FREQUENCIES

534.1.087 : 621.395.616 **2276**

Absolute Calibration of a Capacitance-Type Vibration Pickup.—(*Tech. News Bull. nat. Bur. Stand.*, Jan. 1958, Vol. 42, No. 1, pp. 1–3.) See also 332 of February (Koidan).

534.143 : 537.528 **2277**

The Electroacoustic Efficiency of a Spark Discharge in Water.—N. A. Roï & D. P. Frolov. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Feb. 1958, Vol. 118, No. 4, pp. 683–686.)

534.232 : 537.228.1 **2278**

Transients in Piezovibrators.—P. V. Ponomarev. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 243–253.) An investigation of the operation of a radiating quartz plate excited by pulses of arbitrary shape.

534.232 : 537.228.1 **2279**

Shift of Resonant Frequencies of a Plane Piezoelectric Radiator with Resistive Load.—A. A. Anan'eva. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 282–285.) Investigation of $BaTiO_3$ ceramics and other piezoelectric materials showed the considerable influence of a resistive load on measurements of resonance frequencies. Results are presented graphically.

534.232-8 **2280**

New Technique for Measuring Transducer Blocked Impedance.—G. A. Sabin. (*J. acoust. Soc. Amer.*, Feb. 1958, Vol. 30, No. 2, pp. 146–150.) A pulse method for measurements on underwater transducers is described and results for a magnetostrictive transducer resonant at 60 kc/s are given.

534.232-8 : 621.318.134 **2281**

Testing Experimental Ferrite Ultrasonic Receivers.—I. P. Golyamina, A. D. Sokolov & V. I. Chulkova. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 288–290.) The acoustic properties of some Ni-Zn ferrites used in ultrasonic receivers are reported and the frequency response of Ni ferrite and Ni-Zn ferrites are shown graphically.

534.232-8 : 621.395.623.52 **2282**

Theory of Ultrasonic Concentrators.—L. G. Merkulov. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 230–238.) The properties of conical, exponential and catenary horns are investigated and their characteristics are calculated. Experiments show that the highest magnification is obtained with the catenary type of horn.

534.241 **2283**

Frequency Dependence of Echoes from Bodies of Different Shapes.—R. Hickling. (*J. acoust. Soc. Amer.*, Feb. 1958, Vol. 30, No. 2, pp. 137–139.) It is shown by calculation that the frequency dependence of an echo is different for a prolate spheroid, an infinite cylinder and a sphere.

534.6 : 519.272 **2284**

Correlation Method of Measurement of Acoustic Relations.—S. G. Gershman & E. F. Orlov. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 285–288.) Results of experiments carried out in a building of the Acoustical Institute of the U.S.S.R. are shown graphically and a method for the calculation of transmission losses is described. See also 2501 of 1955 (Goff).

534.76 : 534.78 **2285**

Stereophonic Listening and Speech Intelligibility against Voice Babble.—I. Pollack & J. M. Pickett. (*J. acoust. Soc. Amer.*, Feb. 1958, Vol. 30, No. 2, pp. 131–133.) Experiments with earphones confirm that directional information may improve intelligibility.

534.78 **2286**

Masking of Speech by Noise at High Sound Levels.—I. Pollack & J. M. Pickett. (*J. acoust. Soc. Amer.*, Feb. 1958, Vol. 30, No. 2, pp. 127–130.) Over a wide range of conditions a high level of background noise was found to cause deterioration of speech intelligibility for a constant speech/noise ratio.

534.861 **2287**

Experimental Study of Statistical Properties of Musical and Speech Signals in Broadcasting.—B. A. Fersman. (*Akust. Zh.*, July 1957, Vol. 3, No. 3, pp. 274–281.) An experimental installation for the investigation of the distribution function of instantaneous values of four types of signal is examined. Speech signals have been found to follow a Cauchy type of distribution.

621.395.625.3 : 621.317.42 **2288**

The Determination of the Magnetization of Magnetic Recording Tape.—Schmidbauer. (See 2505.)

AERIALS AND TRANSMISSION LINES

621.315.212 **2289**

Attenuation in Continuously Loaded Coaxial Cables.—G. Raisbeck. (*Bell Syst. tech. J.*, March 1958, Vol. 37, No. 2, pp. 361–374.) A theoretical study of the attenuation in cables loaded with magnetic material is presented. Three parameters are involved, and when the line has optimum dimensions, two parameters determine the attenuation. Graphical and analytical relations are given in forms which can be applied to practical problems.

621.372.2 **2290**

Experimental Check of Formulas for Capacitance of Shielded Balanced-Pair Transmission Line.—B. G. King, J. McKenna & G. Raisbeck. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 922–923.) See 1646 of 1957 (Gent).

621.372.43 **2291**

General Synthesis of Quarter-Wave Impedance Transformers.—H. J. Riblet. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 36–43. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.) See also 1250 of 1955 (Collin).

621.372.8 : 537.226 **2292**

Losses in Dielectric Image Lines.—D. D. King & S. P. Schlesinger. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 31–35. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.)

621.372.8 : 621.372.413 **2293**

Resonance Properties of Ring Circuits.—F. J. Tischer. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 51–56. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.)

621.372.821 : 621.372.832.6 **2294**

Strip-Line Hybrid Junction.—H. G. Pascalar. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 23–30. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.)

621.372.825 **2295**

Calculation of Parameters of Ridge Waveguides.—Tsung-Shan Chen. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 12–17. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 575.)

621.372.85+621.315.212] : 621.318.134 **2296**

Ferrite Components in Microwave Systems.—B. L. Humphreys. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 341–345.) A review of circuit techniques.

621.396.67 : 621.397.62 : 621.317.3 **2297**

Measuring TV Aerial Performance.—F. R. W. Strafford. (*Wireless World*, Feb., March & June 1958, Vol. 64, Nos. 2, 3 & 6, pp. 67–69, 120–123 & 294–298.) A minimum height of 25 ft and distance/height ratio of 15 are recommended for the test transmitting and receiving aerials; conditions for accurate measurements and the selection of a suitable test site are discussed. The choice of standard aerials and of radiating and detecting equipment is considered, and methods of presenting gain and directional response characteristics are given. Techniques of aerial impedance measurement and effects of feeder mismatching are discussed.

621.396.67.095 **2298**

Use of a Fictitious Magnetic Current for the Solution of the Problem of Radiation of an Aerial over a Surface with the Heterogeneous Boundary Conditions of Leontovich.—O. N. Tereshin. (*Radiotekhnika, Mosk.*, April 1957, Vol. 12, No. 4, pp. 24–31.) The conditions are defined in terms of a coefficient α, the ratio of the tangential components of the electric and magnetic fields.

621.396.677.43 **2299**

The Behaviour of Two Concentric Rhombic Aerials.—W. Kronjäger & K. Vogt. (*Nachrichtentech. Z.*, Oct. 1957, Vol. 10, No. 10, pp. 494–496.) Measurements of coupling factor, matching and gain of two rhombics, one of which is placed inside the other, show that the influence of the loaded outer aerial on the inner is negligible. The installation discussed is used for reception in a 4 : 1 range of frequencies, each aerial covering a range of 2 : 1.

621.396.677.5 **2300**

Super Loop Antenna.—S. Adachi, J. R. McDougal & Y. Mushiake. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B.*, June 1957, Vol. 9, No. 1, pp. 1–8.) Measurements of input impedance and directional properties at 120–370 Mc/s are presented for a multi-element aerial consisting of circular loops, positioned concentrically around a supporting pipe of relatively large diameter and connected in parallel at their feed points.

621.396.677.5 **2301**

Theoretical Formulation for Circular Loop Antennas by Integral Equation Method.—S. Adachi & Y. Mushiake. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, June 1957, Vol. 9, No. 1, pp. 9–18.) An integral equation for the current distribution in a loaded circular loop aerial is solved by the successive approximation method, thus leading to expressions for the transmitting and receiving characteristics of a loaded or short-circuited loop aerial at h.f.

621.396.677.71 **2302**

Radiation and Surface Currents from a Slot on an Infinite Conducting Cylinder.—F. H. Northover. (*Canad. J. Phys.*, Feb. 1958, Vol. 36, No. 2, pp. 206–217.) The saddle-point method of calculating the far field, as used by Papas (2715 of 1949) and others, is not valid for points close to the aerial. The basis for a method which overcomes this defect is described. The ranges of validity of this method and of the Papas procedure are discussed.

621.396.677.73 **2303**

A New Broad-Band Microwave Antenna System.—R. W. Friis & A. S. May. (*Commun. & Electronics*, March 1958, No. 35, pp. 97–100.) The antenna is an electromagnetic horn capped by a sector of a paraboloidal reflector. It is intended for simultaneous operation in the bands 3 700–4 200 Mc/s, 5 925–6 425 Mc/s, and 10 700–11 700 Mc/s. Details of gain and directivity are given.

621.396.677.85 : 629.19 **2304**

Loaded-Lens Antenna tracks Missiles.—L. S. Miller. (*Electronics*, 28th March 1958, Vol. 31, No. 13, pp. 44–46.) An artificial dielectric lens consisting of concentric foam-plastic hemispheres each covered with metal disks is used to provide a circularly polarized feed for a 60-ft paraboloid at 216–245 Mc/s. A waveguide feed allows independent control of E- and H-plane patterns, and separate horizontal- and vertical-polarization outputs provide maximum polarization versatility.

AUTOMATIC COMPUTERS

681.142 **2305**

Simulation of Nonlinear Field Problems.—I. C. Hutcheon. (*Brit. Commun. Electronics*, Feb. 1958, Vol. 5, No. 2, pp. 96–99.) Some three-dimensional problems can be solved by using a resistance network in conjunction with a single a.c. amplifier, a function generator, a scanning switch and a number of simple analogue storage units.

681.142 : 512.831 **2306**

Matrix Analysis of Logical Networks.—E. J. Schubert. (*Commun. & Electronics*, March 1958, No. 35, pp. 10–13.) "A novel method of matrix algebra is derived for the analysis of networks representing logical systems too complex for conventional approaches using Boolean algebra. It is shown that truth tables of system blocks may be treated as matrices and that logical operations on such matrices can be performed efficiently."

681.142 : 517.512.2 **2307**

An Analogue Computer for Fourier Transforms.—D. G. Tucker. (*J. Brit. Instn Radio Engrs*, April 1958, Vol. 18, No. 4, pp. 233–235.)

681.142 : 621.311.25 : 621.039 **2308**

Analogue Computers in the Nuclear Power Programme.—J. C. Nutter. (*Instrum. Practice*, Jan. 1958, Vol. 12, No. 1, pp. 46–51.)

681.142 : [621.314.63 + 537.525.92 **2309**

New Applications of Impedance Networks as Analogue Computers for Electronic Space-Charge and for Semiconductor Diffusion Problems.—G. Čremošnik, A. Frei & M. J. O. Strutt. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 868–877.) The general second-order partial differential equation is transformed to the equivalent equation of finite differences and hence relations for analogue impedance networks are derived. Resistance chains are applied to the case of one-dimensional high-vacuum diodes yielding results within about 1% of the exact solutions. *RC* networks are applied to semiconductor diffusion with space and surface recombination, giving some new results for *p-n* diodes. A plane resistance network applied to a triode with space charge gives the anode-current/grid-voltage curve within a few per cent of the published value. See also 1493 of May (Čremošnik & Strutt) and back references.

681.142 : 621.314.7 **2310**
A Basic Transistor Circuit for the Construction of Digital Computing Systems.—P. L. Cloot. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 213–220.) A circuit, comprising one transistor, one capacitor and three resistors, forms the basis of a complete digital computing system for use where high operation speed is not essential. The construction of well known computer circuits from the basic circuit is described and a complete logical computing system, using 184 circuits, is developed to demonstrate their application.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 **2311**
Printed Wiring.—H. Nagatsu & H. Sakai. (*Rep. elect. Commun. Lab., Japan*, Dec. 1957, Vol. 5, No. 12, pp. 18–29.) An assessment and detailed description of processes in the production of photo-etched circuits. The properties of the copper foil used are examined and particular attention is paid to the photographic and enamelling techniques and to dip-soldering.

621.314.22 **2312**
Small Transformers with Low Magnetic Leakage.—A. Lang. (*Elektrotech. Z., Edn A*, 1st Jan. 1958, Vol. 79, No. 1, pp. 10–14.) The external field characteristics of various types of laminated and C-core transformers are examined to determine the form of construction with lowest leakage.

621.314.22 : 621.372.512.2 **2313**
Wide-Band Transformer Characteristics.—A. C. Hudson. (*Electronic Radio Engr*, June 1958, Vol. 35, No. 6, pp. 228–234.) "A parameter f_r is defined for wide-band transformers, which represents the series resonance of the leakage inductance and the primary and secondary stray capacitances. It is shown that this parameter may be determined from the low-frequency requirements on the transformer, and then used as a guide to the attainable high-frequency response."

621.318.042.1 : 538.24 **2314**
Terminal Properties of Magnetic Cores.—T. C. Chen & A. Papoulis. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 839–849.) The terminal properties of the cores can be obtained from their dynamic step response; it is shown that for thick (2 mils or more), thin (1/8 mil) and intermediate-sized cores there is a unique relation between the flux in the core and the applied coulomb-turns. The analysis is verified by experiment.

621.318.44 **2315**
New Techniques for Winding Subminiature Coils.—W. F. Kallensee. (*Electronic Ind.*, Jan. 1958, Vol. 17, No. 1, pp. 70–71.) Details of a method for winding self-supporting coils of adhesive-coated wire.

621.318.57 : 621.314.63 **2316**
Solid-State Thyratron switches Kilowatts.—R. P. Frenzel & F. W. Gutzwiller. (*Electronics*, 28th March 1958, Vol. 31, No. 13, pp. 52–55.) The operation of a controlled silicon rectifier and its application in various switching circuits and converters are described.

621.318.57 : 621.396.96 **2317**
Broad-Band Balanced Duplexers.—C. W. Jones. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 4–12. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 575.)

621.319.4 **2318**
Research of Condensers and its Results.—S. Hayashi, K. Kinugawa, K. Kudo, M. Chiba & S. Sakurai. (*Rep. elect. Commun. Lab., Japan*, Dec. 1957, Vol. 5, No. 12, pp. 30–34.) Voltage, temperature and frequency characteristics are given for a number of types of capacitor, particularly the $BaTiO_3$ and silvered mica types.

621.319.4.001.4 **2319**
Accelerated Life Testing of Capacitors.—G. J. Levenbach. (*Trans. Inst. Radio Engrs*, June 1957, No. PGRQC-10, pp. 9–20. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1957, Vol. 45, No. 8, p. 1166.)

621.319.43 **2320**
Precision Variable Capacitors for High-Grade Electronic Equipment.—A. A. Turnbull. (*Brit. Commun. Electronics*, Dec. 1957, Vol. 4, No. 12, pp. 756–759.) Two types of small variable capacitor are described which were designed to meet stability and accuracy specifications comparable with larger types. One gives a linear frequency variation over 150° of shaft rotation; the other gives a sinusoidal variation of frequency about a mean value as the rotor shaft is turned continuously.

621.372.413 **2321**
Excitation of Higher-Order Modes in Spherical Cavities.—R. N. Ghose. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 18–22. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, pp. 575–576.)

621.372.44 : 621.372.6 **2322**
Some General Properties of Nonlinear Elements: Part 2—Small-Signal Theory.—H. E. Rowe. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 850–860.) The analysis of nonlinear capacitor modulators, demodulators and negative-conductance amplifiers, with only two signal frequencies present, gives results that agree with the general energy relations of Part 1 [2988 of 1956 (Manley & Rowe)]. In addition, the gain, bandwidth, terminal admittances, and sensitivity to changes in terminal admittances or in local-oscillator drive, are given together with their dependence upon the amount of nonlinearity. Generally the bandwidth of these devices approaches zero as the nonlinearity approaches zero.

621.372.44 : 621.372.632 **2323**
General Properties of Frequency-Converting Networks. — S. Duinker. (*Philips Res. Rep.*, Feb. & April 1958, Vol. 13, Nos. 1 & 2, pp. 37–78 & 101–148.) A generalized analysis of frequency-converting systems consisting of networks containing nonlinear inductors, capacitors and resistors, and including also nonlinear *L*, *C* and *R* coupling elements. The mesh and nodal equations for a general nonlinear network are given, and it is assumed that a certain high-voltage and current distribution (the so-called fundamental state) results from the presence of (carrier) current and voltage sources. First-order perturbational equations are then derived corresponding to the disturbance of the fundamental state by small (signal) voltage or current sources. Linear differential equations with time-dependent coefficients are obtained, which yield, for a periodic fundamental state, equations in matrix form. Methods of simplifying the matrices are shown, and the method is used to examine stability, conversion gain and loss. As an example, a polyphase magnetic modulator is analysed.

621.372.5 **2324**
Parallel Four-Terminal Networks.—F. E. Rogers. (*Electronic Radio Engr*, June 1958, Vol. 35, No. 6, pp. 207–211.) A simple relation exists between the output/input voltage ratio for any number of four-terminal networks in parallel and their parameters of output admittance and short-circuit transfer admittance. The relation enables the voltage ratio to be stated without solving the networks.

621.372.512.3 : 621.374.4 **2325**
A Wide-Band Multiplier Unit.—G. T. Sassoon. (*R.S.G.B. Bull.*, Feb. 1958, Vol. 33, No. 8, pp. 360–362.) Capacitive coupling leads to easily adjustable wide-band couplers. A practical design of a wide-band frequency-multiplier unit incorporating these and covering all bands from 3·5 to 28 Mc/s is given.

