WIRELESS ENGINEER

ELECTRONIC & RADIO ENGINEER

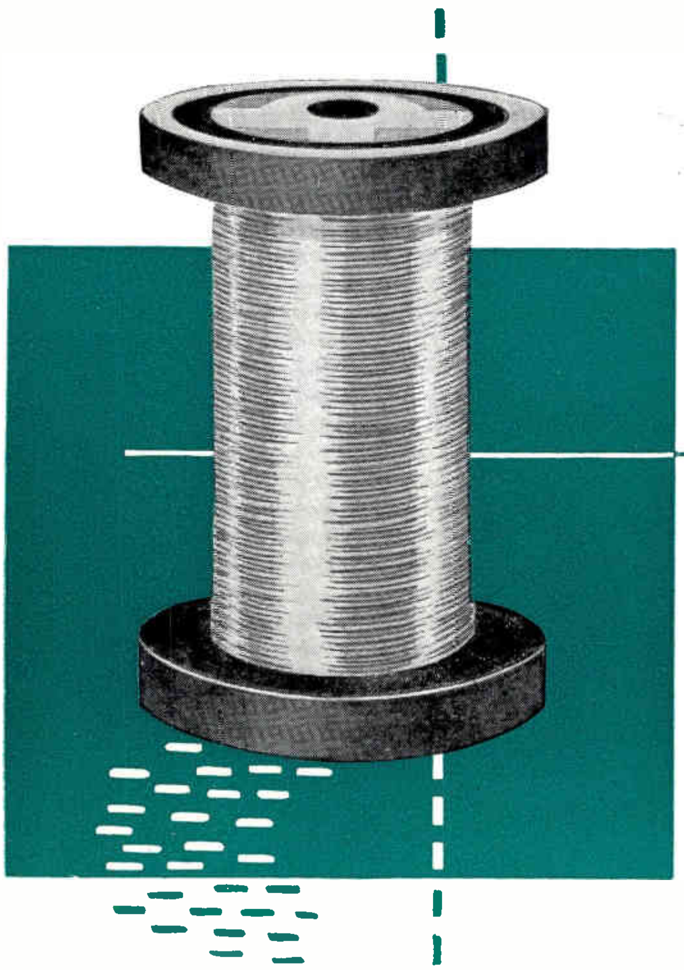
Incorporating **WIRELESS ENGINEER**

In this issue

Subjective Sharpness of Television Pictures
Parallel-T RC Selective Amplifiers
Limited-Gain Operational Amplifiers
Scattering of E.M. Waves by Long Cylinders

Three shillings
and sixpence

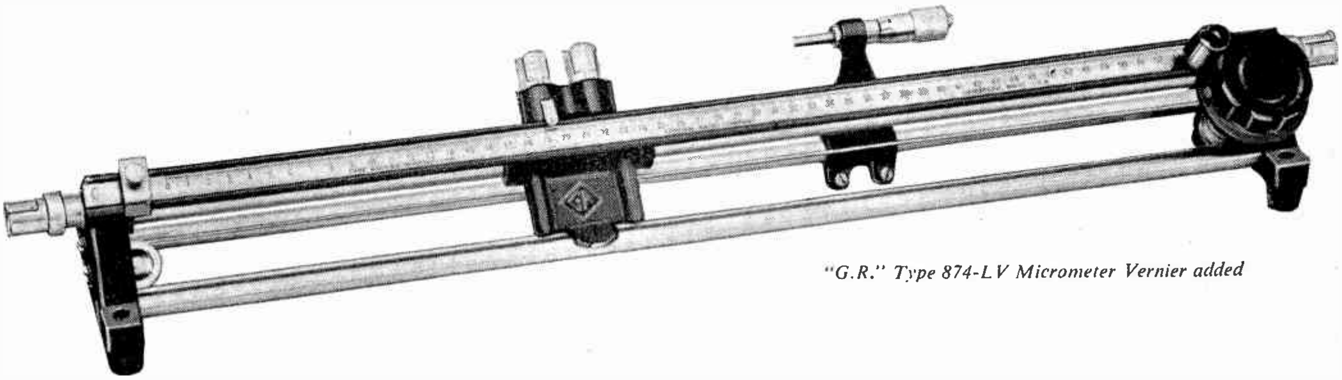
APRIL 1958 Vol 35 *new series* No 4



Trade Mark

U.H.F. MEASURING EQUIPMENT

"G.R." Type 874-LBA Slotted Line



"G.R." Type 874-LV Micrometer Vernier added

This precision-made instrument is an invaluable laboratory tool for impedance measurements and determinations of VSWR or mismatch, at any frequency from 300 to 5,000 Mcs/s. With some sacrifice in accuracy measurements can also be made down to 150 Mc. and above 5,000 Mc.

Very superior electrical characteristics and many operating conveniences, coupled with compactness and light weight, make it extremely useful in field work as well as in the test-room or development laboratory.

In no other way are the fundamentals of high-frequency wave propagation more effectively illustrated than by the Slotted Line. It clearly indicates positions of minima and maxima, determines voltage amplitudes and effects of different load terminations, etc.

" GENERAL RADIO " also offer a *complete* range of coaxial elements and other accessories for use with their Type 874-LBA Slotted Line, such as: Oscillators, Detectors, 874-UB " Balun ", Component Mount, 50-ohm, short-circuit, open-circuit and other terminations, attenuators, line-stretchers, micrometer vernier, filters, tees, coaxial adaptors, etc., etc. Indeed, only from " GENERAL RADIO " can you obtain ALL the equipment necessary for ALL needful measurements, ALL designed to work together flexibly and rapidly, and ALL being moderately priced. The " G.R." Catalogue " O " (258 pages) gives all information and is available promptly against all bona fide written enquiries, on official letter-headings.

Features of the "G.R." 874-LBA Slotted Line

BASIC FREQUENCY RANGE:
300 to 5,000 Mc/s.

CHARACTERISTIC IMPEDANCE:
50 ohms ±1%.

PROBE TRAVEL:
50 Cm.: scale calibrated in mm.

PROBE PENETRATION: Fully adjustable.

DIELECTRIC: Air.

ACCURACY:
Constancy of Probe Penetration—±1½%. VSWR of Terminal Connectors less than 1.025 at 1,000 Mc. and less than 1.07 at 4,000 Mc.

CRYSTAL RECTIFIER: 1N21BR.

PORTABILITY: Weighs only 8½ lbs.

MICROMETER VERNIER: 874-LV attachment, and a harmonic filter is necessary where VSWR's greater than 10 are to be measured.

PRICES: 874-LBA Slotted Line ... £121. 0. 0.
874-LV Micrometer Vernier £13.15. 0.

Claude Lyons Ltd.

76 OLDHALL STREET, LIVERPOOL 3 LANCS. TELEPHONE: CENtral 4641/2
VALLEY WORKS, HODDESDON, HERTS. TELEPHONE: HODdesdon 3007-8-9

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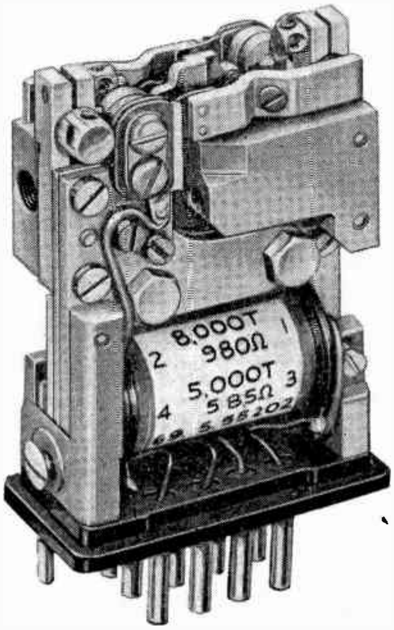
INSTRUMENTS, ELECTRONICS & AUTOMATION

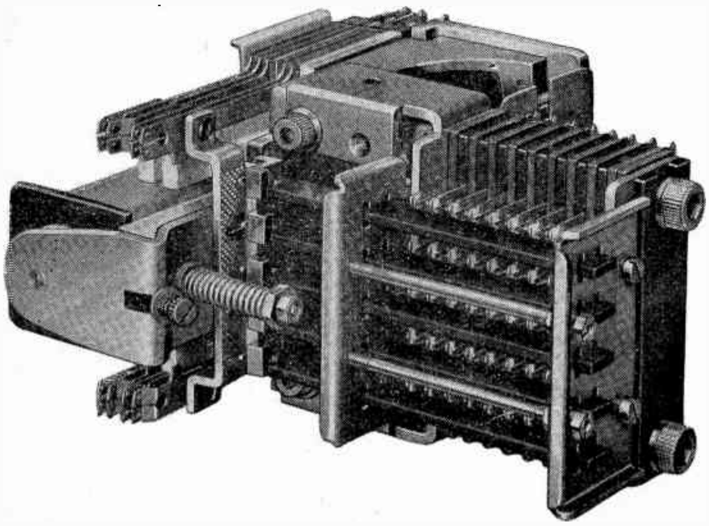
When you have a problem in HIGH SPEED SWITCHING, CONTROL, AMPLIFICATION, IMPULSE REPETITION or MULTI-POINT SWITCHING —

there is no question about the two components that will go farther than others towards giving you the efficiency you *must* have —

THE CARPENTER POLARIZED RELAY

It operates quickly on extremely short, weak or ill-defined impulses of varying polarity — has a close operate/release differential and retains accurate adjustment for long periods in wide variations of temperature, acceleration and vibration. Five basic types are available in a series of specialized versions, with a wide range of single and multiple windings.





T.M.C. CROSSBAR BRIDGE

A compact, versatile multi-point switching unit. Comprising 5 banks of 8 make-contacts, one unit can select any 1 of 5 points or, in association with a simple change-over relay, any 1 of 10 points (4-way). Two or more Bridges can provide multiple outlets, and, interacting electrically, can provide complex counting and stepping facilities. For multi-point switching in *Telemetering*, *Process Control*, *Serial testing* equipment, etc. the Bridge has many advantages to offer.

See demonstrations of these components at our STAND 302 "Instruments, Electronics & Automation" Exhibition, Olympia, London 16th-25th April, and STAND 72 R.E.C.M.F. Exhibition, London 14th-17th April.

or write for technical data to

TMC TRADE MARK

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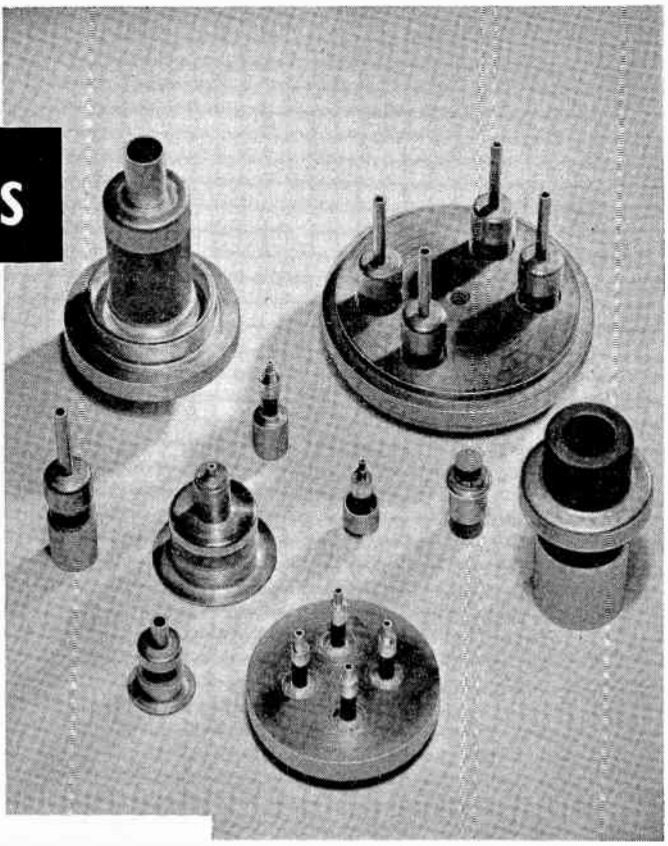
Hollingsworth Works · Martell Road · West Dulwich, S.E.21 · Telephone: GIPsy Hill 2211

FERRANTI

CERAMIC TO METAL SEALS



Ferranti Ceramic to Metal Seals have been developed to overcome the limits imposed upon valve performance by the use of conventional glass to metal seals. They perform so successfully that uses are apparent in other fields, particularly as terminals for vacuum and pressure vessels operated at elevated and sub-normal temperatures.

CHARACTERISTICS

Operating Temperature	intermittent 700°C maximum—continuous 300°C - 450°C dependant on atmosphere.
Operating Pressure	at least 100 atmospheres, depending on direction of compression.
Mechanical Strength	shearing force = 2,500 lbs per square inch of seal area.
Electrical Insulation	breakdown voltage in air is greater than 24 kV per inch of ceramic between seals.
Leakage Resistance (Typical)	10^{13} to 10^{15} ohms, at room temperature between two metal rings separated by 0·3″ of clean ceramic surface on a seal 0·440″ in diameter.
High Frequency Performance (Typical)	loss of 56 watts when 1 kW C.W. is passed through a seal incorporated in a ceramic-filled X-band circular wave-guide.

APPLICATIONS

The mechanical and electrical properties listed above suggest a wide variety of uses, a few of which are given below:-

Microwave and ordinary valve envelopes.
Semi-conductor envelopes.
Terminations for single and multicore cables.
Terminals and leads for vacuum and pressure vessels in atomic energy projects, thermocouple seals for furnaces, etc.

F
FERRANTI

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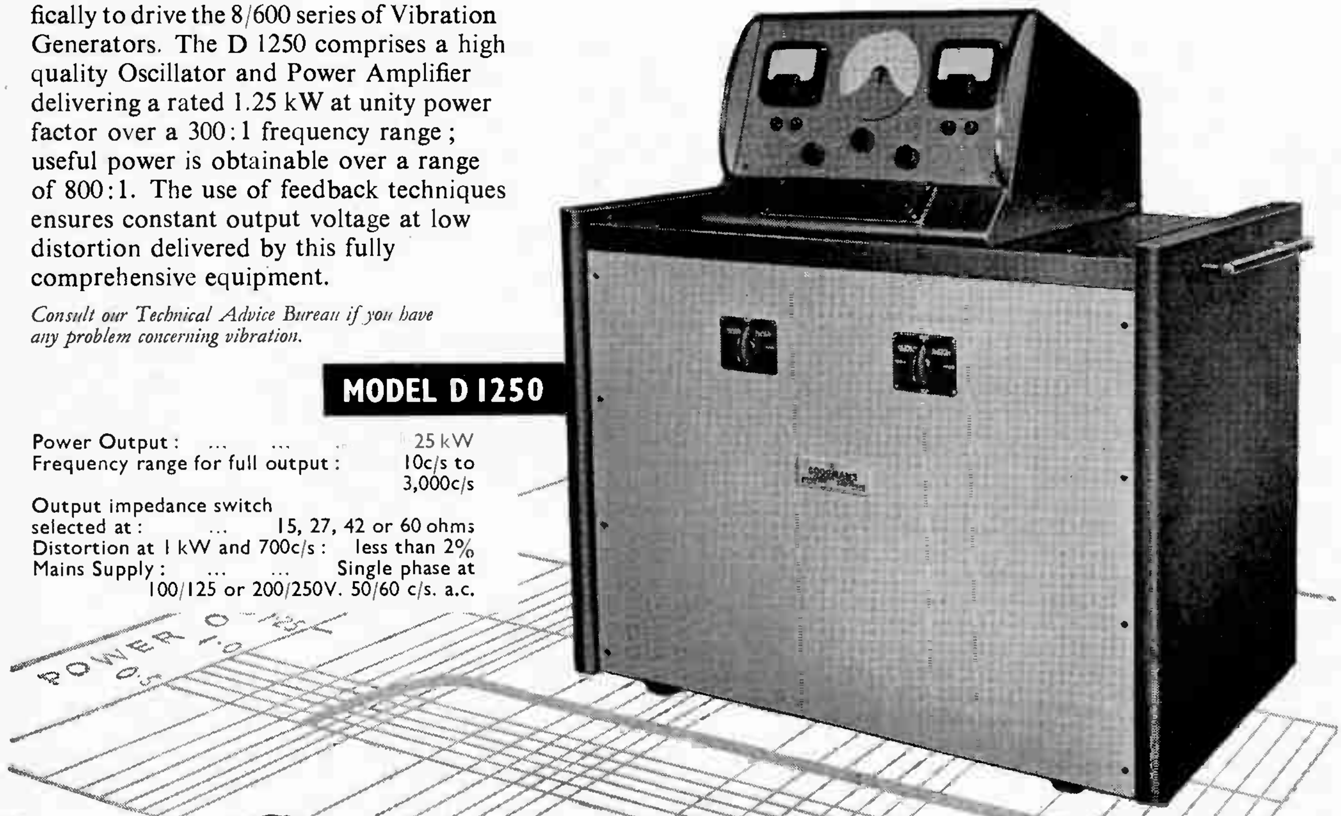
FE181

VIBRATION EQUIPMENT

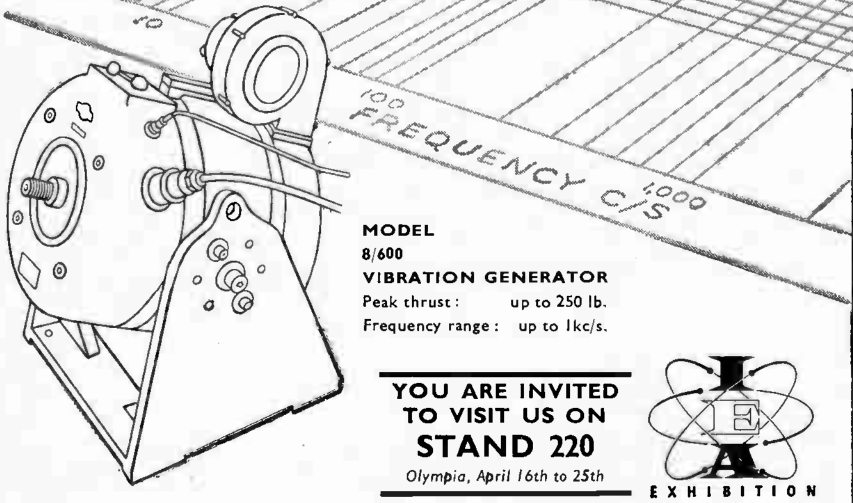
The D. 1250 OSCILLATOR — POWER AMPLIFIER from GOODMANS range of Vibration Equipment is designed specifically to drive the 8/600 series of Vibration Generators. The D 1250 comprises a high quality Oscillator and Power Amplifier delivering a rated 1.25 kW at unity power factor over a 300:1 frequency range; useful power is obtainable over a range of 800:1. The use of feedback techniques ensures constant output voltage at low distortion delivered by this fully comprehensive equipment.

Consult our Technical Advice Bureau if you have any problem concerning vibration.



MODEL D 1250

Power Output: ... ... 1·25 kW
Frequency range for full output: 10c/s to 3,000c/s
Output impedance switch selected at: ... 15, 27, 42 or 60 ohms
Distortion at 1 kW and 700c/s: less than 2%
Mains Supply: ... ... Single phase at 100/125 or 200/250V. 50/60 c/s. a.c.



MODEL 8/600 VIBRATION GENERATOR

Peak thrust: up to 250 lb.
Frequency range: up to 1kc/s.