621.372.54 : 621.372.412 **2326**
Ceramic I.F. Filters match Transistors.—D. Elders & E. Gikow. (*Electronics*, 25th April 1958, Vol. 31, No. 17, pp. 59–61.) $BaTiO_3$ resonant filters can replace i.f. transformers with improvement in size, cost, ruggedness and insertion loss. Their input and output impedances are compatible with those of transistors.

621.372.543 **2327**
Transient Phenomena in Band Filters.—A. Sabbatini. (*Note Recensioni Notiz.*, Sept./Oct. 1957, Vol. 6, No. 5, pp. 625–638.) Analysis of a filter consisting of two coupled circuits.

621.373.4 **2328**
Simultaneous Asynchronous Oscillations in Class-C Oscillators.—M. I.

Disman & W. A. Edson. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 895–903.) It is shown, theoretically and practically, that self-starting asynchronous oscillations are possible in a class-C pentode oscillator. Transient and steady-state solutions are obtained for such an oscillator with one and two degrees of freedom.

621.373.42.029.64 : 621.3.018.41(083.74) **2329**

S.H.F. Frequency Standard uses Double Conversion.—Thompson, Vetter & Waters. (See 2489.)

621.373.421.11 **2330**

Simultaneous Oscillations at Two Frequencies in a Self-Oscillatory System with Automatic Bias.—G. M. Utkin. (*Radiotekhnika, Mosk.*, April 1957, Vol. 12, No. 4, pp. 64–66.) An investigation of the self-biasing by grid current in an oscillator confirms that oscillations at two different frequencies may occur. See also 1362 of May.

621.373.421.14 : 621.3.018.41 (083.74) **2331**

An Atomic Reference Oscillator.—Capelli. (See 2490.)

621.373.421.14.029.6 **2332**

Frequency Stabilization of a Microwave Oscillator with an External Cavity.—I. Goldstein. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 57–62. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 1, p. 576.)

621.373.431.1.018.756 **2333**

Multivibrator Circuit for Millimicrosecond Pulses.—W. Gruhle. (*Elektronik*, Sept. 1957, Vol. 6, No. 9, pp. 261–263.) Pulse rise times of about 20 mμs and a repetition frequency of several Mc/s can be achieved with the valve circuit described. Pulse width can be adjusted continuously without affecting the rise time.

621.373.44 **2334**

Precision Pulse Generator.—H. L. Miller. (*Trans. Inst. Radio Engrs*, June 1956, Vol. NS-3, No. 3, pp. 18–21. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1956, Vol. 44, No. 8, p. 1084.)

621.374.32 : 621.314.7 **2335**

Dekatrons and Electromechanical Registers Operated by Transistors.—G. B. B. Chaplin & R. Williamson. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 231–236. Discussion, pp. 266–272.) Transistor circuits are described for driving cold-cathode decade tubes and electromechanical counters. In both cases the transistor circuit combines the waveform-shaping properties of the thermionic valve with the economy of power achieved with the cold-cathode valve. High reliability of performance is also attained.

621.374.4 : 621.318.4 : 621.397.6 **2336**

Magnetic-Core Dividers for ITV Sync Generators.—A. Rose. (*Electronics*, 11th April 1958, Vol. 31, No. 15, pp. 76–77.) An accurate and stable digital divider can be made with a pair of magnetic cores having rectangular hysteresis loops. Division by factors up to 17 has been obtained in a single stage in the frequency range 10–50 kc/s.

621.375.1 **2337**

On the Problem of Synthesis of Amplifier Circuits.—S. V. Samsonenko. (*Radiotekhnika, Mosk.*, April 1957, Vol. 12, No. 4, pp. 45–57.) A new mathematical method is described for the analysis of transients in amplifiers and a means of synthesis is suggested for multistage systems considering different types of distortion.

621.375.13 + 621-52 **2338**

Design of Conditionally Stable Feedback Systems.—J. Ôizumi & M. Kimura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, June 1957, Vol. 9, No. 1, pp. 19–38.) Conditionally stable systems are shown to have a greater feedback than that obtainable from unconditionally stable feedback circuits. The design of such systems is analysed mathematically and a practical method for automatically preventing oscillation occurring in them is suggested. Experimental results obtained with a suitable circuit are presented.

621.375.3 **2339**

Some Aspects of Half-Wave Magnetic Amplifiers.—G. M. Ettinger. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 237–248. Discussion, pp. 266–272.) The performance of half-wave magnetic amplifiers is analysed and a theory presented which allows for a finite control-circuit resistance, or rectifier reverse conductance, and imperfect core properties. Experimental results obtained with amplifiers using different cores show good agreement with theory. Various methods of bias are reviewed and a conductance-controlled amplifier is described.

621.375.3 **2340**

A Fast-Response Full-Wave Magnetic Amplifier.—G. E. Lynn. (*Commun. & Electronics*, March 1958, No. 35, pp. 37–41.)

621.375.3 **2341**

The Effective Feedback Factor of Self-Balancing Magnetic Amplifiers.—W. A. Geyger. (*Commun. & Electronics*, March 1958, No. 35, pp. 66–69.)

621.375.4.011.21 **2342**

Three Output Immittance Theorems.—H. Stockman. (*Electronic Ind.*, Jan. 1958, Vol. 17, No. 1, pp. 61–63..156.) A discussion of various theorems having practical application in the analysis of transistor circuits.

621.375.4.018.783 **2343**

Nonlinear Distortion in Transistor Amplifiers.—G. Meyer-Brötz & K. Felle. (*Elektronische Rundschau*, Oct. 1957, Vol. 11, No. 10, pp. 297–301.) Following an analysis of the causes of distortion in common-emitter circuits the distortion factor is derived from the static characteristics. The dependence of distortion on temperature, collector bias, input signal level and load resistance is determined, and calculated values are compared with results of measurements.

621.375.4.024 **2344**

Some Transistor Input Stages for High-Gain D.C. Amplifiers.—G. B. B. Chaplin & A. R. Owens. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 249–257. Discussion, pp. 266–272.) Several types of input stage are investigated and methods of reducing drift are suggested. It is shown that drift of operating point in transistor d.c. amplifiers depends on the type of circuit, the lowest drift being obtained using the transistor as a chopper.

621.375.4.024 **2345**

A Transistor High-Gain Chopper-Type D.C. Amplifier.—G. B. B. Chaplin & A. R. Owens. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 258–266. Discussion, pp. 266–272.) A modulated system is employed consisting of a transistor input chopper, a high-gain transistor a.c. amplifier and an output chopper. The system has an open loop gain of 500 V/μA with a band extending from direct current to 25 c/s. Peak output is ± 10 V and current drift referred to input is 4×10^{-9} A in the range 20° C–50° C. The voltage drift at the input is less than 100 μA.

621.375.4.029.3 **2346**

Designing Transistor A.F. Power Amplifiers.—M. B. Herscher. (*Electronics*, 11th April 1958, Vol. 31, No. 15, pp. 96–99.) The amplifier described delivers 45 W to a 4-Ω load. It contains a driver stage, a phase splitter using a transistor pair, and a push-pull output stage.

621.375.9 : 538.569.4.029.6 **2347**

Operation of a Solid-State Quantum-Mechanical Amplifier.—M. W. P. Strandberg, C. F. Davis, B. W. Faughnan, R. L. Kyhl, & G. J. Wolga. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1988–1989.) "The operation of an S-band solid-state quantum-mechanical amplifier operated at 4·2°K with a computed noise temperature of less than 4·5°K is compared with similar devices and is used as proof of the condition of discrete phonon saturation which has been previously postulated." See also 1691 of June (McWhorter & Meyer).

621.375.9 : 538.569.4.029.6 **2348**

The Solid-State Maser—a Supercooled Amplifier.—J. W. Meyer. (*Electronics*, 25th April 1958, Vol. 31, No. 17, pp. 66–71.) A review of maser techniques with descriptions of two-level molecular-beam and three-level solid-state devices. Amplifier and oscillator characteristics, noise measurements and future developments are discussed.

621.375.9 : 538.569.4.029.6 **2349**

A U.H.F. Solid-State Maser.—R. H. Kingston. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 916.) Maser action at a frequency of 300 Mc/s has been obtained by using a cavity mode at the pumping frequency and a lumped resonant circuit at the amplifying frequency.

621.375.9 : 538.569.4.029.6 **2350**

Design Considerations for Circulator Maser Systems.—F. R. Arams & G. Krayer. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 912–913.) A general expression for effective maser system noise temperature is calculated and a working approximation is derived. The noise and stability requirements for a low-noise circulator maser system are given in terms of maser gain and of dissipative losses and s.w.r. in various parts of the system.

621.375.9 : 538.569.4.029.6 : 621.396.822 **2351**

System-Noise Measurement of a Solid-State Maser.—A. L. McWhorter & F. R. Arams. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 913–914.) A complete amplifier system with an effective input noise temperature of 20 ± 5°K is described. The sources of noise are analysed and it is shown that the noise temperature of the maser proper is close to the theoretical value.

621.375.9 : 538.569.4.029.64 **2352**

Electron Spin and Phonon Equilibrium in Masers.—Bloembergen. (See 2385.)

621.375.9 : 621.372.2 **2353**

A Travelling - Wave Parametric Amplifier.—A. L. Cullen. (*Nature, Lond.*, 1st Feb. 1958, Vol. 181, No. 4605, p. 332.) A lossless transmission line with periodically varying distributed inductance and a constant distributed capacitance will support a growing current wave. For maximum amplification the oscillator used for providing the periodic variation of inductance must be synchronized with the incoming signal. The noise figure of such an amplifier should be very favourable. An experimental investigation is in progress, and related schemes are being studied in which precise frequency and phase relations are not demanded.

621.375.9.029.6 : 537.311.33 **2354**

Proposed Negative-Mass Microwave Amplifier.—H. Krömer. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, p. 1856.) It is pointed out that negative masses occur in semiconductors at energies close to the band edge if the energy contours are re-entrant there. It is suggested that such a semiconductor might be used as the active element in a microwave amplifier, although preliminary experiments with germanium did not show the effect. See also 2901 of 1954 (Shockley & Mason).

GENERAL PHYSICS

533.6.011 **2355**

On Chandrasekhar's Theory of Turbulence.—P. C. Jain. (*Proc. nat. Inst. Sci. India.*, Part A, 26th Nov. 1957, Vol. 23, No. 6, pp. 504–513.) The theory (*Proc. roy. Soc. A*, 5th April 1955, Vol. 229, No. 1176, pp. 1–19) has been generalized to the case of axisymmetric turbulence, and eight differential equations in eight defining scalars of double and triple correlation tensors deduced. Two of these equations replace that of de Kármán & Howarth (*Proc. roy. Soc. A*, 21st Jan. 1938, Vol. 164, No. 917, pp. 192–215) in the theory of isotropic turbulence.

533.6.011 **2356**

On Decay of Energy Spectrum of Isotropic Turbulence.—N. R. Sen. (*Proc. nat. Inst. Sci. India*, Part A, 26th Nov. 1957, Vol. 23, No. 6, pp. 530–533.)

535.215 : 537.224 **2357**

Electric Field Distribution in Polarized Photoconductors.—H. Kallmann & J. R. Freeman. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1506–1508.) The distribution is determined from changes which occur in the persistent polarization [see 2633 of 1955 (Kallmann & Rosenberg)] under various polarizing conditions.

537.12 : 537.311.1 **2358**

The Fundamental State of an Electron Gas.—H. Koppe. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 135–155.) Detailed discussion with reference to Mayer's treatment of electron correlation energy (*Phys. Rev.*, 15th Dec. 1955, Vol. 100, No. 6, pp. 1579–1586).

537.226 **2359**

Force Densities in Dielectrics.—É. Durand. (*C. R. Acad. Sci., Paris*, 16th Sept. 1957, Vol. 245, No. 12, pp. 1003–1006.)

537.311.1 **2360**

A New Approach to the Theory of Electrical Conductivity of Solids.—B. Meltzer. (*Physica*, Feb. 1957, Vol. 23, No. 2, pp. 118–124.) Since the electrical conductance of a body is a measure of the fluctuations of charge in a state of thermodynamic equilibrium an analysis of these fluctuations gives the value of the conductivity. The value derived agrees with that of Drude but is found to be half that for the Lorentz-Sommerfeld model.

537.311.1 **2361**

The Boltzmann Equation in the Theory of Electrical Conduction in Metals.—D. A. Greenwood. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 585–596.) The Boltzmann equation for electrical conduction in metals is derived assuming $\hbar/\tau \ll kT$, where τ is the collision time. An expression independent of this assumption is considered but not evaluated.

537.311.1 **2362**

Theory of Electrical Conductivity of Anisotropic Inhomogeneous Media.—A. Nedoluha. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 248–252.)

537.311.31 : 538.63 **2363**

Quantum Theory of the Electrical Conductivity of Metals in a Magnetic Field.—I. M. Lifshits. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1509–1518.) The kinetic equation for the density matrix is derived and the asymptotic values of the kinetic coefficients in a strong magnetic field are analysed.

537.56 **2364**

On the Theory of Spectral-Line Broadening in Plasma.—V. I. Kogan. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Feb. 1958, Vol. 118, No. 5, pp. 907–910.)

537.56 **2365**

Structure of Shock Waves in a Plasma.—V. D. Shafranov. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1453–1459.) Investigation of a shock wave in a plasma taking account of the difference in electron and ion temperatures. Three cases are examined: (*a*) nonstationary shock wave, (*b*) stationary shock wave, (*c*) stationary shock wave in a strong magnetic field.

537.56 **2366**

Electrodynamical Acceleration of Plasma Bunches.—L. A. Artsimovich, S. Yu. Luk'ianov, I. M. Podgornyĭ & S. A. Chuvatin. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 3–8.) A method is described of generating a gas-discharge plasma by the electric explosion of a 0·02-mm-diameter copper wire in a discharge chamber evacuated to 1–2 × 10^{-6} mm Hg. A 75-μF capacitor is used with initial voltage 30 kV.

537.56 **2367**

Quantum Kinetic Equation for Plasma taking Account of Correlation.—Yu. L. Klimontovich & S. V. Temko. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 132–134.) A quantum kinetic equation for a system of particles with Coulomb interaction is derived taking into account the correlation of the mutual position of the charged particles.

537.56 **2368**

On the Effective Field in a Plasma.—B. B. Kadomtsev. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 151–157.) Equations for the partial distribution functions are used to compute the effective field acting on charged particles in a plasma. The effective field differs from the mean field by a small quantity of the order of $1/N$ where N is the number of particles within a sphere of radius equal to the Debye radius.

537.56 : 538.56 **2369**

The Dispersion Equation for Plasma Waves.—N. G. Van Kampen. (*Physica*, July 1957, Vol. 23, No. 7, pp. 641–650.) The validity of different formulae is discussed. When collisions are infrequent no dispersion equation exists except in special cases. See also 2029 of 1956.

537.56 : 538.569 **2370**

On the Dynamics of a Bounded Plasma in an External Field.—L. M. Kovrizhnykh. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 72–76.) An investigation of the dynamics of a quasi-neutral plasma formation located in the field of a plane e.m. wave. The method of successive approximations is used. Within the limits of the assumptions of the analysis a plasma bunch tends to spread out.

537.56 : 538.6 : 52 **2371**

An Energy Principle for Hydromagnetic Stability Problems.—I. B. Bernstein, E. A. Frieman, M. D. Kruskal & R. M. Kulsrud. (*Proc. roy. Soc. A*, 25th Feb. 1958, Vol. 244, No. 1236, pp. 17–40.) The problem of the stability of static, highly conducting, fully ionized plasmas is investigated by means of an energy principle developed from one introduced by Lundquist (122 of 1952). The method is applied to the general axisymmetric system and to plasmas which are completely separated from the magnetic field by an interface.

537.56 : 538.63 : 538.56 **2372**

Radiation of Plasma in a Magnetic Field.—B. A. Trubnikov. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Feb. 1958, Vol. 118, No. 5, pp. 913–916.) Mathematical analysis of e.m. radiation in plasma taking account of electron spin in a uniform magnetic field.

537.56 : 538.63 : 538.566 **2373**

Coherent Scattering and Radiation of Electromagnetic Waves by Plasma in an Inhomogeneous Magnetic Field.—G. A. Askar'yan. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1576–1577.) The propagation of quasi-plane waves in a waveguide filled with plasma in a longitudinal magnetic field and the radiation in an electron plasma of a bunch which travels through a spatially periodic axially symmetric magnetic field are discussed. The plasma tends to focus the bunches and a longitudinal focusing effect results from the attraction between the induced currents flowing in the same direction.

538.114 **2374**

Models Demonstrating the Phenomena of Ferromagnetic Hysteresis.—H. Wilde. (*Nachrichtentech. Z.*, Oct. 1957, Vol. 10, No. 10, pp. 497–502.) Extension and refinement of the Preisach model. See also 842 of 1957 (Feldtkeller & Wilde).

538.222 : 539.2 **2375**

Contribution to the Phenomenological Theory of Paramagnetic Relaxation in Parallel Fields.—N. K. Belousova & I. G. Shaposhnikov. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 238–242.) The role of spin-lattice interaction is examined and the applicability of the results of this analysis to paramagnetic absorption at high frequencies and temperatures of the order of 20°K is considered.

538.3 **2376**

The Classical Electromagnetic Equations expressed as Complex Four-Dimensional Quantities.—E. F. Bolinder. (*J. Franklin Inst.*, March 1957, Vol. 263, No. 3, pp. 213–223.) It is possible to compress the classical e.m. equations into three complex four-dimensional quantities representing respectively e.m. fields, currents and charges, and vector and scalar potentials. This notation is useful in problems relating to waveguides, cavities and aerials for microwaves.

538.3 : 52 **2377**

Theory of Magnetohydrodynamic Waves using the Energy Pulse Tensor of Abraham.—E. Richter. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 253–261.) See also 2423 of 1957.

538.3 : 52 **2378**

Dynamo Effect in Magnetohydrodynamic Theory.—S. Colombo. (*Rev. gén. Élect.*, June 1957, Vol. 66, No. 6, pp. 325–332.)

538.52 : 538.566 **2379**

Surface Currents Induced by Short-Wavelength Radiation.—J. A. Cullen. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1863–1867.) "The integral equation given by Fock [1412 of 1947] for the surface currents induced in a perfectly conducting convex object by short-wavelength plane electromagnetic waves is solved directly by the method of Fourier transforms. The result agrees with Fock's indirect approach. Some general properties of the equation are also discussed."

538.561 : 537.56 : 53.087 : 621.383.27 **2380**

The Cherenkov Effect Produced by Single Particles in Gases.—A. Ascoli-Balzanelli & R. Ascoli. (*Nuovo Cim.*, 1st Dec. 1957, Vol. 6, No. 6, pp. 1392–1408. In English.) The application of photomultipliers for detecting Cherenkov radiation is discussed and successful experimental work is reported.

538.566 : 535.42 **2381**

A New Formulation of Scalar Diffraction Theory for Restricted Aperture.—J. M. Cowley & A. F. Moodie. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 533–545.)

538.566 : 537.53 **2382**

Radiation of Charged Particles in Flight Past Ideally Conducting Bodies.—Yu. N. Dnestrovskiĭ & D. P. Kostomarov. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Sept. 1957, Vol. 116, No. 3, pp. 377–380.) The problem is discussed in the nonrelativistic approximation and the induced charge and currents are calculated initially without considering retardation. The radiation due to the current system determined is then estimated. The dipole approximation is used to derive expressions for the rate at which energy is radiated. The radiation resistance for a waveguide system is calculated and its dependence on the particle velocity and on the radius/wavelength ratio is shown graphically.

538.566.029.65 : 537.228.5 **2383**

Stark Effect at 2·0 and 1·2 Millimetres Wavelength: Nitric Oxide.—C. A. Burrus & J. D. Graybeal. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1553–1556.) "Stark-effect measurements have been made for the first time in the 1–2-mm wave region. Results of Stark-effect measurements on the $J = \frac{1}{2} \rightarrow 3/2$ and $J = 3/2 \rightarrow 5/2$ transitions of the $^2\Pi_{\frac{1}{2}}$ ground state of $N^{14}O^{16}$, falling at 2·0 and 1·2 mm, respectively, are reported, and are shown to fit closely existing theory; the electric dipole moment of NO in the ground state is found to be $0{\cdot}158 \pm 0{\cdot}006$ Debye unit."

538.569.4.029.6 **2384**

Electric Susceptibility of Ethyl Chloride in the Centimetre Region.—G. P. Srivastava. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 242–247. In English.) Results of measurements at 8 780 Mc/s over a pressure range 10–75 cm Hg and a temperature range 263°–353°K are discussed. See also 3841 of 1957 (Krishnaji & Srivastava).

538.569.4.029.64 : 621.375.9 **2385**

Electron Spin and Phonon Equilibrium in Masers.—N. Bloembergen. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 2209–2210.) It is pointed out that the successful operation of $KCo(Cr)(CN)_6$ in three-level steady-state masers is incompatible, regardless of the operating frequencies of the maser, with the assumption that the relaxation rates are determined by the interaction between the lattice vibrations and the helium bath [see *ibid.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 302–311 (Giordmaine et al.)].

539.1.01 : 538.569.4 **2386**

Identity of Spin Temperature and Thermodynamic Temperature.—A. Abragam & W. Proctor. (*C. R. Acad. Sci., Paris*, 23rd Sept. 1957, Vol. 245, No. 13, pp. 1048–1050.)

539.2 : 537.311.1 **2387**

Generalized Mobility Theory.—M. Lax. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1921–1926.) A formal theory of mobility is presented that does not depend on the existence of a transport equation. In particular the Hamiltonian describing the electron plus the scattering system is not decomposed into an unperturbed part plus a perturbation. Only the applied field is treated as small. The result is shown to reduce to the usual transport result when the scattering perturbation is weak, without assuming the existence of a relaxation time. The relation between a many-electron treatment and the one-electron treatment is demonstrated for the case of Fermi as well as Boltzmann statistics.

539.2 : 537.311.31 : 061.3 **2388**

The Band Theory of Metals.—(*Nature, Lond.*, 22nd Feb. 1958, Vol. 181, No. 4608, pp. 525–527.) Brief report of a conference held by the Physical Society at Imperial College, London, 20th–21st December 1957. Some 20 papers are noted.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164 : 523.35 **2389**

A 'Lunar' Radio Interferometer.—G. A. Gurzadyan. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Feb. 1958, Vol. 118, No. 6, pp. 1094–1097.) In a two-aerial interferometer used in radio astronomy one of the aerials can be replaced by the surface of the sea. In the instrument described the ionosphere of the moon is used in place of the second aerial. With metre wavelengths the minimum angle of refraction of the lunar ionosphere is found to be of the order of 0·1′–0·2′.

523.164.3 : 523.35 **2390**

Radio Interference Phenomena due to the Ionosphere of the Moon.—G. A. Gurzadyan. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Feb. 1958, Vol. 118, No. 5, pp. 884–887.) An attempt to explain the increase of solar radio emission observed before and after a solar eclipse, by interference effects due to refraction of radio waves in the lunar ionosphere.

523.164.32 **2391**

A Remarkable Solar Radio Event.—(*Nature, Lond.*, 22nd Feb. 1958, Vol. 181, No. 4608, pp. 542–543.) Increased solar radio noise at a frequency of 200 Mc/s was recorded at Nera Observatory, Netherlands, from 0848 U.T. on 4th November 1957. The maximum value was approximately 900 times the noise level of the quiet sun. Fluctuations with a period of 0·2–0·3 s occurred and are attributed to solar phenomena.

523.164.32 : 621.396.677.833 **2392**

A Radio Image of the Sun on 3·2-cm Wavelength.—V. V. Vitkevich, A. D. Kuz'min, A. E. Salomonovich & V. A. Udal'tsov. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Feb. 1958, Vol. 118, No. 6, pp. 1091–1093.) In June 1957 at a station of the Physical Institute in the Crimea a new stationary radio telescope, with a parabolic reflector of 31 in. diameter, was put into commission. Isophotes of the sun obtained at 3·2 and 10·0 cm λ are shown.

523.165 **2393**

Energy Spectrum of Cosmic Radiation.—H. Alfvén & E. Åström. (*Nature, Lond.*, 1st Feb. 1958, Vol. 181, No. 4605, pp. 330–331.) A theory is given for the observed spectrum f = const. p^{-n} where n = 2·5 for a wide range of values of momentum p.

523.5 : 621.396.11 **2394**

Diurnal Variations in the Number of Shower Meteors Detected by the Forward Scattering of Radio Waves: Part 3—Ellipsoidal Theory.—C. O. Hines. (*Canad. J. Phys.*, Jan. 1958, Vol. 36, No. 1, pp. 117–126.) The ellipsoidal geometry inherent in the forward-scatter process is developed fully. The more rigorous results obtained are compared at various stages in the analysis with those given by the simplifying 'cylindrical approximation' developed in Part 1 (738 of 1956). The 'potentially observable' trails are first located and their distribution determined. The fraction of these providing a signal exceeding some given level is found and the total number of these observable trails as a function of trail orientation is determined by integrating over the spatial distribution. Part 2: 1046 of 1956 (Forsyth et al.).

523.5 : 621.396.11 **2395**

Resonance Effects in the Theory of Meteor Observability.—D. R. Moorcroft & C. O. Hines. (*Canad. J. Phys.*, Jan. 1958, Vol. 36, No. 1, pp. 134–136.) The observability theory for forward scattering is revised to cover the resonance that occurs when there is a component of the incident electric vector transverse to the axis of the meteor trail. Normalized contours are given.

523.75 : 523.165 **2396**

Solar Activity and Cosmic Radiation.—R. R. Brown. (*J. Geomag. Geoelect.*, 1957, Vol. 9, No. 2, pp. 79–85.) Experimental observations of cosmic radiation during intense solar activity are presented and examined with particular attention to the period around the solar flare of 23rd February 1956. The results obtained are discussed in relation to models of solar activity involving corpuscular streams.

523.75 : 550.385.4 **2397**

On the Asymmetry of the Heliographic Distribution of the Storm-Producing Probability of Flares.—Y. Hakura. (*J. Radio Res. Labs, Japan*, Jan. 1958, Vol. 5, No. 19, pp. 57–64.) The probability is greatest for flares near the centre of the disk and for those in the NE and SW quadrants. The greater probability for flares in the western hemisphere is not statistically significant.

523.755 : 523.72 **2398**

The Distribution over the Emitting Area of X Radiation from the Solar Corona and the Residual Intensity during Total Solar Eclipses.—G. Elwert. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 187–199. In German.) The E layer is assumed to be ionized by X rays of approximate wavelength 50–100 Å. The intensity of coronal radiation in this waveband is calculated and the results lead to a considerable degree of limb brightening and to emission of radiation at radii greater than that of the visible disk. The intensity of the residual radiation during total eclipses can be calculated and is estimated to have been about 20% in 1952 and 1954, and about 10% in 1955.

550.385 : 523.75 **2399**

Magnetic Activity following a Solar Flare.—M. A. Ellison. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 214–215.) The conclusions reached by Watson (3853 of 1957) that solar flares and increases of magnetic activity are not correlated is not correct. When the magnitude of the flare and its position on the disk are taken into account, there is a positive correlation as was shown by Newton (see e.g. 3398 of 1944).

550.389.2 **2400**

International Geophysical Year.—H. Odishaw. (*Science*, 17th Jan. 1958, Vol. 127, No. 3290, pp. 115–128.) A report on the U.S. program with a brief account of some of the activities during the first five months.

550.389.2 : 629.19 **2401**

Estimating the Life of a Satellite.—J. M. C. Scott. (*Nature, Lond.*, 28th Dec. 1957, Vol. 180, No. 4600, pp. 1467–1468.)

550.389.2 : 629.19 **2402**

Lifetime of an Artificial Earth Satellite.—D. C. M. Leslie. (*Nature, Lond.*, 8th Feb. 1958, Vol. 181, No. 4606, pp. 403–404.) The approximate formulae given by Fejer (2087 of July) and Scott (2401 above) are compared with those given by a more exact method and found to be accurate provided the eccentricity is not large.

550.389.2 : 629.19 **2403**

British Radio Observation of the Satellite.—(*Brit. Commun. Electronics*, Dec. 1957, Vol. 4, No. 12, pp. 770–772.) Brief reports of the observations made on the first Russian satellite at the B.B.C. Station, Tatsfield, the Radio Research Station, Slough, and Ministry of Supply establishments are given. These include measurements of the signal characteristics, field strength, Doppler frequency shift and bearing and elevation of the satellite.

551.510.535 **2404**

The Effect of Diffusion on the Vertical Distribution of Ionization in a Quiet F Region.—V. C. A. Ferraro & I. Özdoğan. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 140–149.) Electrons are assumed to disappear by attachment to neutral molecules and by diffusion. The coefficient of attachment is assumed to be constant, while the diffusion constant is calculated according to classical theory. Using rocket measurements of air density and temperature, the vertical distribution of electrons can be calculated. The results are at variance with those actually observed, possibly because the diffusion constant is much less than that assumed.

551.510.535 **2405**

Ambipolar Diffusion in the F_2 Layer.—A. Dalgarno. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 219–220.) Diffusion in the F_2 layer probably depends on the diffusion of O^+ in O; the coefficient is estimated to be about one quarter of the value normally assumed to be applicable in the F_2 layer.

551.510.535 **2406**

Horizontal Wind Systems in the Ionospheric E Region Deduced from the Dynamo Theory of the Geomagnetic *Sq* Variation.—(*J. Geomag. Geoelect.*, 1957, Vol. 9, No. 2.)

Part 3—H. Maeda (pp. 86–93). Wind systems are deduced from *Sq* data for the solstice seasons in both northern and southern hemispheres and the diurnal and semi-diurnal components are examined. A comparison of the theoretical system with observed winds is made.

Part 4—S. Kato (pp. 107–115). Wind velocities are calculated taking account of Coriolis forces and assuming the electron density at night to be 1/12th that at noon.

Parts 1 & 2: 1080 of 1957.

551.510.535 **2407**

Non-Chapmanlike Variations in the Ionospheric E and F_1 Layers. Effect of the *Sq* Current System: Part 2.—T. Shimazaki. (*J. Radio Res. Labs, Japan*, Jan. 1958, Vol. 5, No. 19, pp. 35–56.) Variations in f_0E and f_0F_1 are examined in various months in addition to those of March which have already been discussed (see 3472 of 1957). It is shown that, in non-equinoctial months, f_0E is larger in the winter hemisphere than in the summer hemisphere for the same solar zenith distance. This unsymmetrical distribution is interpreted as being due to the *Sq* current system effect in the E layer, the *Sq* current system developing more strongly in summer than in winter. Somewhat similar results are observed for the F_1 layer but detailed examination shows that they can be explained by the temperature difference between the summer and winter hemispheres.

551.510.535 : 523.164 **2408**

Radio-Star Scintillations at an Equatorial Station.—J. R. Koster. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 100–109.) Scintillation of radio stars has been measured at 45 Mc/s in Ghana for four years. It is unusually intense, occurs only at night, and is correlated with spread echoes from the F layer and with equatorial scatter of radio signals near sunspot maximum.

551.510.535 : 523.164 **2409**

The Diurnal and Seasonal Variations of Spread-F Ionospheric Echoes and the Scintillations of a Radio Star.—B. H. Briggs. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 89–99.) The degree of spreading is greater at Inverness (57°N) than at Slough (52°N) but the diurnal and seasonal variations are similar and can be correlated with the scintillations of the radio star in Cassiopeia provided an allowance is made for the zenith angle of the star. See also 1737 of June.

551.510.535 : 523.746 **2410**

A Method of Determining the Correlation between f_0F_2 and Sunspot Number.—H. Shibata. (*J. Radio Res. Labs, Japan*, Jan. 1958, Vol. 5, No. 19, pp. 65–73.) The monthly median critical frequency f and the smoothed sunspot number R can be represented for a month m and time of day t by $f\ (m,t) = a\ (m,t)\ R + b\ (m,t)$. The continuous periodic functions a and b are examined using data obtained at six places in Japan.

551.510.535 : 523.75 : 621.396.11 **2411**

Solar Activity and the Ionosphere.—C. M. Minnis. (*Nature, Lond.*, 22nd Feb. 1958, Vol. 181, No. 4608, pp. 543–544.) Observations at Slough showed that the intensity of the ultraviolet and X radiation responsible for the existence of the E layer of the ionosphere, was higher in September, October and November 1957 than ever before recorded, and that on individual days in November the critical frequency of the F_2 layer at noon was above 16 Mc/s. See also 1153 of April (Bennington).

551.510.535 : 550.385 **2412**

Disturbances of the Ionospheric F_2 Region associated with Geomagnetic Storms: Part 3—Auroral Latitudes.—T. Sato. (*J. Geomag. Geoelect.*, 1957, Vol. 9, No. 2, pp. 94–106.) Results of a statistical treatment show that the deviation of f_0F_2 on a disturbed day is similar to the negative disturbance in middle latitudes (see Part 2: 127 of January); the deviation of h_pF_2 is different. Main features of the F_2 disturbances can be accounted for by vertical electron drift.

551.510.535 : 621.396.11 **2413**

Electron Density Fluctuations and Scattering of Radio Waves in the Ionosphere.—Ya. L. Al'pert. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 213–225.) Discussion of theoretical and experimental determinations of electron density fluctuations and of energy scattered at v.h.f. indicates that in vertical-incidence soundings at frequencies below the critical frequency the scattered field is chiefly due to waves first scattered in the forward direction and then reflected from higher levels in the ionosphere.

551.510.535 : 621.396.11 **2414**

Control of the Ionosphere by means of Radio Waves.—V. A. Bailey. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 216–217.) A note on the possibility of using radio waves near the gyrofrequency of electrons in the ionosphere to control the electron temperature and hence the rates of diffusion, attachment and recombination.

551.510.535 : 621.396.11.029.62 **2415**
: 629.136.3

An Artificial Ionosphere.—(*Brit. Commun. Electronics*, Feb. 1958, Vol. 5, No. 2, p. 117.) The release from rockets of potassium into the high atmosphere produces an ionized cloud, lasting for about an hour, suitable for returning microwave signals.

551.594.2 **2416**

Waveform Studies of Electric Field Changes during Cloud-to-Cloud Lightning Discharges.—B. A. P. Tantry, R. S. Srivastava & S. R. Khastgir. (*Proc. nat. Inst. Sci. India*, Part A, 26th Nov. 1957, Vol. 23, No. 6, pp. 499–503.) Assuming that cloud-to-cloud discharges may be identified from the sign of the field change and the time separation of the individual strokes, such discharges exhibit all the features known to be associated with ground discharges. They include: (*a*) precursors, (*b*) multiple strokes, (*c*) return strokes, (*d*) slow field changes and 'hook' components, (*e*) junction field changes between strokes.

551.594.5 **2417**

Auroral Display Observed from Unusually Low Geomagnetic Latitudes.—B. McInnes. (*Met. Mag. Lond.*, April 1957, Vol. 86, No. 1018, pp. 114–117.) Observations of the aurora display of 8th September 1956 are reported with brief details of some related effects.