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Olympia, April 16th to 25th

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GOODMANS Vibration Equipment

GOODMANS INDUSTRIES, LIMITED
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GD.20

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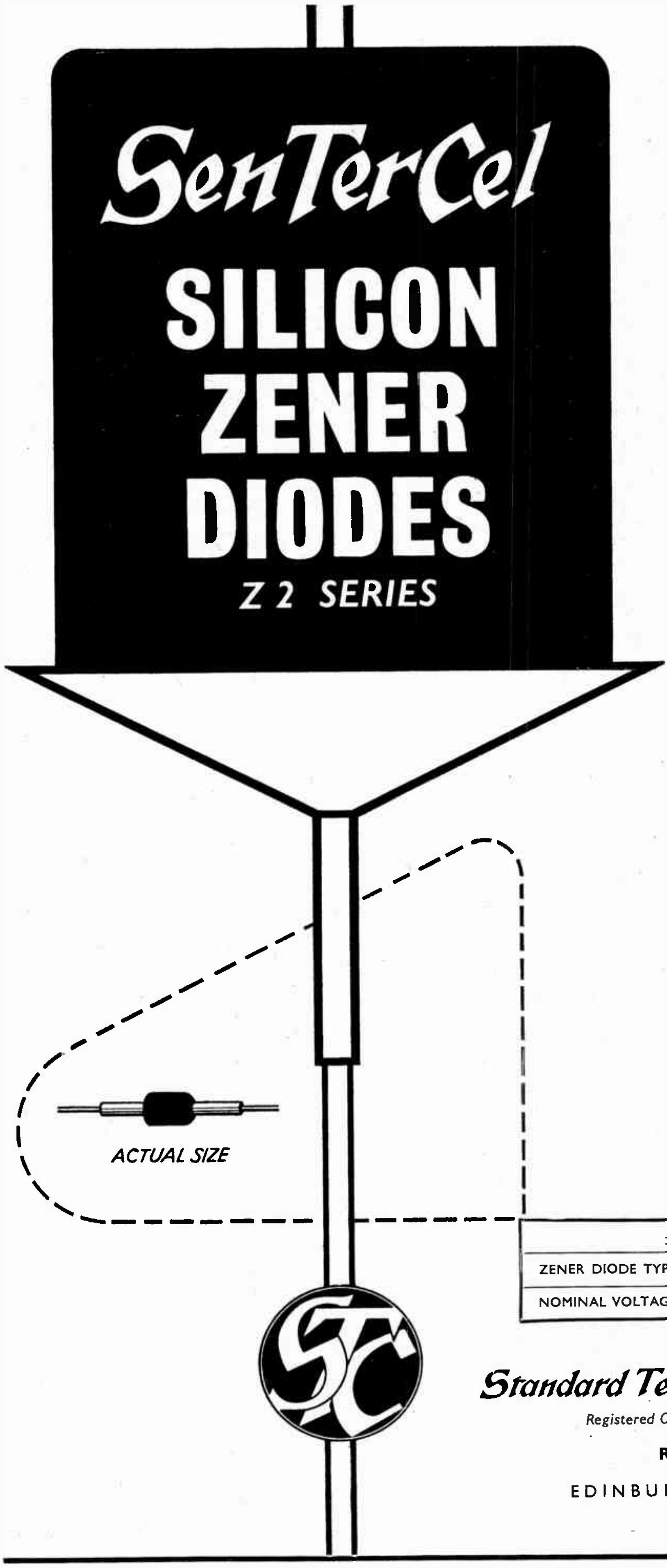
AXIOM WORKS · LANCELOT RD. · WEMBLEY · MIDDX.
ENGLAND

We are interested in the following:—

☐ General Catalogue
☐ Equipment to vibrate load of ....lb at ....g(........in) ....c/s to........c/s
☐ Suitable driving equipment
☐ Other items ....
☐ A vibration system for the duty of ....

Name ....
Company ....
Address ....

Please mark details required. R20



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ZENER DIODE TYPE	Z2A33	Z2A47	Z2A68	Z2A100	Z2A150
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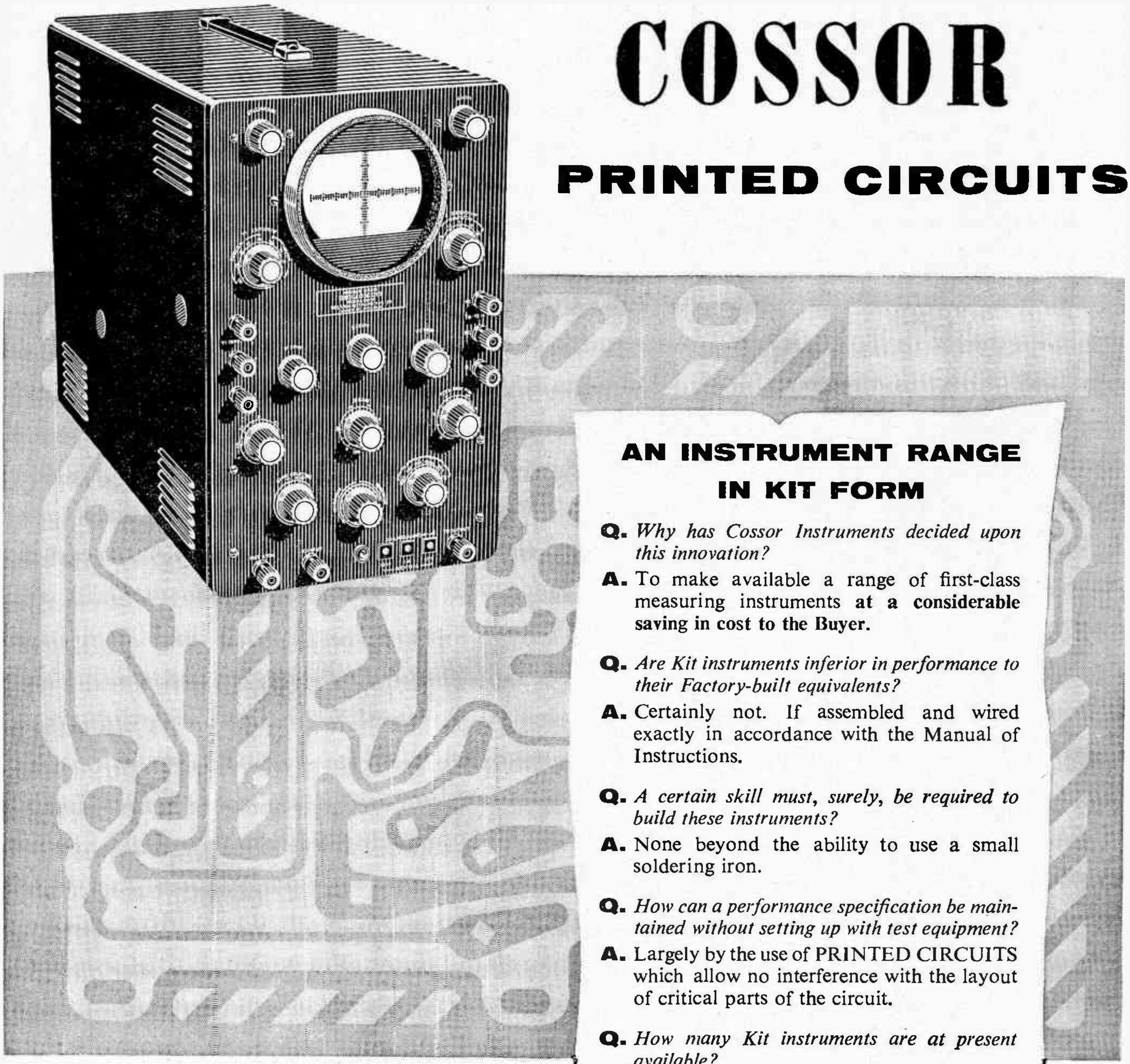
COSSOR

PRINTED CIRCUITS



AN INSTRUMENT RANGE IN KIT FORM

Q. *Why has Cossor Instruments decided upon this innovation?*

A. To make available a range of first-class measuring instruments **at a considerable saving in cost to the Buyer.**

Q. *Are Kit instruments inferior in performance to their Factory-built equivalents?*

A. Certainly not. If assembled and wired exactly in accordance with the Manual of Instructions.

Q. *A certain skill must, surely, be required to build these instruments?*

A. None beyond the ability to use a small soldering iron.

Q. *How can a performance specification be maintained without setting up with test equipment?*

A. Largely by the use of PRINTED CIRCUITS which allow no interference with the layout of critical parts of the circuit.

Q. *How many Kit instruments are at present available?*

A. Three. Two Oscilloscopes, a Single-Beam and a Double-Beam, and a Valve Voltmeter. Others will follow shortly.

Q. *Could I have more information on these interesting instruments?*

A. With the greatest of pleasure. Just write to:

Model 1071K ***Double Beam*** *Kit Oscilloscope*
List Price £69.0.0
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From 200 cycles/sec. to 90 Mc/sec.

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

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

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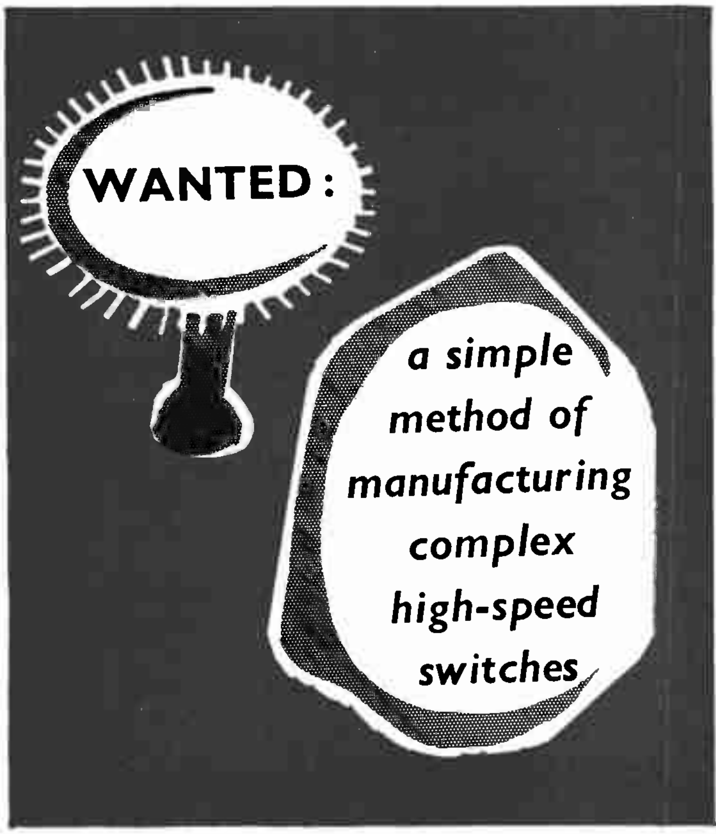
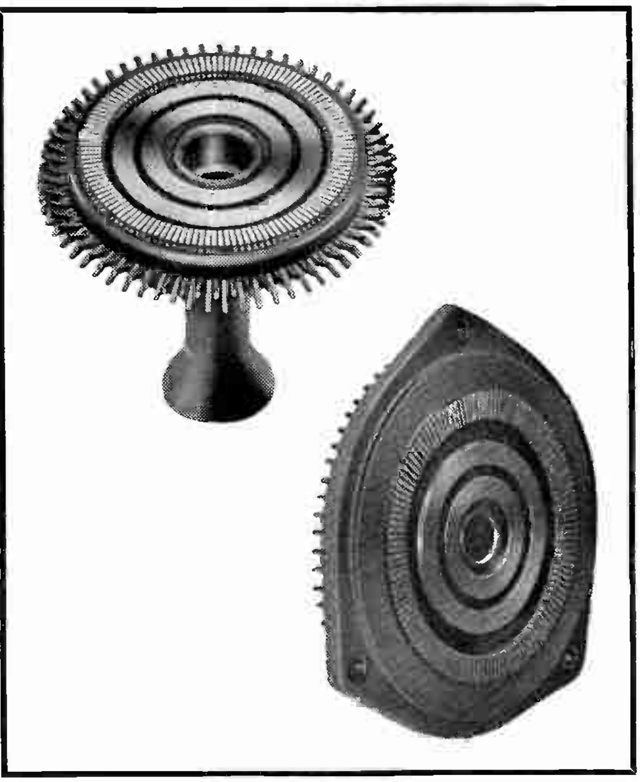


These switches for machine tool control, made only by EMI Electronics Ltd., are excellent examples of advanced resin moulding technique. In planning simplified production based on resin moulding, the big problem was to find a resin formulation capable of fulfilling the exacting performance requirements.