551.594.5 **2418**

Horizontal Movements of Aurora.—J. S. Kim & B. W. Currie. (*Canad. J. Phys.*, Feb. 1958, Vol. 36, No. 2, pp. 160–170.) The drift of auroral forms has been measured at three stations in Canada south of the auroral zone. No evidence is found of a motion due to the earth's rotation relative to a fixed point in space. The distribution and magnitude of the drift speeds parallel and normal to the geomagnetic meridians are similar to those found for ionospheric drifts. The drift speeds increase with geomagnetic activity particularly in an E-W direction.

551.594.5 : 621.396.96 **2419**

A Continuously Recording Automatic Auroral Radar.—A. G. McNamara. (*Canad. J. Phys.*, Jan. 1958, Vol. 36, No. 1, pp. 1–8.) A low-power 50-Mc/s system with fixed aerials is used to give a continuous photographic record. The echo intensity is also integrated electronically to give a chart record which is immediately available.

551.594.6 **2420**

Abnormal Polarization of the Atmospheric Pulses Reflected Successively from the Ionosphere.—S. R. Khastgir. (*Nature, Lond.*, 8th Feb. 1958, Vol. 181, No. 4606, pp. 404–405.) Elliptic patterns of gradually decreasing eccentricity and tilt angle have been observed in cathode-ray direction finders used to record atmospherics due to lightning discharges at a distance of about 500 km.

551.594.6 : 621.396.11.029.45 **2421**

Electromagnetic Radiation from Lightning Strokes.—E. L. Hill. (*J. Franklin Inst.*, Feb. 1957, Vol. 263, No. 2, pp. 107–119.) A theory is given of the spectral distribution and the amount of v.l.f. radiation emitted from a vertical lightning stroke from cloud to ground. The radiated energy has a maximum intensity at about 11 kc/s and a total width at half-maximum of 12 kc/s. The total energy radiated in one leader and return stroke is estimated to be about 220 000 joules.

551.594.6 : 621.396.822 **2422**

Atmospheric Radio Noise at Frequencies between 10 kc/s and 30 kc/s.—J. Harwood. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 293–300.) The results are described in terms of a voltage envelope at the output of a narrow-band receiver (300 c/s between 3 dB points). The r.m.s. voltage deduced by integrating the measured amplitude distributions varies between four and eight times the average voltage. The noise is always much more impulsive than fluctuation noise. The structure of the voltage envelope is considered in two ways: (*a*) the amplitude distribution of pulse peaks, and (*b*) the amplitude distribution of voltage. See also 445 of 1957 (Horner & Harwood).

LOCATION AND AIDS TO NAVIGATION

621.396.933.1 **2423**

Marconi Doppler Navigator.—(*Wireless World*, June 1958, Vol. 64, No. 6, pp. 260–261.) Further technical details are given and performance during a demonstration flight is described. See also 3146 of 1957.

621.396.96 : 551.594.5 **2424**

A Continuously Recording Automatic Auroral Radar.—McNamara. (See 2419.)

621.396.969.33 **2425**

A Doppler Collision Course Indicator for Use at Sea.—H. R. Whitfield & C. M. Cade. (*J. Inst. Nav.*, Jan. 1958, Vol. 11, No. 1, pp. 81–87.) A system is proposed for the instantaneous measurement of the radial component of relative velocity. This, together with the range and bearing data from a p.p.i., gives the aspect of the target. Possible errors in deducing true aspect and speed are discussed and the mode of operation and principal design parameters of a suitable Doppler radar are given.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 537.311.33 **2426**

Temperature Dependence of Photo-electromotive Force.—F. F. Kodzhespirov. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1593–1594.) The difficulties of reconciling experimental results with theories concerning photoelectric effects are considered. The temperature dependence of the e.m.f. is discussed with reference to amorphous Se exposed to X rays of different intensity.

535.215 : 546.472.21 **2427**

Photovoltages Larger than the Band Gap in Zinc Sulphide Crystals.—S. G. Ellis, F. Herman, E. E. Loebner, W. J. Merz, C. W. Struck & J. G. White. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, p. 1860.) Ultraviolet illumination produces photovoltages greater than the band gap in ZnS crystals grown from the vapour. The crystals contained random stacking faults.

535.215 : 546.482.21 : 538.6 **2428**

Surface Photoconductivity of Cadmium Sulphide Modified with Magnetic Field.—S. Tanaka & T. Masumi. (*J. phys. Soc. Japan*, Jan. 1958, Vol. 13, No. 1, pp. 22–32.) Report and discussion of measurements made on CdS single crystals.

535.37 **2429**

Energy Storage in ZnS and ZnCdS Phosphors.—H. Kallmann & E. Sucov. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1473–1478.) It was found that the trap concentration was of the order of 10^{15}/ cm^3 in all the phosphors examined. The observed slow rate of decay of the stored energy is explained by predominant retrapping in traps of various depths. It is concluded that the decay of stored electrons can take place radiatively and/or nonradiatively and that the rate for both processes is of the same order of magnitude. It is also shown that the light sum under a stimulation curve does not give a true picture of the trap population.

535.37 **2430**

Temperature Characteristics of Barium Strontium Lithium Silicate Phosphors.—A. H. McKeag. (*J. electrochem. Soc.*, Feb. 1958, Vol. 105, No. 2, pp. 78–81.) "The changes in spectral emission, efficiency, and temperature stability are discussed which result from changes in the following variable parameters in phosphor preparation: the Ba : Sr ratio, Li content, base : acid ratio, activator concentrations, and firing conditions."

535.371.07.002.2 : 621.397.62 : 535.623 **2431**

Photodeposition of Luminescent Screens.—M. Sadowsky & P. D. Payne, Jr. (*J. electrochem. Soc.*, Feb. 1958, Vol. 105, No. 2, pp. 105–107.) The method of deposition of the phosphor on a c.r. tube screen for colour television is described, with details of how to overcome faults which are liable to occur.

535.376 **2432**

The Luminescence Response of Phosphors to Low-Energy Ion Bombardment.—C. F. Eve & H. E. Duckworth. (*Canad. J. Phys.*, Jan. 1958, Vol. 36, No. 1, pp. 104–116.) The response of ZnS-Ag and Zn_2SiO_4-Mn was determined as a function of the ion energy and the form of the dependence derived. Experimental results for ZnS-Ag are consistent with a theory that the bombarding particles penetrate the phosphor as neutral atoms and produce luminescence by electronic excitation.

537.226/.227 : 546.431.824-31 **2433**

Surface- and Space-Charge Effects in Ceramic Barium Titanate.—W. R. Büssem & E. C. Subbarao. (*Naturwissenschaften*, Oct. 1957, Vol. 44, No. 19, p. 509.) Interim report on the results of X-ray tests on $BaTiO_3$ specimens subjected to ultraviolet radiation and to various surface treatments.

537.226/.227 : 546.817.824 **2434**

Dielectric Properties of Lead Titanate and its Solid Solutions at Low Temperature under Strong Electric Field.—S. Nomura & J. Kobayashi. (*J. phys. Soc. Japan*, Jan. 1958, Vol. 13, No. 1, pp. 114–115.)

537.226.3 : 621.315.616 **2435**

On Dielectric Absorption in Plastic Insulators.—A. Kelen. (*Phys. Chem. Solids*, April 1957, Vol. 2, No. 2, pp. 150–151.)

537.226.31 **2436**

Motion of a Charge Parallel to the Axis of a Cylindrical Channel in a Dielectric.—L. S. Bogdankevich & B. M. Bolotovskiĭ. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1421–1428.) The field produced by a charge moving parallel to the axis of a cylindrical channel is determined. The field and energy losses of the charge are calculated for various conditions of the medium.

537.226.31 **2437**

Anomalous Dispersion of Permittivity in Feldspars.—V. A. Ioffe & G. I. Khvostenko. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Feb. 1958, Vol. 118, No. 4, pp. 709–712.)

537.228.1 : 549.514.51 **2438**

Theory of Plane Elastic Waves in a Piezoelectric Crystalline Medium and Determination of Elastic and Piezoelectric Constants of Quartz.—I. Koga, M. Aruga & Y. Yoshinaka. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1467–1473.) In the case of a plane elastic wave only the component of piezoelectric polarization in the direction of wave propagation produces a restoring force against mechanical strain. Thus the elastic and piezoelectric constants can be determined from experimental data. A discrepancy between theory and experiment in the results of earlier workers [see 1452 of 1941 (Atanasoff et al.)] is explained.

537.228.1 : 549.514.51 **2439**

Anelasticity of Synthetic Crystalline Quartz at Low Temperatures.—J. C. King. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1552–1553.) "This is a preliminary report of an examination of the anelasticity and frequency-temperature behaviour of synthetic quartz resonators, vibrating in thickness shear at low temperatures. The stress-induced relaxation absorption occurring at 50°K for a frequency of vibration of 5 Mc/s in natural quartz is found to be three orders of magnitude larger in some samples of the synthetic quartz."

537.311.33 **2440**

Theory of Interstitial Impurity States in Semiconductors.—P. E. Kaus. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1944–1952.) "The ionization energy and expectation value of the radius corresponding to the states of several interstitial impurities in Ge and Si are calculated. The range of the valence electron of the impurity is divided into two regions; an inner region, which is treated microscopically, and an outer region, which is treated macroscopically. The separation radius, which is primarily a function of the host crystal, is a parameter of the calculation. At a critical separation radius a rapid change of ionization energy and wave-function results. The calculations are carried out for several impurities in column I of the periodic table."

537.311.33 **2441**

On the Exciton Mechanism for Capture of Current Carriers in Homopolar Semiconductors.—V. L. Bonch-Bruevich. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1470–1478.) The capture of charge carriers by structural defects with energy transfer to small-radius excitons is investigated. The probability of this transfer and the temperature dependence of the recombination coefficients are also examined.

537.311.33 **2442**

On the Mass-Action Laws in Degenerate Semiconductors.—F. W. G. Rose. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 699–701.) An analysis which defines the effective density of the energy levels in the conduction band and valence band. See also 482 of February.

537.311.33 **2443**

Decay Law for the Concentration of Non-equilibrium Charge Carriers in Semiconductors.—G. M. Guro. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 158–165.) The kinetics of the decay of non-equilibrium carriers in a semiconductor containing a small concentration of traps is examined. In addition to monomolecular and bimolecular modes, a constant-rate recombination is identified. This occurs at low temperature when the recombination cross-section for minority carriers is much larger than that for the majority carriers.

537.311.33 **2444**

On the Equivalence of the Fokker-Planck Method and the Free-Energy Method for the Calculation of Carrier Density Fluctuations in Semiconductors.—K. M. Van Vliet. (*Physica*, March 1957, Vol. 23, No. 3, pp. 248–252.) The Fokker-Planck method is applied to a semiconductor with transitions between the valence band, the conduction band and one kind of impurity level. The results are consistent with the equations of Burgess (793 of 1956) obtained by the thermodynamic free-energy method.

537.311.33 **2445**

Electrical and Optical Properties of some $M_2^{Vb}N_3^{VIb}$ Semiconductors.—J. Black, E. M. Conwell, L. Seigle & C. W. Spencer. (*Phys. Chem. Solids*, 1957, Vol.2, No. 3, pp. 240–251.) A number of compounds of group Vb and VIb elements, with chemical formula $M_2^{Vb}N_3^{VIb}$, have been synthesized as single crystals of fairly good purity. Experimental results are interpreted according to the usual semiconductor model. Energy gaps range from 1·7 eV for Sb_2S_3 to 0·16 eV for Bi_2Te_3, decreasing as the atomic number of the components increases.

537.311.33 : 537.29 **2446**

Influence of an External Electric Field on the Adsorption Properties of a Semiconductor.—F. F. Vol'kenstein & V. B. Sandomirskiĭ. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Feb. 1958, Vol. 118, No. 5, pp. 980–982.) A theoretical treatment to determine the variation of absorption due to a uniform electric field directed perpendicularly to the semiconductor surface.

537.311.33 : 546.23 **2447**

The Field-Strength Dependence of the Electrical Conductivity of Selenium Single Crystals.—H. Gobrecht & H. Hamisch. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 218–232.) Report on investigations of the influence of a strong alternating field on the conductivity perpendicular to the field. It is concluded that the dependence of conductivity on field strength may be explained as being due to the influence of the field on the concentration of charge carriers. See also 2448 below.

537.311.33 : 546.23 : 537.228.1 **2448**

The Piezoelectric Effect in Selenium.—H. Gobrecht, H. Hamisch & A. Tausend. (*Z. Phys.*, 3rd April 1957, Vol. 148, No. 2, pp. 209–217.) Report on investigations of the influence of ultrasonic oscillations on the conductivity of single-crystal Se. The piezoelectric constant was determined by a dynamic as well as a quasi-static method.

537.311.33 : 546.24 **2449**

On the Structure of the Electron Spectrum in Lattices of the Tellurium Type.—Yu. A. Firsov. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1350–1367.) The general properties of the energy spectrum of an excess electron are studied by group-theory methods. The possibility of contact between the zone in definite directions in the **k**-space is considered and the effect of the spin-orbit interaction is examined. In semiconductors of this type there can be two kinds of carriers of the same sign.

537.311.33 : 546.26-1 : 538.21 **2450**

The Electric and Magnetic Properties of Graphite.—R. R. Haering & P. R. Wallace. (*Phys. Chem. Solids*, 1957, Vol. 3, Nos. 3/4, pp. 253–274.)

537.311.33 : [546.28+546.289 **2451**

The Manufacture of High-Purity Germanium and Silicon Crystals.—H. F. Mataré. (*Elektronische Rundschau*, Oct. 1957, Vol. 11, No. 10, pp. 293–296.) Outline of some techniques used, particularly those designed to overcome the greater difficulties inherent in the purification of Si.

537.311.33 : [546.28+546.289 **2452**

Tight-Bonding Calculation of Acceptor Energies in Germanium and Silicon.—R. G. Shulman. (*Phys. Chem. Solids*, April 1957, Vol. 2, No. 2, pp. 115–118.) "The differences of hole ionization energies among the group-III elements acting as acceptors in silicon indicate the necessity of considering the specific contribution of the acceptor atom. Energies of two electron bonds are calculated using Morse curves to approximate interatomic potentials. The additional binding energy increases sharply from gallium to indium in accord with experiment. Arguments are presented to show why varying the acceptor element is less important in germanium."

537.311.33 : 546.28 **2453**

Ionization Rates for Electrons and Holes in Silicon.—A. G. Chynoweth. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1537–1540.) The ionization rate for electrons is found to be higher than that for holes for the following ranges of field strength E: for holes, $(2\cdot5–6\cdot0) \times 10^5$ V/cm; for electrons, $(2\cdot0–5\cdot0) \times 10^5$ V/cm. The results suggest that the field dependence of the ionization rate for holes, and probably for electrons also, could be expressed by $a \exp(-b/E)$.

537.311.33 : 546.28 : 669.046.54/.55 **2454**

Zone Purification of Silicon.—E. A. Taft & F. H. Horn. (*J. electrochem. Soc.*, Feb. 1958, Vol. 105, No. 2, pp. 81–83.) "A procedure is described for the zone purification of high-purity Si using thin-walled quartz boats. Data are given that show that improved lifetime and more uniform electrical resistivity result from the zone purification of commercially available Si."

537.311.33 : 546.289 **2455**

The Low-Temperature Electrical Conductivity of *n*-Type Germanium.—S. H. Koenig & G. R. Gunther-Mohr. (*Phys. Chem. Solids*, 1957, Vol. 2, No. 4, pp. 268–283.) Samples of reasonably pure *n*-type Ge single crystals were investigated at temperatures between 1·8 and 5·1°K. The dominant conductivity mechanism is associated either with the conduction band or with an impurity conduction mechanism according to whether the temperature is greater or less than about 4°K. The mechanism of impurity conduction fits a model proposed by Mott (see 2088 of 1957). The conductivity of the conduction band deviates from Ohm's law at fields as low as 3% of the threshold field for the breakdown generally ascribed to impact ionization.

537.311.33 : 546.289 **2456**

Electrical Methods for Determining the Positions of Dislocation Regions in Germanium.—C. A. Hogarth & A. C. Baynham. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 647–653, plate.) In Ge crystals grown on a deformed seed, edge dislocations occur. This gives regions of high electrical conductivity in *n*- and *p*-type samples. The positions of dislocation walls determined from point-contact rectification measurements or by plotting potential against distance at constant current, agree with the positions located by etch-pit examination of the surfaces. Evidence of the *p*-type character of dislocations in *n*-type Ge is shown.

537.311.33 : 546.289 **2457**

The Fermi Level in Germanium at High Temperatures.—J. S. Blakemore. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 692–694.) Some results are given of calculations on the Fermi level in doped and intrinsic Ge from room temperature to 1 200°K.

537.311.33 : 546.289 **2458**

Precipitation of Nickel and Copper from Supersaturated Solutions in Germanium.—P. Penning. (*Philips Res. Rep.*, Feb. 1958, Vol. 13, No. 1, pp. 17–36.) During the anneal of Ge samples supersaturated with Ni, unidentified acceptors with a low activation energy are observed in addition to the acceptor levels arising from substitutional Ni atoms. A mechanism is suggested in which these acceptors are associated with vacancies left behind in the lattice when a substitutional Ni atom jumps to an interstitial site. The interstitial Ni atoms then diffuse rapidly to nuclei of precipitation. Copper gives rise to a similar behaviour. The presence of a high density of vacancies during the precipitation of one of these elements greatly facilitates the penetration of the other, permitting concentrations much larger than the saturation value. See also 3779 of 1956.