FOUND: *Epikote Resins*
with 'all the qualities'

Yes, *all* of them — no distortion with temperature fluctuation. Excellent adhesion. Excellent electrical and mechanical properties. Low shrinkage on cure. High resistance to heat, water and chemicals. By using an Epikote Resin formulation as support and insulator, EMI Electronics Ltd., have produced an exceptionally efficient component, made without recessing or other machine operations.

Have you a production problem for which Epikote Resins are the perfect answer? Full technical information will be gladly given.

EPIKOTE EPOXY RESINS

unrivalled for potting and casting

SHELL CHEMICAL COMPANY LIMITED. *In association with Petrochemicals Ltd.—Styrene Products Ltd.*
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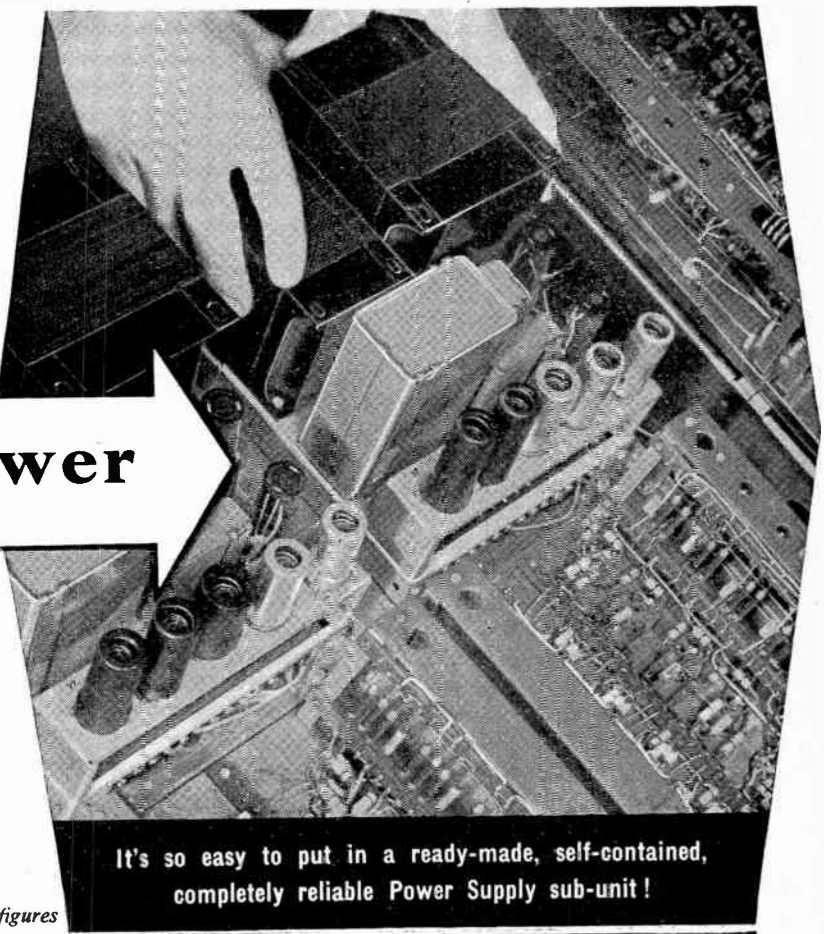
SS 20

improved

SOLARTRON

‘packaged’ power

sub-units



It's so easy to put in a ready-made, self-contained, completely reliable Power Supply sub-unit!

these are design centre performance figures

	SRS 156	AS 516	AS 517	AS 616	AS 619	AS 754	AS 755
VOLTAGE	±150V	±250V or ±300V	±250V or ±300V	±250V or ±300V	±150V to ±180V	±250V or ±300V	±250V or ±300V
CURRENT	0-40mA	0-50mA	10-100mA	0-1A	0-200mA	0-200mA	0-500mA
A.C. OUTPUTS	6.3V 4A C.T. 6.3V 1A	6.3V 4A C.T. 6.3V 1A	6.3V 4A C.T. 6.3V 2A C.T. 6.3V 1A	6.3V 10A C.T.	6.3V 4A C.T. 6.3V 4A C.T. 6.3V 1A C.T.	6.3V 6A C.T. 6.3V 6A C.T. 6·3V 2A	6.3V 10A C.T. 6.3V 5A C.T.
D.C. SOURCE IMPEDANCE	<2Ω	0·3Ω	0·2Ω	0·4Ω	0·5Ω	0·4Ω	0·05Ω
STABILISATION FACTOR	400:1	1000:1	800:1	500:1	400:1	500:1	500:1
A.C. SOURCE IMPEDANCE 40c/s—100kc/s	<2Ω	0·1Ω	0·15Ω	0·2Ω	0·2Ω	0·4Ω	0·2Ω
RIPPLE AND NOISE CONTENT	<350μV	100μV	150μV	300μV	100μV	150μV	300μV
DIMENSIONS	9⅛″×6⅛″ ×6¼″ high	9⅛″×6⅛″ ×6¼″ high	9½″×7⅛″ ×6¼″ high	19″×19″ ×10½″ high	10″×10″ ×6$\frac{7}{16}$″ high	13″ × 6¾″ × 6¼″ high	19″ × 12″ × 8¾″ high
WEIGHT	14½ lbs	14½ lbs	20½ lbs	150 lbs	21¼ lbs	25 lbs	56 lbs

- * Better stability factors
- * More heater power
- * Less ripple and noise
- * No increase in cost for this improved performance
- * Lower source impedance
- * Stability unaffected by capacity loading
- * Delivery ex-stock

SOLARTRON Power Supply Sub-units

THE SOLARTRON ELECTRONIC GROUP LTD • THAMES DITTON • SURREY • ENGLAND

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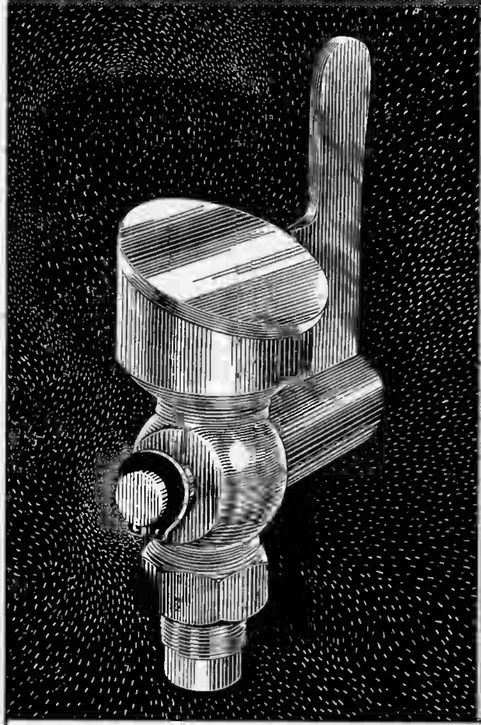
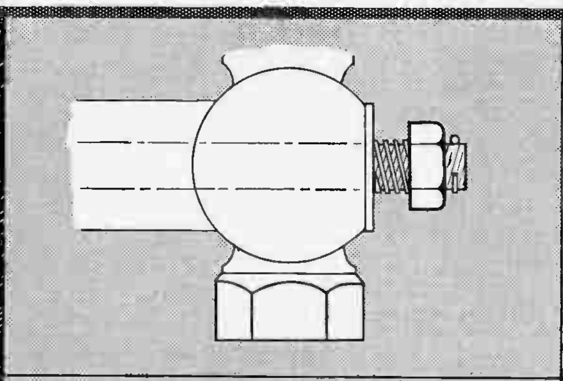
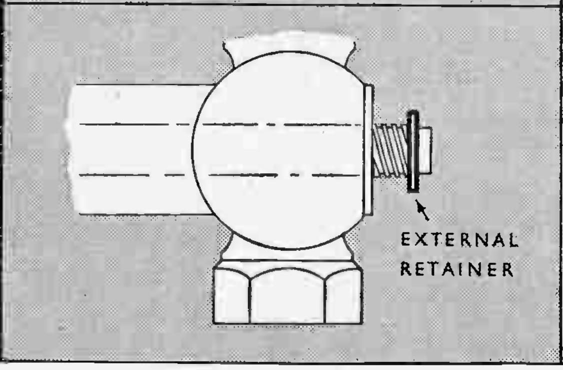
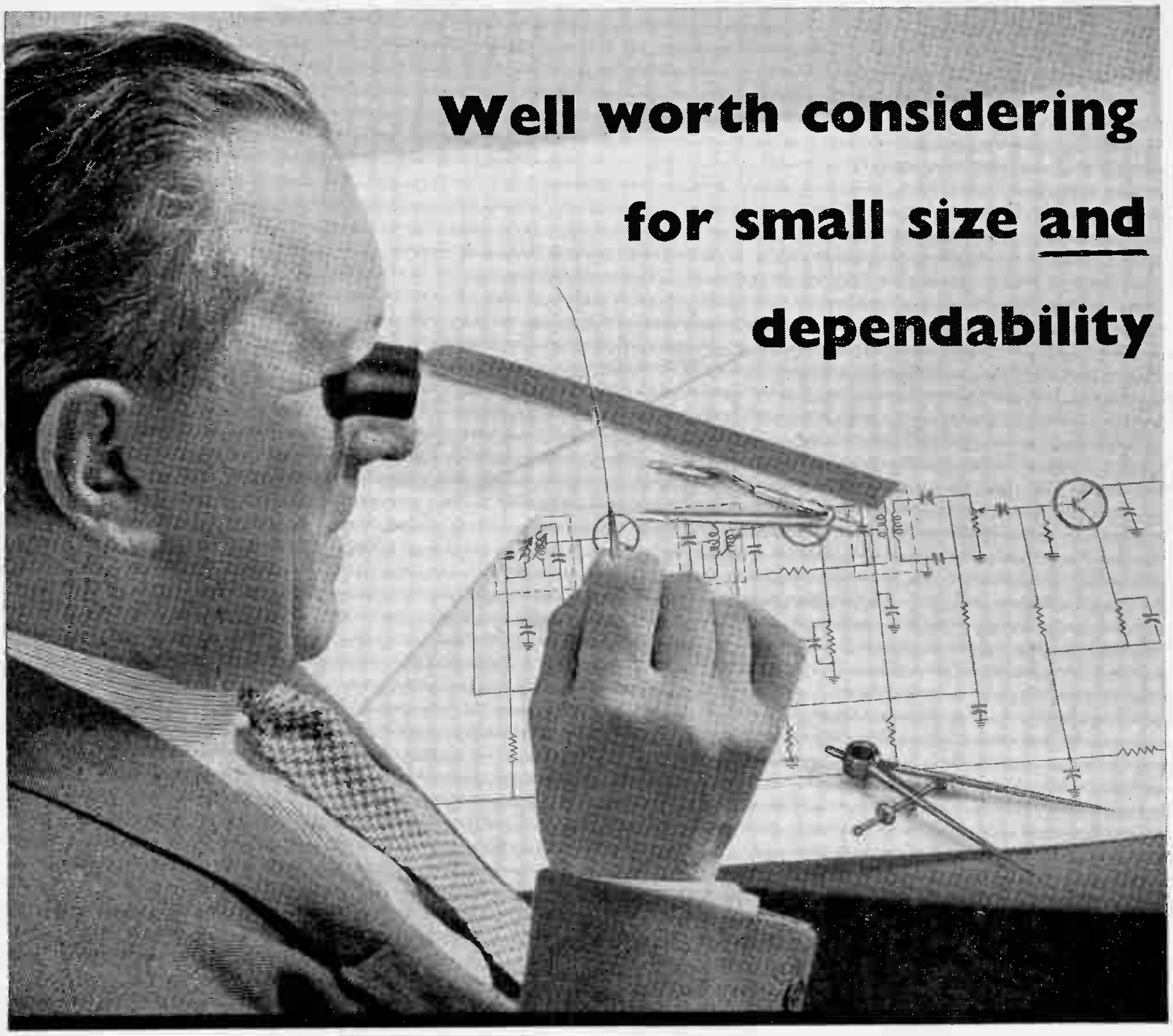





EXTERNAL RETAINER



The new type BTH Germanium Point Contact Rectifiers—

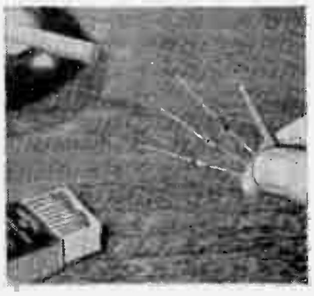
Only ¼ in. long, yet their miniature size is combined with high performance and complete dependability! They offer the following outstanding characteristics:

- HIGH TEMPERATURE STABILITY
- 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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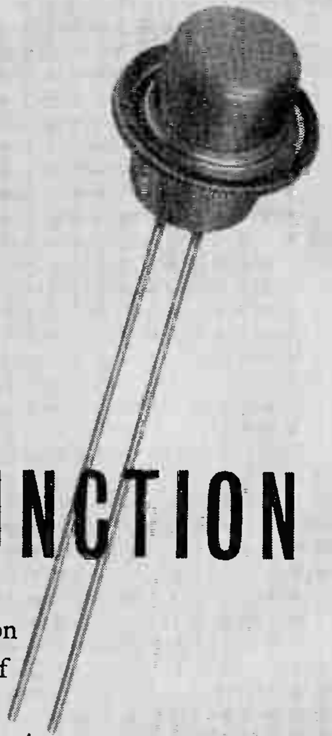


SILICON JUNCTION DIODES

For High Temperature Operation with extremely high ratios of forward to reverse resistance.

SX641 SX642 SX643 SX644

These diodes utilise a recently developed glass—pure copper seal which has made possible the production of devices with really outstanding thermal properties.

LOWER VOLTAGE TYPES
SX641 SX642

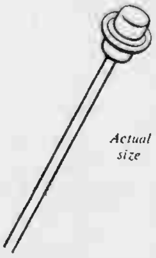
Suitable for use as Second Detectors at frequencies of up to 10 Mc/s and for the majority of other low power circuit functions including Magnetic Amplifiers.

NOW REDUCED
IN PRICE

HIGHER VOLTAGE TYPES
SX643 SX644

Suitable for use as H.T. rectifiers in telecommunication type power supplies, and for Blocking and Gating functions.

Typical ratings for capacitive input circuits at ambient temperatures of less than 75°C are given below:—



Actual size

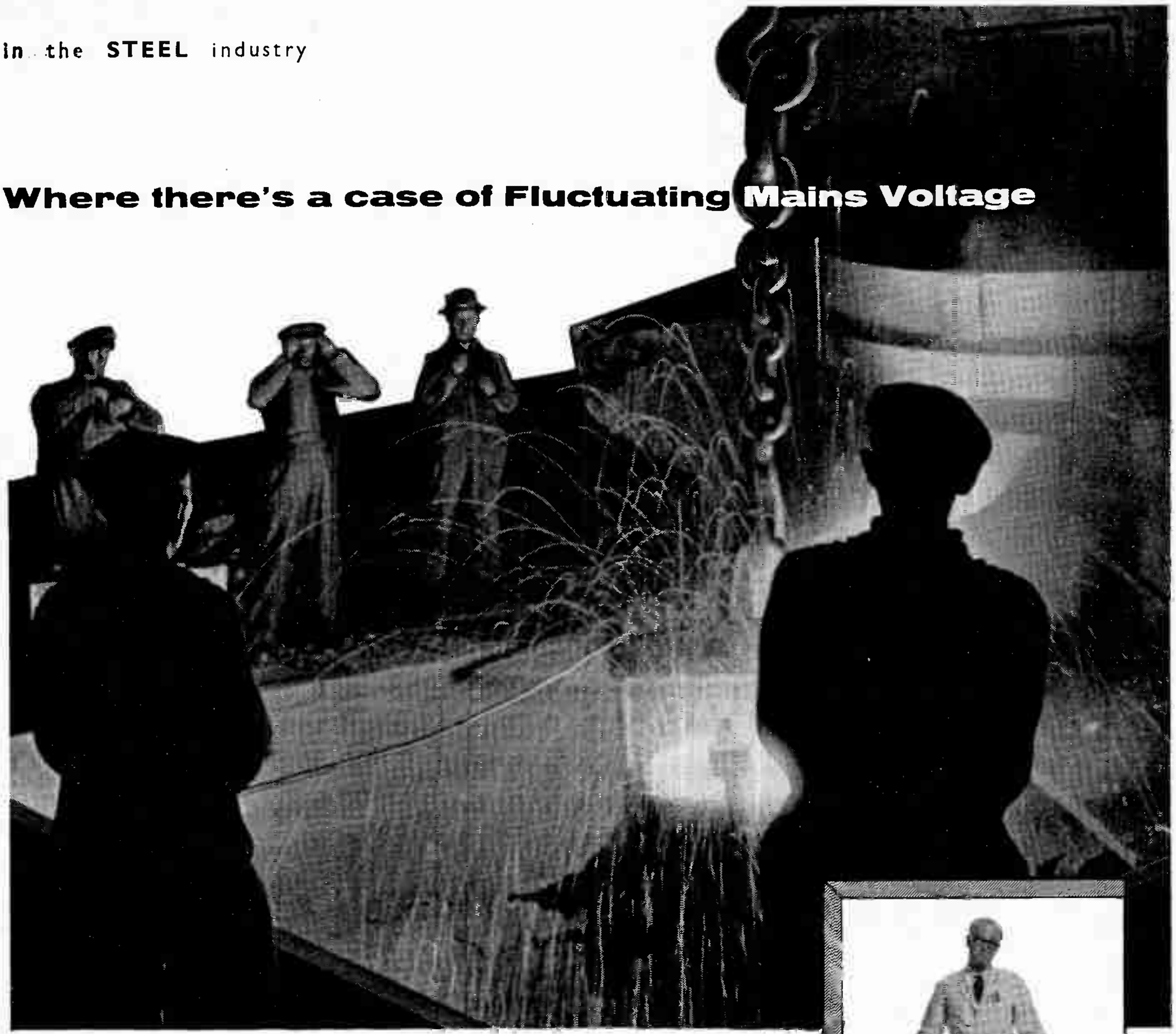
Circuit Arrangement	Number of Diodes	Max. Rectified Current (mA)	R.M.S. Input Voltage (V)		D.C. Output Voltage (V)	
			SX643	SX644	SX643	SX644
Half-wave	1	100	64	106	90	150
Bi-phase	2	200	64-0-64	106-0-106	90	150
Bridge	4	200	128	212	180	300

For further information, write to the G.E.C. Valve and Electronic Department

THE GENERAL ELECTRIC CO. LTD., MAGNET HOUSE, KINGSWAY, LONDON, W.C.2.

in the **STEEL** industry

Where there's a case of Fluctuating Mains Voltage



there's a case for *Advance*

The use of electronic control in Britain's Steel Industry has brought its reward in increased output. But it has also brought its own problem. For this vital control equipment must be protected from the a.c. voltage fluctuations caused by the switching-in and operation of heavily motorised machines, otherwise its effective life will be seriously impaired. Moreover, a voltage drop to a servo mechanism could cause chaos on the production line itself. The safeguarding of control gear and processes against supply voltage variations is one of the functions of the "Advance" Constant Voltage Transformer.