537.311.33 : 546.289 **2459**

Generation of Imperfections in Germanium Crystals by Thermal Strain.—P. Penning. (*Philips Res. Rep.*, Feb. 1958, Vol. 13, No. 1, pp. 79–97.) Thermal strain is induced in a crystal if it is heated inhomogeneously or quenched from a high temperature, the strain resulting in either plastic flow or cracking. The distribution of dislocations resulting from plastic flow shows marked characteristics which can be explained by assuming that plastic flow is proportional to the elastic strain induced by the nonuniform temperature. The sources and effects of these strains on crystals grown from a melt are discussed.

537.311.33 : 546.289 : 535.215 : 538.639 **2460**

On the Anisotropy of the Even (Transverse) Photomagnetic Effect in Germanium Single Crystals.—I. K. Kikoin & Yu. A. Bykovskiĭ. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Sept. 1957, Vol. 116, No. 3, pp. 381–384.) Experiments carried out on a round disk-like specimen of *n*-type Ge in a magnetic field with illumination normal to the field showed that the potential difference between opposite points on the disk varied with the angle of rotation of the specimen about the axis parallel to the field. The anisotropy of the transverse photomagnetic effect in a sample with its (111) axis perpendicular to its surface is shown graphically. No correlation was found between transverse and usual photomagnetic effects. See also 491 of 1957.

537.311.33 : 546.289 : 538.569.4 **2461**

Cyclotron Resonance over a Wide Temperature Range.—D. M. S. Bagguley, J. A. Powell & D. J. Taylor. (*Proc. phys. Soc.*, 1st Oct. 1957, Vol. 70, No. 454A, pp. 759–762.) Cyclotron resonance experiments on Ge doped with Au were carried out at temperatures in the range 4°K to 90°K. See also 1121 of 1957 (Dexter et al.).

537.311.33 : 546.289 : 538.6 **2462**

Direct-Transition Exciton and Fine Structure of the Magneto-absorption Spectrum in Germanium.—S. Zwerdling, L. M. Roth & B. Lax. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 2207–2209.) Report of the observation of exciton absorption just below the direct gap in germanium, together with the fine structure of the related oscillatory magneto-absorption. The exciton was also studied at various magnetic field strengths.

537.311.33 : 546.289 : 538.63 **2463**

Magnetoresistive Phenomena in *n*-Type Ge Semiconductors in Strong Magnetic Fields.—M. I. Klinger & P. I. Voronyuk. (*Zh. eksp. teor. Fiz.*, July 1957, Vol. 33, No. 7, pp. 77–87.) The equilibrium electron concentration, the Hall constant and resistivity are examined and their dependence on the magnetic field strength is discussed. The nature of the anisotropy of these quantities is determined by the anisotropy of the electron mass and by the number and mutual arrangement of the constant-energy ellipsoids.

537.311.33 : 546.46.284 **2464**

Semiconducting Properties of Mg_2Si Single Crystals.—R. G. Morris, R. D. Redin & G. C. Danielson. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1909–1915.) A report of resistivity and Hall-effect measurements on high-purity *n*-type and *p*-type Mg_2Si single crystals obtained from melts of the constituents. The *p*-type crystals were obtained by doping the melt with silver or copper. The room-temperature Hall mobilities were 406 cm²/V.sec for *n*-type and 56 cm²/V.sec for *p*-type material. Carrier concentrations in the saturation region were as low as 8×10^{16} cm^{-3} for *n*-type and 4×10^{17} cm^{-3} for *p*-type samples.

537.311.33 : 546.46.289 **2465**

Semiconducting Properties of Mg_2Ge Single Crystals.—R. D. Redin, R. G. Morris & G. C. Danielson. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1916–1920.) A report of resistivity and Hall-effect measurements on single crystals of Mg_2Ge obtained from melts of the constituents. Undoped crystals which were *n* type had saturated impurity concentrations as low as 3×10^{15} cm^{-3}; silver-doped crystals were *p* type and had saturated impurity carrier concentrations roughly proportional to the amount of added silver. Room-temperature Hall mobilities were 280 cm²/V.sec for electrons and 110 cm²/V.sec for holes.

537.311.33 : 546.682.86 **2466**

Properties of *p*-Type Indium Antimonide: Part 1—Electrical Properties.—C. Hilsum & R. Barrie. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 676–685.) The variation of Hall coefficient and resistivity with applied magnetic field is used to calculate the dependence of electron and hole mobilities on carrier concentration. The decrease of electron mobility with acceptor concentration is explained in terms of polar scattering and electron-hole scattering.

537.311.33 : 546.682.86 : 537.312.9 **2467**

Piezoresistance Constants of *P*-Type InSb.—A. J. Tuzzolino. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1980–1987.) The change of resistance in uniaxial tension was measured for several single-crystal specimens of *p*-type InSb over the range 77°–350°K. The piezoresistance coefficients were found to depend on impurity concentrations. The elasto-Hall constant was measured at 77°K for the purest specimens and the experimental results indicate that, in the extrinsic range, the large piezoresistance is primarily due to stress-induced changes in the tensor mobility of the holes. It is concluded that the valence band extrema occur at, or very near, $\mathbf{K} = 0$ in the Brillouin zone with energy surfaces similar to those of Si and Ge.

537.311.33 : 546.682.86 : 537.32 **2468**

Electron Scattering in InSb.—H. Ehrenreich. (*Phys. Chem. Solids*, April 1957, Vol. 2, No. 2, pp. 131–149.) The effects of electron scattering by acoustical and optical phonons as well as electron-hole scattering on the mobility and thermoelectric power of intrinsic InSb are considered using the conduction band theory of Kane (*ibid.*, Jan. 1957, Vol. 1, No. 4, pp. 249–261). Good agreement is found between theory and experiment for both mobility and thermoelectric power.

537.311.33 : 546.682.86 : 537.32 **2469**

On the Scattering of Electrons by Optical Modes in Indium Antimonide.—M. Rodot. (*C. R. Acad. Sci., Paris*, 23rd Sept. 1957, Vol. 245, No. 13, pp. 1051–1054.) Measurements of the magneto-thermoelectric effect and the Nernst effect in specimens of *n*-type InSb at various temperatures gave results in accordance with the theory of polar scattering. See also 3772 of 1956.

537.311.33 : 546.682.86 : 539.23 **2470**

Resistivity and Hall Effect of Thin Films of Indium Antimonide.—J. Launey & A. Colombani. (*C. R. Acad. Sci., Paris*, 16th Sept. 1957, Vol. 245, No. 12, pp. 1009–1011.) Films were reheated and then cooled to room temperature before tests were made. Hall effect was found to increase with reheat temperature and to be strictly proportional to current intensity. Consistent results were not obtained if the reheat temperature was above 470°C. Resistivity decreased rapidly with increase in film thickness.

537.311.33 : 546.682.86 : 539.23 **2471**

Hall Effect and Magnetoresistance of Thin Films of Indium Antimonide.—J. Launey. (*C. R. Acad. Sci., Paris*, 30th Sept. 1957, Vol. 245, No. 14, pp. 1122–1124.) See also 2470 above.

537.311.33 : [546.873.241 + 546.863.241 + 546.193.241 **2472**

Preparation and some Physical Properties of Bi_2Te_3, Sb_2Te_3 and As_2Te_3.—T. C. Harman, B. Paris, S. E. Miller & H. L. Goering. (*Phys. Chem. Solids*, 1957, Vol. 2, No. 3, pp. 181–190.) The preparation of the above intermetallic compounds from purified elements by several techniques is discussed. Electrical and thermal properties are presented as functions of temperature and impurity concentration. The variation of carrier mobility with temperature is approximately $T^{-5/2}$ for As_2Te_3 and Bi_2Te_3.

537.311.33 : 546.873.241 : 537.323 **2473**

The Electrical Conductivity and Thermoelectric Power of Bismuth Telluride.—H. J. Goldsmid. (*Proc. phys. Soc.*, 1st April 1958, Vol. 71, No. 460, pp. 633–646.) The electrical conductivity and thermoelectric power of Bi_2Te_3 are measured over a range of temperatures and for *n*-type and *p*-type samples. This gives the fundamental semiconductor parameters. Samples show extrinsic conduction and partial degeneration over parts of the temperature range necessitating the use of Fermi-Dirac statistics in interpretation of the results.

537.311.33 : 621.314.63 **2474**

Electrical Breakdown in *p-n* Junctions.—A. G. Chynoweth. (*Bell Lab. Rec.*, Feb. 1958, Vol. 36, No. 2, pp. 47–51.) The sudden breakdown which occurs when an increasing reverse voltage is applied to a *p-n* junction is explained by the creation of free electrons and holes in an avalanche process.

538.22 **2475**

Thermodynamic Theory of 'Weak' Ferromagnetism in Antiferromagnetic Substances.—I. E. Dzyaloshinskiĭ. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1547–1562.) This type of ferromagnetism is attributed to relativistic spin-lattice interaction and magnetic dipole interaction. The dependence of the properties of these substances on the magnetic symmetry of the crystal is examined and the effect of an external magnetic field is also considered.

538.22 **2476**

Solid Solutions between Ferromagnetic and Antiferromagnetic Compounds with NiAs Structure.—F. K. Lotgering & E. W. Gorter. (*Phys. Chem. Solids*, 1957, Vol. 3, Nos. 3/4, pp. 238–249.) The aim of the investigation is to see whether continuous series of mixed crystals between a ferromagnetic and an antiferromagnetic binary compound with NiAs structure exist, and to study the magnetic properties. Several solid solutions were prepared. Saturation magnetizations were measured between 20°K and the Curie temperatures, and susceptibilities between Curie temperatures and 850°–1 100°K. Results are discussed.

538.22 : 537.3 **2477**

Electrical Properties of $FeTiO_3$-Fe_2O_3 Solid Solution Series.—Y. Ishikawa. (*J. phys. Soc. Japan*, Jan. 1958, Vol. 13, No. 1, pp. 37–42.) The d.c. electrical conductivity and Seebeck voltage of the synthesized solid solution $x\text{FeTiO}_3.(1-x)\,\text{Fe}_2\text{O}_3$ of ilmenite (FeTiO_3) and hematite (Fe_2O_3) have been measured for a composition range of $1 \geqslant x \geqslant 0{\cdot}33$, over a range of temperature. Both ordered (ferrimagnetic) and disordered (antiferromagnetic) specimens with the same composition were investigated. See also 1199 of April (Ishikawa & Akimoto).

538.221 **2478**

Temperature Dependence of Ferromagnetic Anisotropy.—W. J. Carr, Jr. (*Phys. Rev.*, 15th March 1958, Vol. 109, No. 6, pp. 1971–1976.) It is shown that Zener's result for Fe may be derived from molecular field theory. In Co a satisfactory agreement with experiment is obtained by using Zener's results together with the postulate that the intrinsic anisotropy varies with the thermal expansion in the manner recently calculated (1389 of May). For Ni the temperature dependence of K_1 seems to require, in addition to the tenth power of the magnetization, a multiplicative factor that is linear with temperature.

538.221 **2479**

Contribution to the Clarification of the Magnetic After-Effect due to Electron Diffusion in Ferrites.—A. v. Kienlin. (*Z. angew. Phys.*, May 1957, Vol. 9, No. 5, pp. 245–250.) On the basis of measurements an interpretation is derived of the Richter effect observed in stoichiometric Ni-Zn ferrites. See e.g. 2041 of 1953 (Feldtkeller & Sorger).

538.221 : 537.533.73 **2480**

Investigation of Ferromagnetic Substances by Electron Diffraction.—S. Yamaguchi. (*Z. Metallkde*, Jan. 1958, Vol. 49, No. 1, pp. 52–54.) An electron beam penetrating a ferromagnetic substance is polarized. This effect can be used to obtain a crystallographic and a magnetochemical analysis of the material.

538.221 : [621.318.124+621.318.134 **2481**

Convention on Ferrites.—(*Proc. Instn elect. Engrs*, Part B, 1957, Vol. 104, Supplement No. 7, pp. 399–570.) The following papers were included among those read at the I.E.E. Convention held in London 29th October–2nd November 1956. See also 1479 of May.

Square-Loop Materials:

(*a*) The Influence of Chemistry on B/H Loop Shape, Coercivity and Flux-Reversal Time in Ferrites.—J. B. Goodenough (pp. 400–411).

(*b*) Ordering in Cobalt-Ferrous Ferrites. —H. P. J. Wijn, H. van der Heide & J. F. Fast (pp. 412–417).

(*c*) Square-Loop Ferrites obtained by Magnetic Annealing of New Compositions. —E. W. Gorter & C. J. Esveldt (pp. 418–421).

(*d*) Pulse-Response Properties of Rectangular-Loop Ferrites.—H. P. J. Wijn & H. van der Heide (pp. 422–427).

(*e*) Ferrites with Constricted Loops and Thermal Magnetic Treatment.—O. Eckert (pp. 428–432).

Discussion (pp. 433–435).

Square Loop Applications:

(*f*) A Magnetic-Core Matrix Store with Direct Selection using a Magnetic-Core Switch Matrix.—W. Renwick (pp. 436–444).

(*g*) The Magnetic Cell: A New Circuit Element.—D. S. Ridler & R. Grimmond (pp. 445–456).

(*h*) A Matrix Store for Data Rate Conversion.—C. J. Quartly, A. L. Cain & R. W. Clarke (pp. 457–463).

(*i*) A Ferrite-Core Switch Matrix for Magnetic-Recording-Head Selection.—D. W. Willis (pp. 464–469).

(*j*) A 1 000-Channel Neutron Velocity Spectrometer using Ferrite Data Storage.—F. H. Wells & J. G. Page (pp. 470–480).

(*k*) A Digital Computer based on Magnetic Circuits.—B. Bambrough (pp. 485–490).

(*l*) Some Applications of Square-Loop Ferrite Cores in Telecommunication Switching Circuits.—W. Six & R. A. Koolhof (pp. 491–501).

(*m*) The Application of Square-Loop Ferrite Material in Data-Processing Systems.—E. P. G. Wright (pp. 502–511).

(*n*) A Core/Transistor Logical Element.—D. Eldridge & P. F. Dorey (pp. 512–522).

Discussion (pp. 481–482, 523–525).

Radio & Television Applications:

(*o*) Some Current and Future Applications for Ferrite Components in Radio and Television Receivers.—V. A. Jones & G. Campbell (pp. 526–533).

(*p*) The Application of Soft Magnetic Ferrites in Television Receiver Line-Scanning Circuits.—R. H. C. Morgan (pp. 534–537).

Discussion (pp. 538–540).

Carrier-Frequency Applications:

(*q*) Ferrites Applied to Carrier-Frequency Equipment.—W. F. Glover & S. E. Buckley (pp. 541–549).

(*r*) Ferrites for Magnetic and Piezomagnetic Filter Elements with Temperature-Independent Permeability and Elasticity.—C. M. van der Burgt (pp. 550–557). See also 2232 and 3566 of 1957.

(*s*) Practical Applications and Measurements on Ferrite Electromechanical Transducers.—G. Bradfield (pp. 558–563).

(*t*) Examples of Electric Wave Filters using Ferrite Inductors.—W. Saraga (pp. 564–567). See also 2047 of 1957.

Discussion (pp. 568–570).

538.221 : 621.318.124 **2482**

Uniaxial Anisotropy in Iron-Cobalt Ferrites.—S. Iida, H. Sekizawa & Y. Aiyama. (*J. phys. Soc. Japan*, Jan. 1958, Vol. 13, No. 1, pp. 58–71.) The uniaxial anisotropy induced by magnetic annealing in both single-crystal and polycrystal line specimens has been studied experimentally. The magnitudes of the anisotropy depend essentially on the degree of oxidation of the specimens and become almost zero after deoxidizing heat treatments.

538.221 : 621.318.134 **2483**

Modern Applications of Ferrites.—F. Mayer. (*Rev. gén. Élect.*, June 1957, Vol. 66, No. 6, pp. 333–339.) A review of the properties and applications of spinel-type ferrites.

538.221 : 621.318.134 : 538.569.4 **2484**

Determination of the Gyromagnetic Ratio and the Magnetic Resonance Damping Coefficient of Ferrites.—H. G. Beljers. (*Philips Res. Rep.*, Feb. 1958, Vol. 13, No. 1, pp. 10–16.) Methods of deriving the g factor and damping coefficient from the results of resonance measurements on polycrystalline ferrite spheres in a microwave cavity with a linearly polarized magnetic field are described. Results for three specimens are tabulated.

538.221 : 621.318.2 **2485**

The Development of Alnico Magnets containing Copper and/or Titanium up to their Present Level of Quality.—W. Zumbusch. (*Z. Metallkde*, Jan. 1958, Vol. 49, No. 1, pp. 1–8.)

538.569.4.029.64 **2486**

Microwave Absorption in Methyl Halides.—Krishnaji & G. P. Srivastava. (*Phys. Rev.*, 1st March 1958, Vol. 109, No. 5, pp. 1560–1563.) It is confirmed that the absorption of 3-cm waves in methyl chloride and methyl bromide is entirely due to the inversion transition at zero wave number. In the J band the additional absorption observed is assigned to the resultant effect of various rotational transitions of the molecule.

538.632 : 546.3-1'48'46 **2487**

Variation of the Hall Coefficient of Mg-Cd Superlattice Alloys with Degree of Order.—K. Yonemitsu & T. Sato. (*J. phys. Soc. Japan*, Jan. 1958, Vol. 13, No. 1, pp. 15–22.) The temperature dependence of the Hall coefficient has been measured for Mg_3Cd, MgCd and MgCd_3. At the order-disorder transformation temperature the Hall coefficient and electrical resistivity are discontinuous in opposite directions. It is considered that the change of Brillouin zone structure on transformation is responsible for the discontinuities.