VISIT US AT THE
I.E.A.
EXHIBITION
— OLYMPIA —
16th to 25th April
STAND NO.
608

"Advance" Constant Voltage Transformers provide a.c. voltage stabilisation of ±1% for input variations of up to ±15% at maximum load. For power requirements from 4 to 6,000 watts, they are automatic and contain no moving parts.

Full technical details available in Leaflet R.28.

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CONSTANT VOLTAGE TRANSFORMERS

GD17

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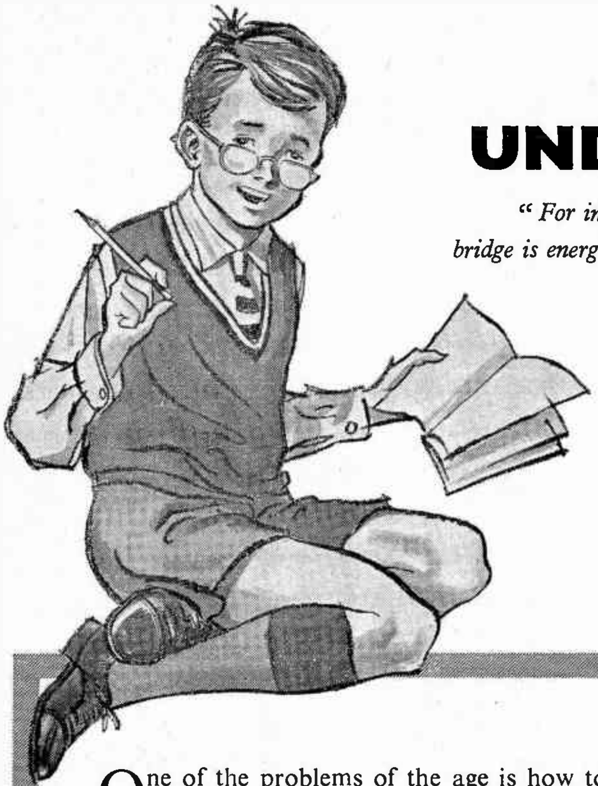

UNDOUBTEDLY . . .

" For inductance or capacitance measurements the measuring bridge is energised by the output from a valve oscillator, the out-of-balance voltage from the bridge being applied to the built-in selective amplifier-detector and moving-coil indicator, while for resistance measurements the indicator is used directly as a centre-zero galvanometer, the necessary d.c. for the bridge being derived from the power pack supplying the oscillator and detector used for inductance and capacitance measurements."

One of the problems of the age is how to keep up with the younger generation. One can support the shame of being unable to help with homework, but the prospect of having to admit ignorance of electronic equipment is intolerable. For instance, what do you know about the Marconi Universal Bridge Type TF 868A? Could you impress our young friend with an authoritative description of the instrument's ingenious mechanical design which provides single-dial measurement of L, C, and R? You owe it to yourself to be really up-to-date about Marconi Instruments—after all, they are important tools of your trade. Start with the TF 868A. Our leaflet V112 contains full details, and we'll gladly send you a copy.

UNIVERSAL BRIDGE
Type TF 868A

Measures inductance or capacitance at 1 or 10 kc/s, resistance at d.c. Measurement Ranges: 1 µH to 100 henrys, 1 µµF to 100 µF, 0.1 ohm to 10 MΩ. Q Range: 0.1 to 10 at kc/s, 1 to 100 at 10 kc/s. Tan δ Range: 0.001 to 0.1 at 1 kc/s, 0.01 to 1.0 at 10 kc/s.

MARCONI
INSTRUMENTS

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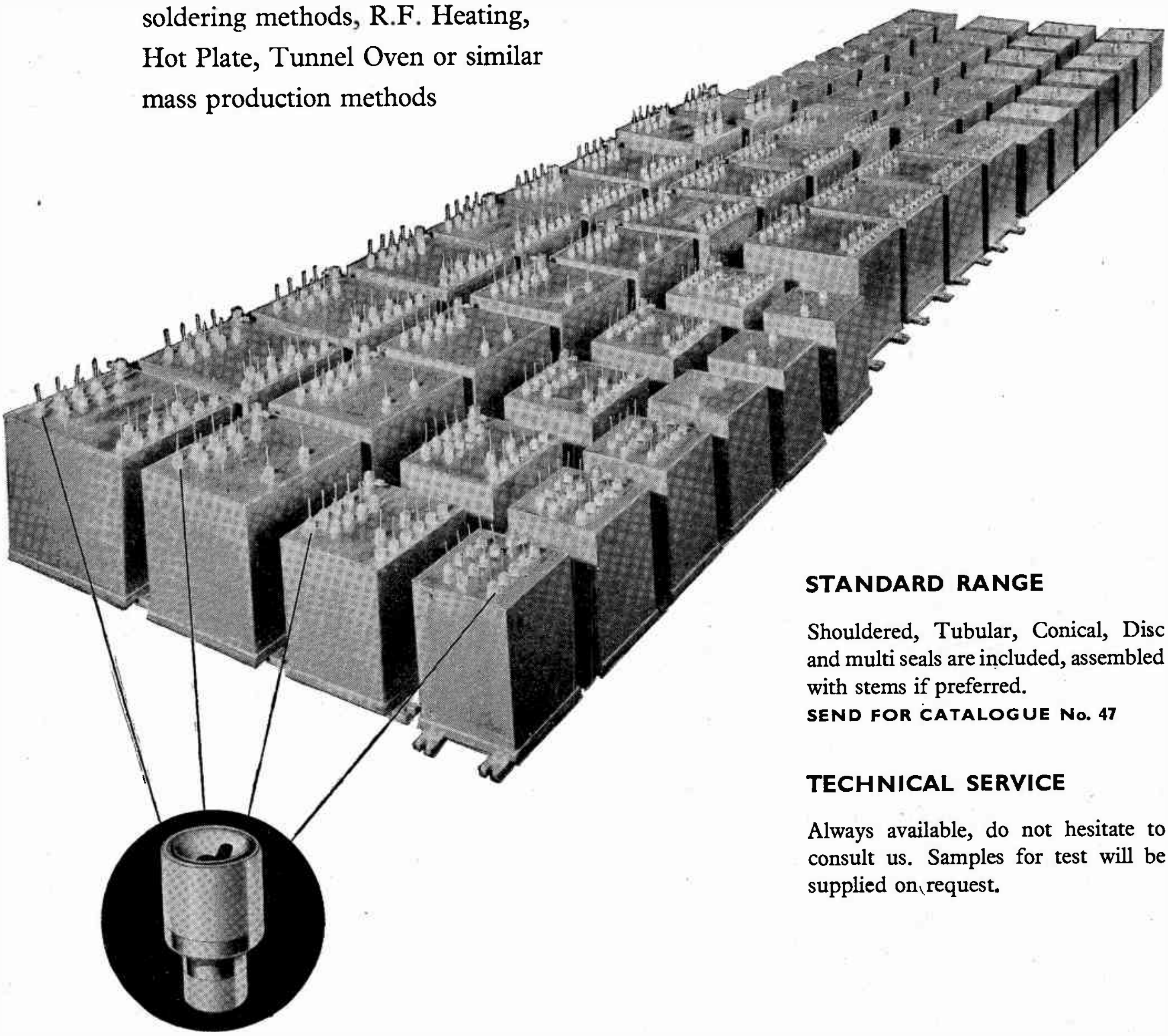

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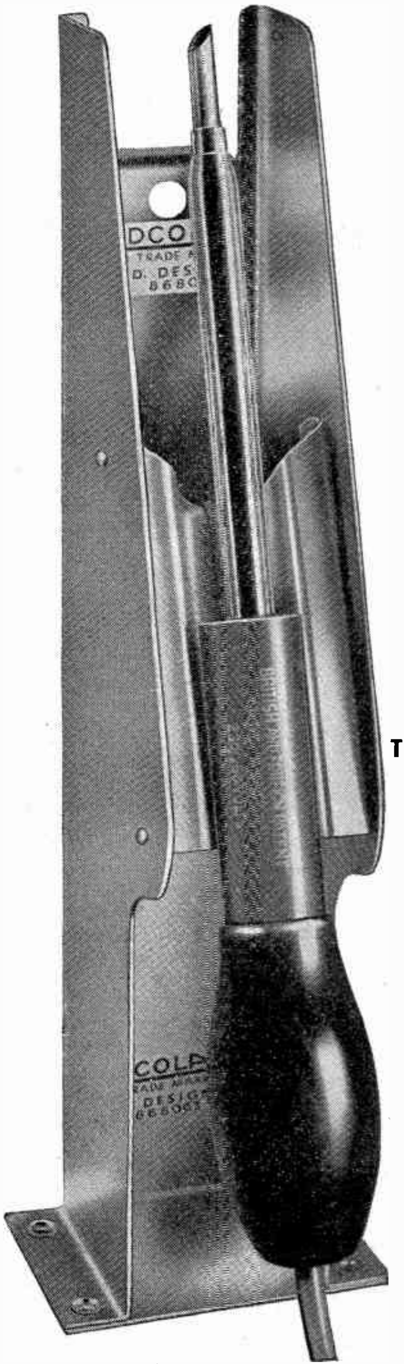
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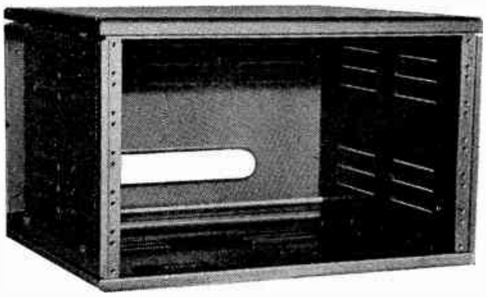

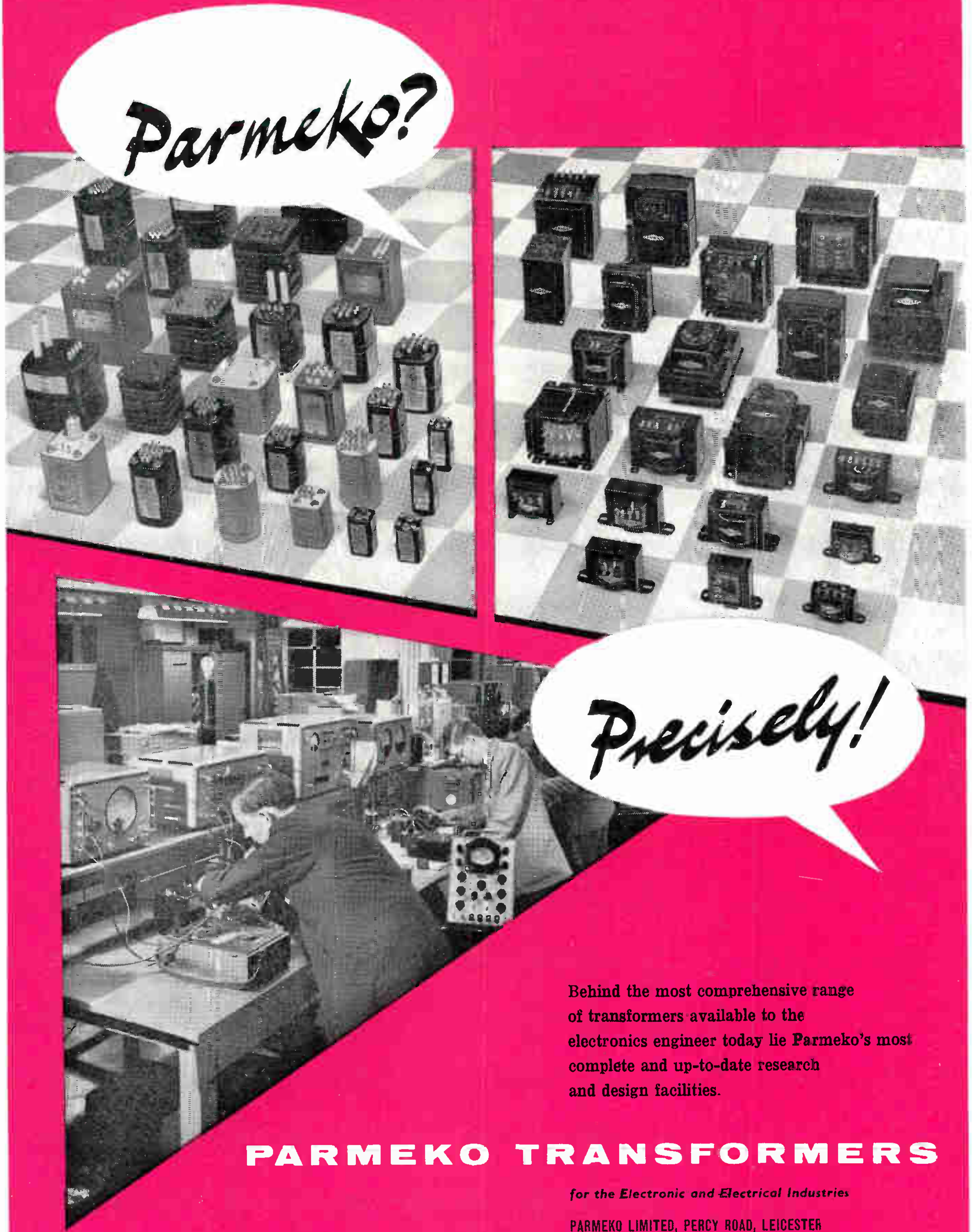
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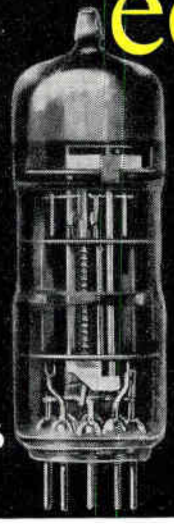

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cg′–a l l (each section)	6.4 pF
ca–a l l (each section)	1.6 pF
cout (two sections in push-pull)	0.95pF
cin (two sections in push-pull)	3.8 pF

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CHARACTERISTICS

(each section) measured at Ia = 25mA, Va = Vg″ = 150V

gm	10.5mA/V
µg′–g″	31

TYPICAL OPERATING CONDITIONS

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f	470	470	Mc/s
Va	180	180	V
Vg″	180	180	V
Ia	2 x 27.5	2 x 20	mA
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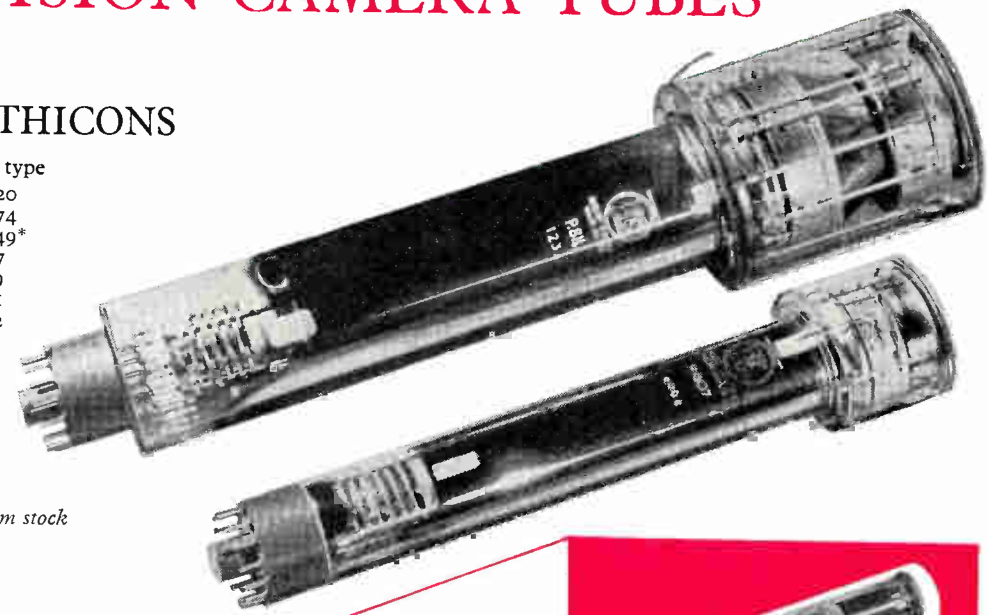

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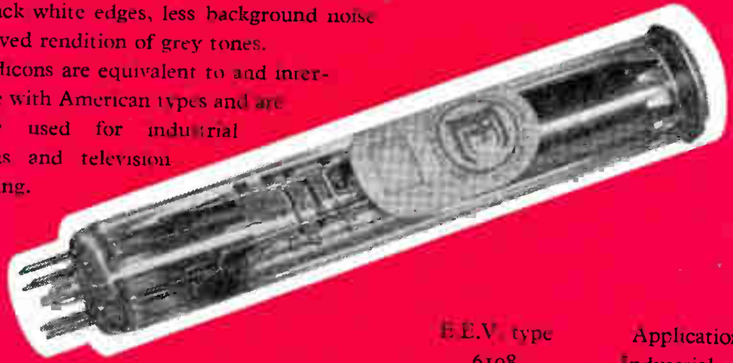

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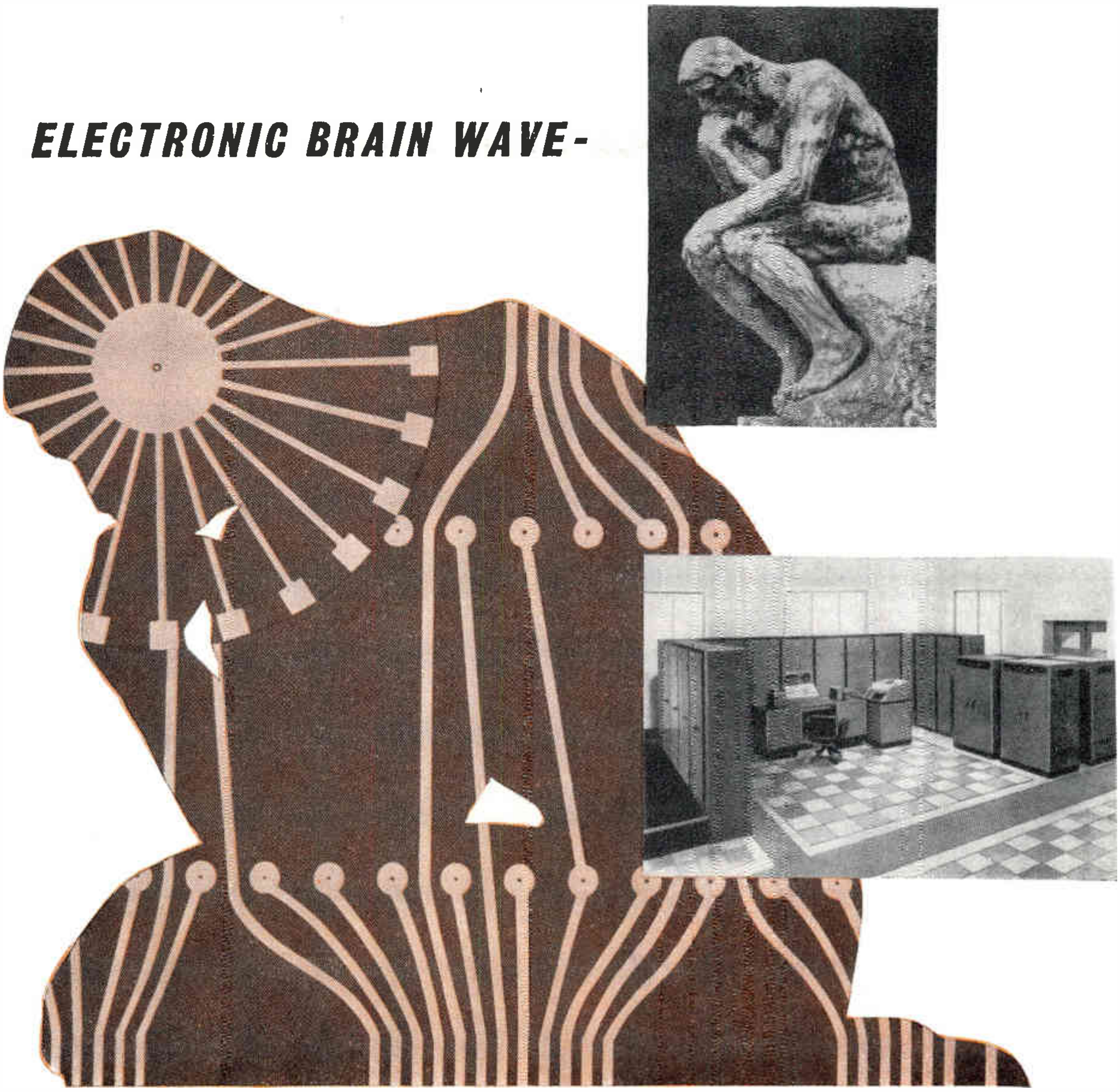

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Permanent Magnets

Design Advisory Service

The Modern Theory of Permanent Magnets

No. 1

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

This article is designed to help towards a better understanding of the changes that take place inside a permanent magnet when it is placed in a magnetic field.