538.632 : 546.59 **2488**

Hall Effect and Susceptibility of Gold.—N. E. Alekseevskiĭ & Yu. P. Gaĭdukov. (*Zh. eksp. teor. Fiz.*, June 1957, Vol. 32, No. 6, pp. 1589–1591.) Tests were made on two samples of Au with differing resistivity/temperature characteristics. See also 3216 of 1957.

MEASUREMENTS AND TEST GEAR

621.3.018.41 (083.74) : 621.373.42.029.64 **2489**

S.H.F. Frequency Standard uses Double Conversion.—M. C. Thompson, Jr, M. J. Vetter & D. M. Waters. (*Electronics*, 11th April 1958, Vol. 31, No. 15, pp. 100–101.) A harmonic from a stable crystal oscillator is mixed with a klystron output. The beat frequency is compared with another crystal-controlled frequency and a correction voltage is applied to the klystron. See also 2874 of 1955 (Peter & Strandberg).

621.3.018.41 (083.74) : 621.373.421.14 **2490**

An Atomic Reference Oscillator.—M. P. G. Capelli. (*Brit. Commun. Electronics*, Feb. 1958, Vol. 5, No. 2, pp. 100–102.)

"One of the spectrum lines of the caesium atom is used as the final frequency reference of a highly stable and accurate frequency source."

621.3.018.41 (083.74) : 621.396.11.029.45 **2491**

Phase Variations of 16-kc/s Transmissions from Rugby as Received in New Zealand.—Crombie, Allan & Newman. (See 2524.)

621.314.7.001.4 (083.74) **2492**

I.R.E. Standards on Solid-State Devices: Methods of Testing Point-Contact Transistors for Large-Signal Applications, 1958.—(*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 878–888.) Standard 58 I.R.E. 28. S1.

621.317.2 : 621.373.4 **2493**

A Precision Microwave Signal Generator.—F. W. Cook. (*A.T.E. J.*, Jan. 1958, Vol. 14, No. 1, pp. 2–10.) Description and circuit details of an instrument covering the frequency range 580–1 220 Mc/s with automatic control of output level and provision for c.w. or square-wave, pulse or sine-wave modulated output.

621.317.3 : 537.311.33 : 537.323 **2494**

Apparatus for Measurement of Thermal E.M.F. in Semiconductors.—J. C. Brice & H. C. Wright. (*J. sci. Instrum.*, April 1958, Vol. 35, No. 4, pp. 146–147.)

621.317.3 : 621.314.7 **2495**

An Investigation of the Current Gain of Transistors at Frequencies up to 105 Mc/s.—F. J. Hyde & R. W. Smith. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 221–228.) Apparatus is described for direct measurement of short-circuit current gain. Results obtained with commercial alloy-junction and surface-barrier transistors are presented, with corrections to give the internal diffusion-current gain, and the effect of stray capacitances on the measurements are considered. The cut-off frequency of the internal current gain is compared with values obtained indirectly from other measurements. The behaviour of alloy-junction transistors agrees closely with existing one-dimensional diffusion theory while surface-barrier transistors show less agreement.

621.317.3 : 621.314.7 : 621.375.1.024 **2496**

Use of Operational Amplifiers in the Measurement of Transistor Parameters.—W. C. Hazel. (*Rev. sci. Instrum.*, March 1958, Vol. 29, No. 3, pp. 235–237.) Methods of measuring the resistance parameters and the current amplification factor are described. Schematic circuit diagrams are included.

621.317.3 : 621.396.67 : 621.397.62 **2497**

Measuring TV Aerial Performance.—Strafford. (See 2297.)

621.317.3 : 621.396.822 : 621.396.41 **2498**

The White-Noise Method of Measuring Crosstalk and Noise Interference in Multichannel Telephone Link Systems.—J. F. Golding. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 349–351.) Description of the methods with details of a commercially available test set.

621.317.32 **2499**

Measurement of Peak Values of Periodic High Voltages in a Steady State.—J. Lagasse, R. Lacoste & G. Giralt. (*Rev. gén. Élect.*, June 1957, Vol. 66, No. 6, pp. 307–324.) The design and construction of a standard variable capacitor is described and the merits of different types of rectifier are discussed. See 1851 of 1957 (Lagasse & Giralt).

621.317.33 : 538.632 **2500**

A Method of Measuring Specific Resistivity and Hall Effect of Disks of Arbitrary Shape.—L. J. van der Pauw. (*Philips Res. Rep.*, Feb. 1958, Vol. 13, No. 1, pp. 1–9.) The method is based on a theorem applicable to a flat specimen of arbitrary shape if the contacts are sufficiently small and located at its circumference; the disk must be free of holes.

621.317.341 : 621.374.015.7 **2501**

Sensitive Measurements of Pulse-Amplifier Gain.—K. A. McCollom, D. R. deBoisblanc & J. B. Thompson. (*Nucleonics*, Jan. 1958, Vol. 16, No. 1, pp. 74–76, 78.) The design of an amplifier comparator is described. Gain stability can be checked to within 0·01%, and amplitude stability of two pulse generators can be compared to within 0·02%.

621.317.35 : 621.391 **2502**

The Measurement of Power Spectra from the Point of View of Communications Engineering: Parts 1 & 2.—R. B. Blackman & J. W. Tukey. (*Bell Syst. tech. J.*, Jan. & March 1958, Vol. 37, Nos. 1 & 2, pp. 185–282 & 485–569.) A comprehensive mathematical and statistical study is presented, with details of analysis and planning for measurement. The analysis is then reconsidered in greater detail and from alternative points of view. Over 50 references.

621.317.361 : 538.569.4 : 539.1.08 **2503**

Measurement and Control of Microwave Frequencies by Lower Radio Frequencies.—R. C. Mackey & W. D. Hershberger. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 64–68. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.)

621.317.411 **2504**

The Determination of the Permeability of Ferromagnetic Materials from the Thermal Noise Voltage of Coils.—W. Nonnenmacher & L. Schweizer. (*Z. angew. Phys.*, May 1957, Vol. 9, No. 5, pp. 239–245.) The complex permeability is measured as a function of frequency without the use of an externally applied alternating field. The thermal fluctuations in the ferromagnetic core material give rise to a noise voltage across the test coil. Results obtained agree well with those from conventional bridge measurements at low field strengths.

621.317.42 : 621.395.625.3 **2505**

The Determination of the Magnetization of Magnetic Recording Tape.—O. Schmidbauer. (*Elektronische Rundschau*, Oct. 1957, Vol. 11, No. 10, pp. 302–305.) The magnetization of tape is best defined on the basis of short-circuit flux measurements; a suitable method is described. The determination of transverse magnetization, where it is undesirable or where it is used for synchronizing purposes, and the calibration of heads used in such measurements are also discussed. A specially designed electronic fluxmeter is briefly detailed.

621.317.44 **2506**

Adaptation of the Iliovici Permeameter to the Investigation of Hypermagnetic Materials.—L. Garde. (*Rev. gén. Élect.*, July 1957, Vol. 66, No. 7, pp. 375–379.) To reduce the error of the instrument which can be excessive in measurements on samples of high permeability, the second yoke is constructed of mumetal.

621.317.715 : 621.375.024 **2507**

Device for Matching a Galvanometer or D.C. Indicator to its Associated Circuit.—T. M. Dauphinee. (*Rev. sci. Instrum.*, March 1958, Vol. 29, No. 3, pp. 240–241.) "Impedance matching is achieved by use of a chopper and transformer circuit which permits near optimum use of the available power."

621.317.727 : 621.316.93 : 621.314.63 **2508**

Silicon Diodes as Protective Meter Shunts.—A. S. Penfold & E. L. Garwin. (*Rev. sci. Instrum.*, March 1958, Vol. 29, No. 3, pp. 252–253.) Relies on the very small current through the diode for forward voltages less than 0·5 V.

621.317.733 **2509**

Impedance Bridges.—J. F. Golding. (*Brit. Commun. Electronics*, Feb. 1958, Vol. 5, No. 2, pp. 104–109.) "After a short discussion of impedance, various types of impedance bridge and their applications are described. A table gives brief specifications of representative bridges available in Britain."

621.317.755 **2510**

Modern Oscilloscope Practice.—(*Electronic Radio Engr*, June 1958, Vol. 35, No. 6, pp. 212–225.) A review of instruments manufactured in the U.K., with simplified circuits of various parts, showing how particular facilities are provided. Oscilloscope cameras are discussed briefly.

621.317.755 : 621.385.832 **2511**

Broadband Oscilloscope Tube.—D. J. Brangaccio, A. F. Dietrich & J. W. Sullivan. (*Bell Syst. tech. J.*, March 1958, Vol. 37, No. 2, pp. 447–460.) The design and operation of a special oscilloscope tube is described. Travelling-wave-valve principles are applied to the design of a helix-type vertical deflection system to give a flat bandwidth characteristic over 600 Mc/s. Novel features are the particular deflection system and a trace on the screen which is readable without other optical means. The construction is similar to commercial tubes except for closer tolerance in electrode alignment. Repetitive pulses a few millimicroseconds in width can be viewed directly.

621.317.784.029.63 **2512**

U.H.F. Power Meter for Operation in the 2 000-Mc/s Communication Band.—J. K. Murray. (*Electronic Engng*, May

1958, Vol. 30, No. 363, pp. 345–348.) Description of the design of a direct-reading power meter for the frequency range 1 700–2 300 Mc/s.

621.317.794 : 621.319.4 **2513**

Notes on Capacitance Bolometers.—H. J. Albrecht. (*Proc. Instn Radio Engrs, Aust.*, April 1957, Vol. 18, No. 4, pp. 128–129.) Note on the principles and design of an instrument comprising temperature-sensitive capacitors, with $BaTiO_3$ or $SrTiO_3$ dielectric, in two tuned circuits.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.365.5 **2514**

Induction Heating of Ferromagnetic Samples at Stepped Frequencies.—J. Minssieux. (*Rev. gén. Élect.*, May 1957, Vol. 66, No. 5, pp. 239–245.) In the case of steel rods, for example, 50-c/s a.c. may be used for heating up to 700°C, and 1 000-c/s a.c. for heating from 700°C to 1 250°C.

621.38 : 62 **2515**

Electronics in Heavy Industry.—G. M. Ettinger. (*Elect. Rev., Lond.*, 3rd Jan. 1958, Vol. 162, No. 1, pp. 30–35.) Survey of applications in measurement, control, data handling, and processing.

621.384.612 **2516**

About Damping and Antidamping Betatron Oscillations, taking into Account Non-Adiabatic Properties of the Radiation.—A. N. Matveev. (*Nuovo Cim.*, 1st Dec. 1957, Vol. 6, No. 6, pp. 1296–1304. In English.) An analysis of betatron oscillations in high-energy electron synchrotrons shows that damping and antidamping effects would only occur if radiation were an adiabatic process.

621.384.612 **2517**

Electron Motion in Synchrotrons in the Presence of Radiation.—A. N. Matveev. (*Nuovo Cim.*, 1st Dec. 1957, Vol. 6, No. 6, pp. 1305–1317. In English.) A stochastic theory of betatron and phase oscillations induced by radiation fluctuations is presented. See also 240 of 1957.

621.387.424 **2518**

Influence of the Effective Length and Diameter of the Cylinder on the Plateau Characteristic of Externally Graphited Glass-Walled Geiger-Müller Counters.—D. Blanc. (*Nuovo Cim.*, 1st Oct. 1957, Vol. 6, No. 4, pp. 974–976. In French.)

PROPAGATION OF WAVES

621.396.11 **2519**

On the Theory of Propagation of Electromagnetic Waves Along a Curved Surface.—J. R. Wait. (*Canad. J. Phys.*, Jan. 1958, Vol. 36, No. 1, pp. 9–17.) The case of vertically polarized waves propagated across a discontinuity in curvature and electrical properties is considered. An approximate expression is derived for the mutual impedance between two aerials on either side of the boundary by using the principle of stationary phase and the concept of surface impedance. The first-order effects of the boundary changes are shown to be additive corrections to the mutual impedance for a homogeneous surface.

621.396.11 **2520**

Some Methods for Studying Wave Propagation in a Uniform Magneto-ionic Medium.—V. A. Bailey. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 118–125.) The Appleton-Hartree expression for complex refractive index can be simplified by using the approximations:

$$(1 + s^2)^{\frac{1}{2}} = (4 + 3s^2)/(4 + s^2)$$
$$\text{or} = s\,(4s^2 + 3)/(4s^2 + 1)$$

The resulting expressions have errors of less than 1% except under limited sets of conditions.

621.396.11 : 551.510.535 **2521**

The Distance Attenuation of Radio Waves Reflected at Vertical Incidence from the Ionosphere.—J. D. Whitehead. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 150–152.) "It is shown that the assumption that the amplitude of radio waves propagated from a point source falls off inversely as the group path P' leads to a value of apparent reflection coefficient of the ionosphere which may be seriously in error, and in particular to a large error in the measurement of the electronic collision frequency in the F region."

621.396.11.029.4 (204) **2522**

On Communication in the Sea by V.L.F. Electromagnetic Waves.—K. Furutsu. (*J. Radio Res. Labs, Japan*, Jan. 1958, Vol. 5, No. 19, pp. 19–33.) A theoretical analysis of the various modes of propagation when either transmitter or receiver or both are immersed shows that a horizontal magnetic bar is the most suitable radiator for use under the sea. Numerical values of input impedance and induced voltage are given for such a radiator under transmitting and receiving conditions. Field-strength curves are presented for various conditions of communication from sea to atmosphere or in the sea.

621.396.11.029.45 **2523**

Differences between the East-West and West-East Propagation of V.L.F. Signals over Long Distances.—D. D. Crombie. (*J. atmos. terr. Phys.*, 1958, Vol. 12, Nos. 2/3, pp. 110–117.) Early suggestions that eastward-travelling waves are less attenuated than those of the opposite sense are supported by recent measurements. Such an effect might be due to the propagation of TM modes in the presence of free electrons in the waveguide formed by the ionosphere and the earth. Propagation along the earth's magnetic field (N–S) would obey the reciprocity law while in E-W directions it would not.

621.396.11.029.45 : 621.3.018.41 (083·74) **2524**

Phase Variations of 16-kc/s Transmissions from Rugby as Received in New Zealand.—D. D. Crombie, A. H. Allan & M. Newman. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 301–304.) It is suggested that the measured diurnal phase changes over the 19 000-km path can be explained to a first approximation by a model in which both earth and ionosphere are considered as perfectly conducting sheets whose separation varies from night to day. See also 1482 of 1956 (Allan et al.).

621.396.11.029.6 : 621.396.81 **2525**

Approximate Methods for the Evaluation of the Field Strength of Ultra Short Radio Waves taking Account of the Influence of the Local Ground Profile.—A. I. Kalinin. (*Radiotekhnika, Mosk.*, April 1957, Vol. 12, No. 4, pp. 13–23.) Interference formulae are used for field strength calculations for unobstructed paths. Where screening occurs the obstacles are represented approximately by spheres of a radius depending on the nature of the obstacle. Formulae for attenuation factor, angle of diffraction and refraction coefficient are derived.

621.396.11.029.62 : 551.510.535 **2526**

Off-Path Propagation at V.H.F.—V. C. Pineo. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 922.) Reception of signals propagated by ionospheric forward scatter at two points about $6\frac{1}{2}°$ to each side of the main transmitted beam shows the influence of meteoric ionization.

621.396.11.029.63/.64 **2527**

Microwave Line-of-Sight Propagation.—M. W. Gough. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 237–247.) A survey of some fundamental investigations made at frequencies above 300 Mc/s, mainly since 1939, in Europe and the U.S.A. Effects of ground reflection are discussed together with the use of passive reflectors and screens. Fading mechanisms and diurnal and seasonal variations in fading due to weather conditions are considered. 45 references.

621.396.11.029.63/.64 : 551.510.52 **2528**

Tropospheric Scatter Propagation.—G. Millington. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 248–252.) "The physical mechanisms of long-distance propagation are reviewed with special reference to forward scattering. Recent research in the United Kingdom is described on the nature of the signal variations and their relation to meteorological conditions. The use of scatter links for wide-band systems is discussed, including the problems of aircraft reflections and mutual interference due to frequency sharing."

621.396.11.029.63/.64 : 551.510.52 **2529**

Tropospheric Scatter Communication.—G. L. Grisdale. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 272–275.) Tropospheric scatter can give reliable communication where terrain limits multi-hop microwave links, but the rapid fading characteristic and limited signal strength over the longer distances impair quality. The principal features and application of this method of propagation are considered.

RECEPTION

621.376.232 : 621.396.822 **2530**

The Behaviour of Phase-Sensitive Detectors.—M. B. Palma-Vittorelli, M. U. Palma & D. Palumbo. (*Nuovo Cim.*, 1st Nov. 1957, Vol. 6, No. 5, pp. 1211–1220. In English.) An analysis is made of the output of different types of phase detector for any given input, and expressions are given for the frequencies present in the output.

621.396.62 : 629.19 **2531**

Microlock.—H. L. Richter, Jr. (*QST*, Dec. 1957, Vol. 41, No. 12, pp. 20–24.) Details are given of a receiver for tracking and receiving telemetry signals from earth satellites, using a phase-locked loop system with very narrow bandwidth, and designed for the reception of signals below the noise threshold of conventional receivers.

621.396.621.54 : 621.372.632 : 621.373.51 **2532**

Crystal Converter for Tropo-scatter Receivers.—P. Gruber. (*Electronics*, 11th April 1958, Vol. 31, No. 15, pp. 78–82.) A noise-figure improvement of 0·5–1·5 dB has been obtained at frequencies of 1–2 kMc/s by replacing a r.f. amplifier by a Type-1N21E silicon diode in a suitably designed coaxial mixer.

621.396.8 **2533**

Carrier-to-Noise Statistics for Various Carrier and Interference Characteristics.—K. K. Clarke & J. Cohn. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 889–895.) "Techniques are presented for the calculation of the statistical properties of the resultant carrier-to-noise ratios of systems subject to both additive and multiplicative noise. The cases considered are those in which the desired signal is either steady or exhibits Rayleigh fading while the interference consists of receiver noise plus an interfering signal which may be steady, Rayleigh fading, Gaussian fading, or Rayleigh fast fading with slow Gaussian fading of the median of the Rayleigh distribution. The results of the indicated calculations are presented graphically."

621.396.812.3 : 621.396.666 **2534**

Theoretical Diversity Improvement in Frequency-Shift Keying.—J. N. Pierce. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 903–910.) If prior information about signal amplitude and phase is not available, square-law combination is the optimum system; if the information is available, coherent combination and detection are the optimum, giving a 3-dB improvement in the error probability at low probabilities.

621.396.822 : 551.594.6 **2535**

Atmospheric Radio Noise at Frequencies between 10 kc/s and 30 kc/s.—Harwood. (See 2422.)

621.396.822 : [621.314.632-71 + 621.396.67-71 **2536**

Cooling of Microwave Crystal Mixers and Antennas.—G. C. Messenger. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 62–63. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.)

621.396.823 **2537**

Contribution to the Prediction of Radio Interference Capacity of Very-High-Voltage Lines.—J. Meyer de Stadelhofen & W. Walter. (*Tech. Mitt. PTT*, 1st Nov. 1957, Vol. 35, No. 11, pp. 456–479. In French & German.) The characteristics of corona effects and the reduction of corona by the use of stranded conductors are described. Details are given of measurements carried out near power lines and on insulators and lines in a laboratory to determine the attenuation of the interference field perpendicular to and along the lines, and to assess the usefulness of laboratory tests in predicting the level of interference due to power lines under various atmospheric conditions.

STATIONS AND COMMUNICATION SYSTEMS

621.376 : 621.396.5 **2538**

Selection of Modulation for Speech Communication.—G. J. Kelley. (*Electronics*, 28th March 1958, Vol. 31, No. 13, pp. 56–58.) The type used (a.m., f.m., s.s.b., or d.s.b. suppressed-carrier) can be chosen by a logical evaluation. The factors compared for particular applications include signal-noise performance, distortion and stability.

621.391 **2539**

Error Rates in Data Transmission.—S. Reiger. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 919–920.) Error rate reduction using correcting codes and binary signalling is compared with the gain obtained using other forms of signalling such as quantized f.m.

621.391 : 621.3.018.75 **2540**

A Statistical Description of Coincidences among Random Pulse Trains.—S. Stein & D. Johansen. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 827–830.) The derivation of the statistical distribution of the coincidence durations may be reduced to the problem of finding the probability of occurrence of a coincidence at any instant between trains of randomly occurring pulses with a known pulse length distribution.

621.391 : 621.3.018.78 **2541**

Characteristics of Signal Transmission and Distortion.—R. Possenti. (*Note Recensioni Notiz.*, Sept./Oct. 1957, Vol. 6, No. 5, pp. 639–667.) The concepts of group velocity and phase distortion are discussed and a method of evaluating signal distortion is derived.

621.391 : 621.397.2 **2542**

Statistical Encoding for Text and Picture Communication.—W. S. Michel. (*Commun. & Electronics*, March 1958, No. 35, pp. 33–36.) An investigation of the channel capacity required for various types of black and white copy based on one-dimensional differential co-ordinates with statistically matched encoding. A binary digital transmission channel is assumed.

621.396.2.029.631.64 **2543**

A Survey of Microwave Radio Communication.—W. J. Bray. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 226–236.) "This article is a broad survey of line-of-sight and tropospheric forward-scatter radio communication systems operating in the frequency range from about 1 000 to 10 000 Mc/s, for the transmission of multichannel telephony and television signals. The applications of such systems, the transmission requirements, the factors determining the choice of frequency, and the basic equipment techniques are described. Possible future developments are also considered."

621.396.32 : 621.396.812.3 **2544**

Measurement of Fading Margin in V.H.F. Radio Teletype Circuits.—K. Kawakami & H. Akima. (*J. Radio Res. Labs, Japan*, Jan. 1958, Vol. 5, No. 19, pp. 1–18.) 'Fading margin' is defined as the ratio of signal strength with fading to that without fading for the same signal/noise ratio. A method of measuring this quantity is described and results obtained with three different teletype v.h.f. circuits are presented. It is concluded that knowledge of the quantity facilitates the design of communication circuits and enables the noise and fading susceptibility of various systems to be compared.

621.396.41 **2545**

Wide-Band Ultra-High-Frequency Over-the-Horizon Equipment.—R. A. Felsenheld, H. Havstad, J. L. Jatlow, D. J. LeVine & L. Pollack. (*Commun. & Electronics*, March 1958, No. 35, pp. 86–93.) A description of wide-band equipment for use in toll-quality multichannel telephone or television circuits in the frequency range 680–900 Mc/s. The receivers can be used for dual or quadruple diversity by combining the signals at the 70-Mc/s intermediate frequency.

621.396.41 : 621.376.3 : 621.396.813 **2546**

Distortion in Frequency-Division Multiplex F.M. Systems due to an Interfering Carrier.—R. G. Medhurst, E. M. Hicks & W. Grossett. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 282–292.) It is found that the distortion generated in a particular channel is accurately proportional to the relative level of the interfering carrier. Thus it is possible to plot curves of distortion in the worst channel against frequency separation of wanted and unwanted carriers; a permissible level of interfering carrier may be deduced. Numerical results are shown for various carrier frequency separations that arise under the C.C.I.R. frequency plan for 6 r.f. channels each carrying 600 speech channels. Provided the frequency separation is less than 9 Mc/s the theory is in good agreement with experiment. See also 4005 of 1957 (Medhurst & Hodgkinson).

621.396.41 : 621.396.65 **2547**

The Miami-Havana Radio System and its Integration into the Telephone Networks.—K. P. Stiles, F. G. Hollins, E. T. Fruhner & W. D. Siddall. (*Commun. & Electronics*, March 1958, No. 35, pp. 94–96.) A description of a system developed for radio and television facilities, its method of operation and the results obtained.

621.396.41 : 621.396.822 : 621.317.3 **2548**

The White-Noise Method of Measuring Crosstalk and Noise Interference in Multichannel Telephone Link Systems.—Golding. (See 2498.)

621.396.65 **2549**

Surveying for Microwave Relay Systems.—L. E. Strazza & R. C. S. Joyce. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 262–267.) Surveys in well-mapped regions may involve only site and path inspections, but path tests are necessary if accuracy of path profiles plotted from maps is doubtful. Short-duration tests are used to confirm a calculated mean path loss and hence to confirm tower heights, aerial gains and other system parameters. Long-term tests give statistical information on fading where the path profile is unfavourable. Typical test methods are described.

621.396.65 **2550**

A Radio Link System with Frequency Modulation for 120 Telephony Channels in the Frequency Range 1 700–2 300 Mc/s (FM 120/2 000).—H. Holzwarth, G. Bosse, C. Colani & E. Seibt. (*Nachrichtentech. Z.*, Oct. 1957, Vol. 10, No. 10, pp. 485–493.)

621.396.65 : 621.385.029.63/.64 **2551**

All-Travelling-Wave-Tube Systems.—S. Fedida. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 283–290.) The intrinsic features of microwave repeaters using travelling-wave valves throughout are discussed and their advantages are described. Possible sources of distortion are analysed and methods of reducing their effect to acceptable values are indicated.

621.396.65 : 621.385.029.64 **2552**

S.H.F. Radio Links using Travelling-Wave Output Amplifiers.—G. Dawson & T. K. M. Korytko. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 276–282.) The problems that have arisen in the planning and installation of wide-band links in many different countries are discussed. A description is given of the techniques which have been evolved in the design of the equipment and the associated supervisory and switching gear.

621.396.65 : 621.385.3.029.63 **2553**

Broad-band Microwave Systems employing U.H.F. Triodes.—G. W. S. Griffith & B. Wilson. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 297–302.) The advantages of economy and convenience in arranging for the modulation, demodulation and as much amplification as possible to be carried out at an intermediate frequency are discussed. In the 1 700–2 300-Mc/s band, triodes are the most satisfactory means of providing amplification. Systems employing such principles are described.

621.396.65 : 621.396.41 **2554**

Microwave Radio Toll Systems.—E. W. Anderson. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 267–271.) The function of the radio link in telephone networks is outlined and some of the problems requiring solution in the design of a short-haul or toll system carrying one supergroup are discussed. A typical system is described. The basic defect of nonlinearity in a f.m. system is avoided by using over-all feedback with a resulting over-all simplification of the repeater station.

621.396.65 : 621.397.7 **2555**

Portable U.H.F.-S.H.F. Links in the B.B.C. Television Service.—T. H. Bridgewater. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 291–296.) The various types of system in use for television outside-broadcast purposes are described together with their auxiliary and test equipment.

621.396.65.029.64 **2556**

Microwave Link Development in the Radio Laboratories of the Post Office Engineering Department.—C. F. Floyd & R. W. White. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 253–261.) The contribution made to wide-band microwave system development is surveyed. Work includes the design and installation of complete point-to-point 4-kMc/s television and multichannel telephony systems and associated test equipment. A description is given of a single-hop f.m. telephony link carrying 240 channels on an over-sea path of 55 miles, and a two-hop link for television transmission. Recent work on system equipment and measuring techniques is discussed and trends of development are considered.

SUBSIDIARY APPARATUS

621.314.214 **2557**

Variable-Output Mains Transformer.—H. E. Styles. (*Wireless World*, June 1958, Vol. 64, No. 6, pp. 262–263.) A design using eight windings giving voltages 2^0, $2^1 \ldots . 2^7$ V produces any integral output voltage between 0 and 255 V by simple switching. Auto-transformer operation above 255 V extends the range to 511 V.

621.314.58 : 621.314.7 **2558**

Voltage Conversion with Transistor Switches.—P. L. Schmidt. (*Bell Lab. Rec.*, Feb. 1958, Vol. 36, No. 2, pp. 60–64.) The conversion of a d.c. voltage to a.c. at high audio frequencies at powers of 10–100 W is achieved using a circuit containing transistors and modern magnetic-core components.

621.314.63 : 546.28 **2559**

Failure-Rate Studies on Silicon Rectifiers.—N. F. Bechtold & C. L. Hanks. (*Commun. & Electronics*, March 1958, No. 35, pp. 49–56.) A study of the failure rate of rectifiers rated at 100–400 peak inverse volts under various conditions. Average failure rates for the 400-V type are approximately three times those for the 100-V type under identical conditions.

621.316.72 : 621.314.7 **2560**

Transistor filters Ripple.—F. Oakes & E. W. Lawson. (*Electronics*, 11th April 1958, Vol. 31, No. 15, p. 95.) In a low-voltage d.c. power supply a capacitance of 3 500 μF is effectively multiplied by 60 by the addition of a junction transistor operating as an impedance transformer.

621.316.722.1.076.25 **2561**

Magnetic Voltage Stabilizers.—W. Taeger. (*Elektronik*, Sept. 1957, Vol. 6, No. 9, pp. 265–269.) The use of saturation effects in a number of basic designs is described; brief details of a valve-controlled automatic stabilizer circuit are given.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 : 535.623 : 778.5 **2562**

The Optics of the Lenticular Colour-Film Process.—R. Kingslake. (*J. Soc. Mot. Pict. Telev. Engrs*, Jan. 1958, Vol. 67, No. 1, pp. 8–13.) This type of film, which has a possible use in colour television, is described and the capabilities and limitations are given. Designs of cameras, projectors, printers and viewers are outlined.

621.397.5 : 778.5 **2563**

Monochrome Television Film Standards.—K. B. Benson & J. R. Whittaker. (*J. Soc. Mot. Pict. Telev. Engrs*, Jan. 1958, Vol. 67, No. 1, pp. 1–5.) Each stage in the production, processing and reproduction of the film is examined. Standards for optimum results with the combined film and television system are developed.

621.397.6 : 621.374.4 : 621.318.4 **2564**

Magnetic-Core Dividers for ITV Sync Generators.—Rose. (See 2336.)

621.397.611.2 **2565**

Beam Temperature, Discharge Lag and Target Biasing in some Television Pick-Up Tubes.—B. Meltzer & P. L. Holmes. (*Brit. J. appl. Phys.*, April 1958, Vol. 9, No. 4, pp. 139–143.) The relation between discharge lag and the acceptance of the electron beam by the target is analysed and related to the calculated effective beam temperature. Results are given of the direct measurement of lag and its reduction by target biasing, and anomalous beam temperatures are discussed with regard to electron-optical measurements.

621.397.62 : 535.623 : 535.371.07.002.2 **2566**

Photodeposition of Luminescent Screens.—Sadowsky & Payne. (See 2431.)

621.397.62 : 621.396.662.078 **2567**

Sound Signal tunes TV Automatically.—C. W. Baugh, Jr, & L. J. Sienkiewicz. (*Electronics*, 25th April 1958, Vol. 31, No. 17, pp. 54–58.) The amplitude of the 4·5-Mc/s intercarrier sound signal is used to control the local-oscillator frequency. As the picture level is held constant by a.g.c., this provides automatic fine tuning of the receiver.

621.397.7 **2568**

The Tasks and Structure of the Swiss Television Network.—W. Gerber. (*Tech. Mitt. PTT*, 1st Nov. 1957, Vol. 35, No. 11, pp. 441–455. In German & French.) After tracing the development of the Swiss television network on the basis of the 625-line system details are given of its present extent and its function as part of the Eurovision network. Future planning with regard to system expansion, progress policy and types of links to be used is discussed.

TRANSMISSION

621.396.61 : 621.376.3 **2569**

Design and Construction of a F.M. Transmitter.—D. F. Bowers. (*Brit. Commun. Electronics*, Feb. 1958, Vol. 5, No. 2, pp. 110–114.) The design and performance figures of a 3–6-kW transmitter for band II are given. Parallel operation is also considered.

VALVES AND THERMIONICS

621.314.63 : 621.318.57 **2570**

Solid-State Thyratron switches Kilowatts.—Frenzel & Gutzwiller. (See 2316.)

621.314.63 : 621.373.51 : 621.372.413 **2571**

A Proposed High-Frequency Negative-Resistance Diode.—W. T. Read, Jr. (*Bell Syst. tech. J.*, March 1958, Vol. 37, No. 2, pp. 401–446.) A comprehensive description and analysis is given of a proposed semiconductor diode designed to operate as an oscillator when mounted in a suitable microwave cavity. The frequency would be in the range 1–50 kMc/s, the Q as low as 10 (negative), and the efficiency as high as 30%. Reverse bias establishes a depletion layer of fixed width in a high-resistance region, bounded by very low-resistance end regions. The electric field has a maximum at one end of the space-charge region where hole-electron pairs are generated by avalanche; the holes (or electrons) produce diode current which is delayed about half a cycle relative to the alternating voltage, thus delivering power to the a.c. signal. When the diode is mounted in an inductive microwave cavity tuned to the diode capacitance, an oscillation will build up. It appears possible to exceed 20 W in continuous operation at 5 kMc/s.

621.314.63+537.525.92] : 681.142 **2572**

New Applications of Impedance Networks as Analogue Computers for Electronic Space-Charge and for Semiconductor Diffusion Problems.—Čremošnik, Frei & Strutt. (See 2309.)

621.314.632 : 621.372.632 **2573**

An Analysis of the Diode Mixer Consisting of Nonlinear Capacitance and Conductance and Ohmic Spreading Resistance.—A. C. Macpherson. (*Trans. Inst. Radio Engrs*, Jan. 1957, Vol. MTT-5, No. 1, pp. 43–51. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 576.) See also 4036 of 1957 (Messenger & McCoy).

621.314.632 : 621.372.632 : 621.396.822 **2574**

Mixer Crystal Noise.—N. Houlding. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 917–918.) The effect of mismatches on the noise temperature ratio is considered and it is suggested that the assumption of thermal equilibrium [4036 of 1957 (Messenger & McCoy)] is unsound.

621.314.7 **2575**

Transient Response of Drift Transistors.—R. C. Johnston. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 830–838.) The short-circuit transient response of a drift transistor to a step input current is found by solving the one-dimensional partial differential equation for minority carriers in the base, using a Laplace transform technique. The improvement in rise time due to the built-in field is less in the common-emitter than in the common-base connection. The built-in field lengthens the storage time, but reduces α.

621.314.7 : 621.317.3 **2576**

An Investigation of the Current Gain of Transistors at Frequencies up to 105 Mc/s.—Hyde & Smith. (See 2495.)

621.314.7 : 621.318.57 **2577**

Semiconductor Switching Devices.—(*Electronic Radio Engr*, June 1958, Vol. 35, No. 6, pp. 235–236.) A note on the characteristics of three new types of transistor. Information is condensed from articles in *Trans. Inst. Radio Engrs*, Jan. 1958, Vol. ED-5, No. 1.

621.314.7 : 621.396.822 **2578**

Transistor Noise: its Origin, Measurement and Behaviour.—B. L. H. Wilson. (*J. Brit. Instn Radio Engrs*, April 1958, Vol. 18, No. 4, pp. 207–225.) "The sources of noise in semiconductors and the mathematical techniques needed in their discussion are indicated in order to survey the theory of noise in transistor amplifiers and to consider methods of measurement. The variation of transistor noise with operating point and frequency is discussed and a comparison is made of noise levels in audio amplifiers using transistors and valves respectively."

621.314.7.001.4 (083.74) **2579**

I.R.E. Standards on Solid-State Devices: Methods of Testing Point-Contact Transistors for Large-Signal Applications, 1958.—(*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 878–888.) Standard 58 I.R.E. 28.S1.

621.314.7.002.2 : 546.28 **2580**

Manufacture of Silicon Transistors.—J. T. Kendall. (*Electronic Radio Engr*, June 1958, Vol. 35, No. 6, pp. 202–207.) Current techniques for the production of six types of transistor are described, with a discussion of impurity distributions. Types manufactured in the U.K., and their main characteristics are listed.

621.314.7.004.2 **2581**

Naval Material Laboratory Transistor Reliability Study.—R. E. Martin. (*Trans. Inst. Radio Engrs*, June 1957, No. PGRQC-10, pp. 49–56.) The effects of collector voltage, of collector power dissipation and of ambient temperature on the reliability of low-power *p-n-p* alloy-junction transistors are to be investigated in a special laboratory. 3 000 samples are to be tested for 10 000 h or longer.

621.314.7.004.2 **2582**

Success Story—Transistor Reliability—1956.—C. H. Zierdt, Jr. (*Trans. Inst. Radio Engrs*, June 1957, No. PGRQC-10, pp. 57–68.) The electrical characteristics of a wide range of samples of Ge and Si junction transistors were examined under various temperature conditions and for various periods of time. Preliminary tests of the effects of nuclear radiation are also reported.

621.314.7.004.2 : 546.289 **2583**

Factors in the Reliability of Germanium Power Transistors.—A. B. Jacobsen. (*Trans. Inst. Radio Engrs*, June 1957, No. PGRQC-10, pp. 43–48.) The reliability of a *p-n-p* alloy-junction transistor used for car receivers is discussed.

621.385.004.15 **2584**

A Basic Study of the Effects of Operating and Environmental Factors on Electron-Tube Reliability.—W. S. Bowie. (*Trans. Inst. Radio Engrs*, Feb. 1956, No. PGRQC-6, pp. 46–56. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1956, Vol. 14, No. 2, p. 277.)

621.385.004.15 **2585**

Electron-Tube Life and Reliability—Built-In Tube Reliability.—M. A. Acheson. (*Trans. Inst. Radio Engrs*, April 1956, No. PGRQC-7, pp. 33–43. Abstract, *Proc. Inst. Radio Engrs*, July 1956, Vol. 44, No. 7, p. 957.)

621.385.004.2 **2586**

Environmental Effects on Vacuum-Tube Life.—H. C. Pleak & A. V. Baldwin. (*Trans. Inst. Radio Engrs*, Jan. 1957, No. PGRQC-9, pp. 93–101. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 577.) See 2956 of 1957.

621.385.004.6 **2587**

Prediction of Tube Failure Rate Variations.—M. P. Feyerherm. (*Trans. Inst. Radio Engrs*, Jan. 1957, No. PGRQC-9, pp. 65–71. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 577.)

621.385.029.6 **2588**

Industrial Magnetrons.—W. Schmidt. (*Elektronische Rundschau*, Oct. 1957, Vol. 11, No. 10, pp. 306–309.) The differences in the requirements to be met in radar and industrial applications of magnetrons are summarized. Brief details are given of some German industrial types for operation at 2 400 Mc/s.

621.385.029.6 **2589**

Very - Low - Noise Travelling - Wave Amplifier.—E. W. Kinaman & M. Magid. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 861–867.) A general explanation of beam noise is given with a description of noise-reduction systems lowering noise figures from 9 to 6 dB.

621.385.029.6 **2590**

Backward-Wave Oscillators.—A. G. Stainsby. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 329–334.) These oscillators can be tuned electronically with a frequency range greater than an octave. There are two distinct types of oscillator, the O type which is mainly suited to low-power operation as a local oscillator or test source and the M type which is capable of delivering high powers with good efficiency and is thus suitable for use as a transmitter.

621.385.029.6 **2591**

Understanding the Backward-Wave Oscillator.—D. A. Dunn. (*Electronic Ind.*, Jan. 1958, Vol. 17, No. 1, pp. 72–76.) An explanation of the physical processes which take place in backward-wave oscillators.

621.385.029.63/.64 **2592**

Travelling - Wave Tubes in Communications.—R. B. Coulson. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 302–304.) Their application in microwave repeaters and the use of electromagnets in preference to permanent magnets for beam focusing is discussed.

621.385.029.63/.64 **2593**

Travelling-Wave-Tube Amplifiers.—D. H. O. Allen. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 305–309.) The major advances made towards a reliable and useful valve are discussed with particular reference to noise characteristics and focusing methods. Applications to microwave links and the distortions that may arise under various operating conditions are also discussed.

621.385.029.63/.64 **2594**

S-Band Travelling-Wave Tube with Noise Figure below 4 dB.—M. Caulton & G. E. St. John. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, pp. 911–912.) Modifications to the electrode configuration and voltage profile of a conventional low-noise travelling-wave valve produced a hollow beam and low-velocity drift region near the cathode. Noise figures down to 3·5 dB were obtained.

621.385.029.63/.64 **2595**

Reflex Klystrons.—A. H. Atherton. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 315–320.) A general description of the construction and operation of reflex klystrons is given, with typical examples.

621.385.029.63/.64 **2596**

Multicavity Klystrons.—V. J. Norris. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 321–323.) "A review is given of the factors determining the gain, efficiency and bandwidth of multicavity klystrons, together with results which have been achieved in practice."

621.385.029.63/.64 : 621.372.414 **2597**

Coaxial - Line Velocity - Modulated Oscillator Valves. — D. E. Lambert. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 324–328.) The principle of operation of these valves is discussed, and also their advantages and limitations. A list of valves in production is given.

621.385.029.64 **2598**

Travelling-Wave Tubes for 4 000 Mc/s.—P. F. C. Burke. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 310–314.) Travelling-wave valves may be used in microwave radio links as output amplifiers, low-noise receivers, and intermediate-level amplifiers. Examples are given of applications at frequencies near 4 kMc/s.

621.385.032.213.13 **2599**

High-Emission Hot Cathodes of New Type.—T. Hashimoto, M. Uchida & A. Yokoyama. (*Rep. elect. Commun. Lab., Japan*, Dec. 1957, Vol. 5, No. 12, pp. 4–8.) Research has been carried out on cathodes impregnated with various barium compounds for use in microwave tubes. Work has also been done on sintered cathodes. Details of life tests are given. See also 1927 of June (Hashimoto).

621.385.032.213.13 **2600**

A Sintered Nickel Matrix Cathode.—R. W. Fane. (*Brit. J. appl. Phys.*, April 1958, Vol. 9, No. 4, pp. 149–153.) Cathodes containing alkaline earth carbonates and boron have given 8 000 h life at current densities in excess of 0·5 A/cm². Processing schedules and operating temperatures are similar to those for normal oxide-coated types. Possible mechanisms of operation are discussed.

621.385.032.269.1 **2601**

A New Type of Low-Noise Electron Gun for Microwave Tubes.—M. R. Currie. (*Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 911.) Noise figures of 3·5 dB are obtained with the gun system whose basic features are described.

621.385.1 + 621.387 **2602**

Thermionic and Cold-Cathode Valves.—W. H. Aldous. (*Proc. Instn elect. Engrs*, Part B, May 1958, Vol. 105, No. 21, pp. 273–281.) A full survey of valve development since 1933 is given, with particular emphasis on the decrease in size of all types of valves.

621.385.3 **2603**

On the Amplification Factor of a Triode Valve: Part 2.—E. B. Moullin. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 196–202.) A description of measurements designed to test the degree to which the amplification factor is independent of the anode current (see 2649 of 1957).

621.385.3/.4].029.63/.64 **2604**

Triodes and Tetrodes for U.H.F.-S.H.F. Operation.—C. A. Tremlett & A. D. Williams. (*Electronic Engng*, May 1958, Vol. 30, No. 363, pp. 335–340.) A review of the advances made over the last ten years in the development of space-charge-controlled valves for operation above 300 Mc/s. Valve geometry and performance data for various types are listed.

621.385.832 : 621.317.755 **2605**

Broad - Band Oscilloscope Tube.—Brangaccio, Dietrich & Sullivan. (See 2511.)

MISCELLANEOUS

061 : 621.3 (436) (091) **2606**

75 Years of the Elektrotechnischer Verein Österreichs.—(*Elektrotech. u. Maschinenb.*, 1st May 1958, Vol. 75, No. 9., pp. 177–340.) The history of the Association and its journal are outlined and the contributions of Austrian scientists and engineers in the field of electrical engineering are summarized in a number of papers which include the following:

(*a*) The Development of Electrical Communications in Austria.—H. Schmid (pp. 300–320).

(*b*) The Contribution of Austria in the Development of High-Frequency Engineering.—G. Caspar (pp. 320–326).

(*c*) Austria's Contribution in the Development of Electroacoustics.—F. Lachner (pp. 330–332).

621.3.029.6 (091) **2607**

Microwaves.—E. L. Ginzton. (*Science*, 18th April 1958, Vol. 127, No. 3303, pp. 841–851.) The historical development of microwave research is reviewed and current applications and techniques are outlined.

621.3.049 **2608**

Optimum Tolerance Assignment to Yield Minimum Manufacturing Cost.—D. H. Evans. (*Bell Syst. tech. J.*, March 1958, Vol. 37, No. 2, pp. 461–484.) The basic item considered is a unit with a single nominal design response, having several components with given nominal design values such that the unit response is as required. Tolerance assignment is considered in relation to cost and salvage value of a unit, and statistical results are presented.

621.3.049 : 621-52 **2609**

Automatic Methods in Radio Component Manufacture.—D. Stevenson & R. B. Shepherd. (*J. Brit. Instn Radio Engrs*, April 1958, Vol. 18, No. 4, pp. 227–231.) Two special systems are described: (*a*) an electronic counter technique applied to the accurate control of a high-speed coil winding machine; (*b*) a power press producing ferrous piece-parts protected from jamming by detecting each ejected part magnetically.

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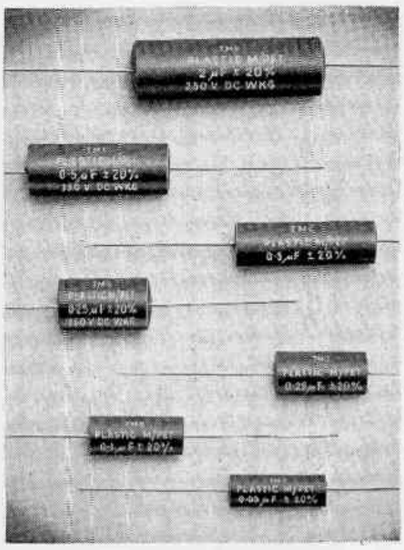

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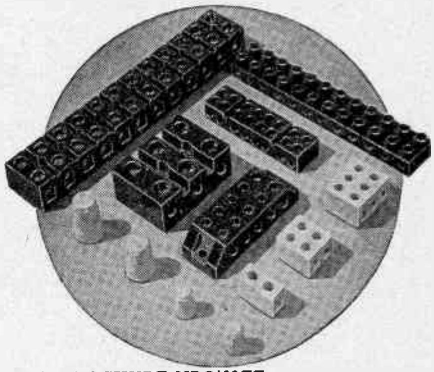

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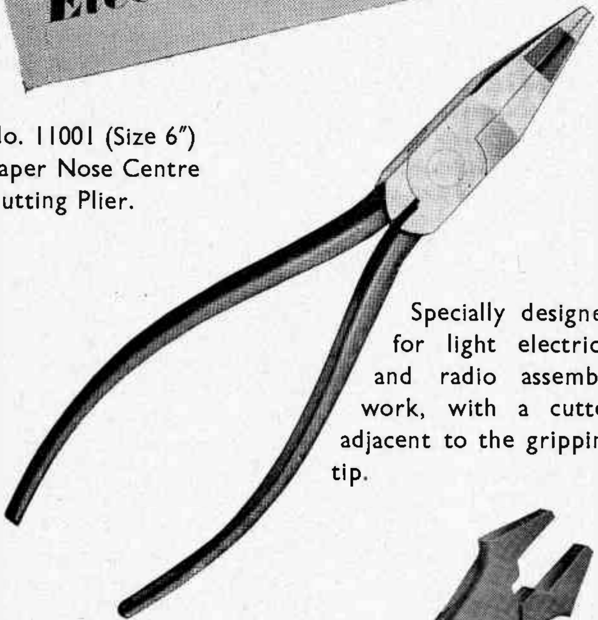

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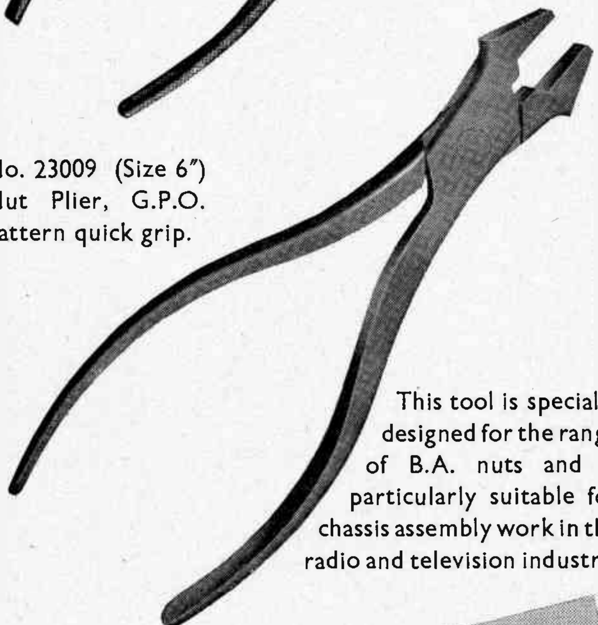

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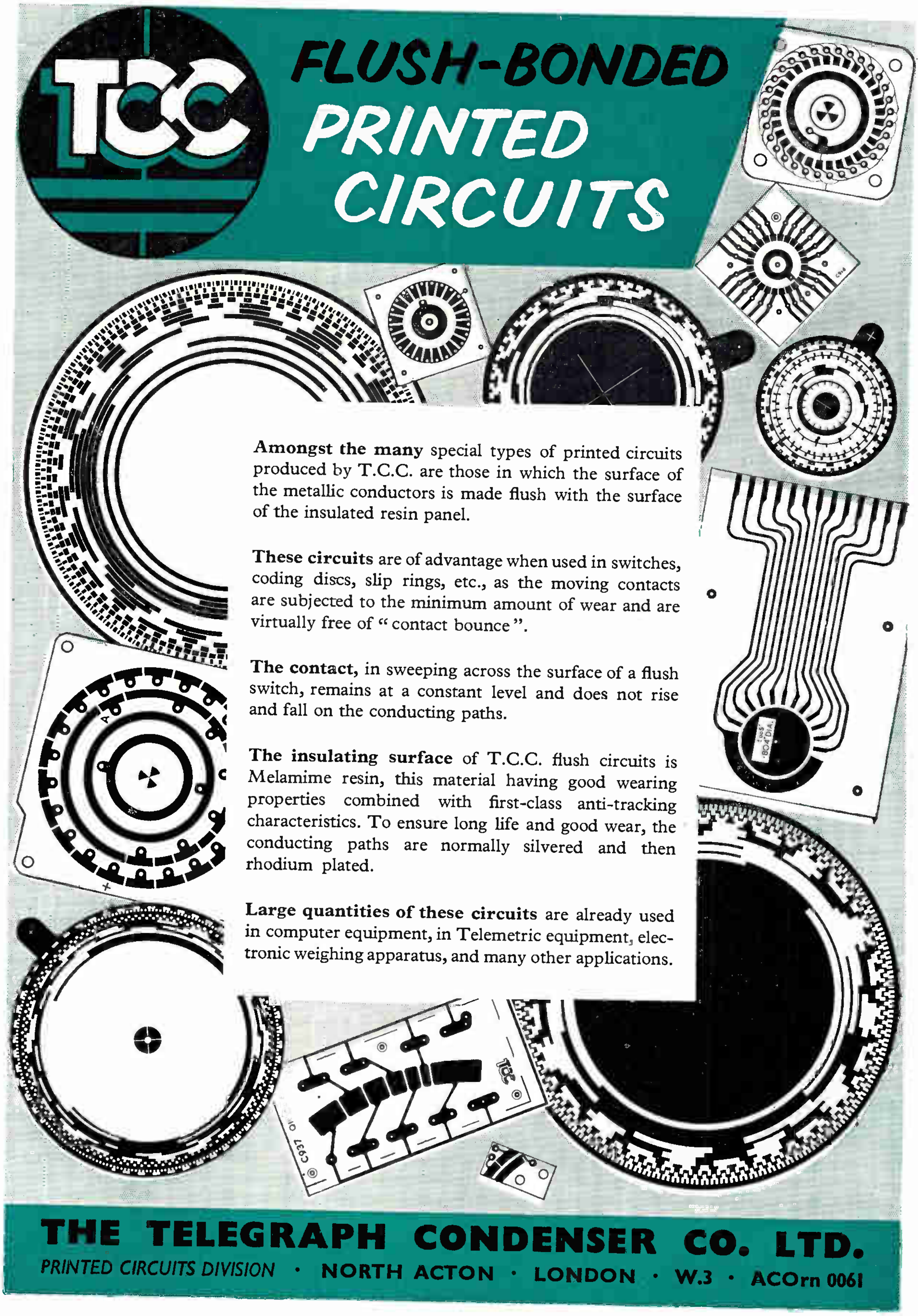
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