According to present theory a magnetic material consists of a number of elemental magnets known as domains. If the material is unmagnetised the magnetic fields from the domains cancel out, due to their random arrangement, so that the collective field is zero.

To magnetise the material, it must be subjected to a magnetic field of sufficient intensity to rotate the domains and bring them into line with the direction of the field. This is known as magnetic saturation and is indicated in Fig. 1 as B_{sat}.

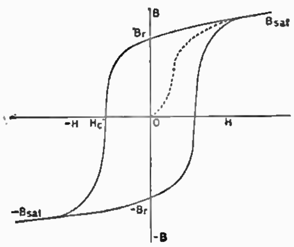

Fig 1

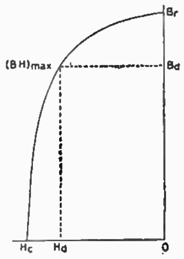

Fig 2

Fig. 1 shows the familiar hysteresis loop of a ferromagnetic material. It represents a cyclic change in magnetic flux density due to the change in magnetising force.

The area of the hysteresis loop is a measure of the energy stored in the material and for permanent magnets the loop should be as large as possible.

Point O represents a magnet material unmagnetised and the dotted curve shows the increase in magnetic flux density with the increase in magnetising force H.

Saturation, denoted B_{sat} is the degree of saturation which will produce the maximum area of hysteresis loop. When the magnetising force is reduced to zero, the magnet retains a high proportion of the flux set up and this value, which is the remanant induction in the magnet, is shown as B_r.

If the magnet is then subjected to a de-magnetising force (a magnetising force in the reverse direction) the value of flux will reduce as the de-magnetising force increases, until the flux falls to zero, at H_c. This value H_c is known as the coercive force of the magnet and is the highest value of H to which the magnet can be subjected before the flux in the magnet reverses. By increasing the de-magnetising force still further the magnet will become saturated in the reverse direction at $-B_{sat}$.

If the magnetising force is once more reversed, the flux density will follow in a manner as indicated on the graph.

The second quadrant (Fig. 2) of the hysteresis loop is the curve usually shown to indicate the characteristics of a permanent magnet. The maximum external energy obtainable from a magnet is obtained where the product of the corresponding values of B and H are maximum. These values are usually designated B_d and H_d and are the values recommended for design purposes.

Partial de-magnetisation of the magnet will occur if ferromagnetic objects come into contact with its sides. This treatment not only alters the direction of magnetisation of the domains, thus decreasing the useful flux at the pole faces, but is liable to cause instability. It is important, therefore, that ferromagnetic objects are kept away from magnetised magnets.

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ELECTRONIC & RADIO ENGINEER

VOLUME 35 NUMBER 4

APRIL 1958

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Design Problems

IT is commonly supposed that theory leads practice and that the designer can sit down with pencil, paper and slide-rule to determine what must be done to obtain a given result. Engineers know that the truth is often the other way round and that theory, so far from leading, often lags behind practice. People have often known how to do things before they knew why.

The theoretical basis of electricity and electronics is firmly established and yet in probably no other branch of engineering are empirical methods so widely used. The main reason is not that the theory is inadequate. It so happens that in parts of our field trial and error methods are so quick and easy that it is much simpler to use them than to carry out the laborious calculations which the theoretical approach would necessitate.

Although the empirical approach may thus still be used even when the general theory is well understood, there are cases when theory is lacking. A case in point is the ultra-linear amplifier which is now widely used in high-quality audio equipment. Its special character is that the pentode screen-grid is connected to a tapping point on the output transformer so that there is negative feedback to the screen-grid. Apparently due to Blumlein as far back as 1937, its operation still lacks a sound theoretical basis, and design is almost entirely experimental.

We dealt with this circuit in our Editorial for August 1955, where we endeavoured to show something of the mechanism by which the improved linearity of the circuit comes about. As far as we can find out, no one has yet worked out a method of paper design for it, whereby one can take the valve curves and work out the optimum load and transformer tapping point, and then calculate the power output and percentage distortion. If anyone has succeeded in doing this, he has kept remarkably quiet about it!

We feel that this sort of thing is unsatisfactory and it irks us that such an apparently simple circuit should lack a proper theoretical basis. Of course, like most large-signal valve problems, it is a non-linear one and, as such, does not lend itself well to an algebraic solution.

Parallel-T RC Selective Amplifiers

By **J. J. Ward** and **P. V. Landshoff** *

SUMMARY. *The most commonly used form of parallel-T selective amplifier requires that the input signal be derived at high impedance. In this article a less familiar form of the circuit, intended for use where the input signal appears at low impedance, is described. A series of equations is derived, from which the performance of the circuit as an amplifier or an oscillator can be calculated with considerable accuracy. The use of the equations for the design of a fixed-tuned amplifier is illustrated, applied to a specimen amplifier working at* 50 *c/s. Finally, brief details are given of the degree of accuracy to be expected in practice in the performance of circuits designed with the aid of the equations.*

Selective amplifiers employing parallel-T filter networks in feedback loops have been widely used at low frequencies in applications where high selectivity, compactness and low cost are important considerations. Although, with careful design and the use of magnetic ferrite cores, circuits using inductors may give comparable values of Q at high audio frequencies, the increase in the bulk and cost of the coils makes them less suitable as the frequency is reduced. Also, since such cores saturate at quite low direct currents, it is usually necessary to shunt-feed them, passing the direct current to the amplifier valve through a separate resistive path. This damps the tuned circuit, allowing only a fraction of its dynamic resistance and unloaded Q to be usefully employed. Even where shunt loading is not necessary, it is rarely possible to operate the coil at a high enough impedance level to overcome this drawback completely.

Fig. 1 shows the most commonly used circuit of a parallel-T selective amplifier. With the component relationships as shown, the network should ideally operate from a zero-impedance source at the anode and be terminated in an open circuit at its output, or grid, end. However, the valve has of necessity a finite output resistance and hence the generator impedance to the network is not, in fact, zero. Also, since the input signal is applied directly between the grid and earth, loading of the network occurs unless R_0, the input source resistance to the network, is large. As a result the Q is adversely affected and the resonant frequency is not the usual value assumed; i.e., $1/2\pi CR$. When R_0 is very small the circuit is completely useless, but there are many applications where the input signal E_{in} is derived at high impedance; in such cases, the amplifier proves quite satisfactory provided some adjustment is made in the values of the components of the filter network.

An alternative form of the parallel-T selective amplifier (Fig. 2) in which the input signal E_{in} is injected into the null point of the network and must be at low impedance will be described here. R_0 in this case may be the output impedance of a cathode-follower, or any other convenient low-impedance source provided, as will be seen below, it is less than about a tenth of the resistance R. Since the null point of the network is isolated from a d.c. point of view from the grid of the amplifier valve, it may be connected to the signal source without a blocking condenser, an important consideration when the frequency is of the order of a tenth of a cycle per second.

Analysis of the Amplifier

As a preliminary, consider the circuit of Fig. 3 (a), which is the usual constant-voltage valve equivalent circuit. The impedance Z represents the loading of the filter network on the amplifier and R_L is the effective anode load. The current through the valve is given by

$$I_a = \frac{-\mu E_g}{r_a + \dfrac{R_L Z}{R_L + Z}}$$

and the output voltage by

$$E = \frac{R_L Z}{R_L + Z} I_a = \frac{-\mu E_g \dfrac{R_L Z}{R_L + Z}}{r_a + \dfrac{R_L Z}{R_L + Z}}.$$

If we write

$$\frac{\mu R_L}{r_a + R_L} = \mathrm{A} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (1)$$

and

$$R_a = \frac{r_a R_L}{r_a + R_L} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (2)$$

so that

$$\frac{R_a}{A} = \frac{r_a}{\mu} = \frac{1}{g_m} \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \qquad (3)$$

we have

$$E = \frac{-AE_g Z}{R_a + Z}$$

From this, we see that the circuit of Fig. 3 (a) may be redrawn as in Fig. 3 (b) and the full equivalent circuit

---

* *Cinema-Television, Ltd.*

of the amplifier is as in Fig. 4. Applying Kirchhoff's laws to the various meshes of this circuit and writing $E/E_{in} = -a$, we have

$$E = -aE_{in} = -AE_g - I_1R_a \quad .. \quad .. \quad (4)$$

Also

$$E_g = E_{in} + I_1R_0 + I_2k/j\omega C + I_3(k/j\omega C + R/n)$$

and

$$-AE_g = E_{in} + I_1(R_a + R + R_0 + 1/j\omega nC) + I_2R - I_3/j\omega nC$$

or, using (4),

$$0 = E_{in}(1 - a/A) + I_1(R_a/A + R_0) + I_2k/j\omega C + I_3(k/j\omega C + R/n) \quad .. \quad .. \quad .. \quad (5)$$

and

$$0 = E_{in}(1 + a) + I_1(R + R_0 + 1/j\omega nC) + I_2R - I_3/j\omega nC \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

In addition

$$0 = I_1R + I_2(R + 1/j\omega C)(1 + k) + I_3k(R + 1/j\omega C) \quad .. \quad .. \quad .. \quad .. \quad (7)$$

and

$$0 = I_1/j\omega nC + I_2k(R + 1/j\omega C) + I_3(k + 1/n)(R + 1/j\omega C) \quad .. \quad .. \quad .. \quad (8)$$

Eliminating I_1, I_2, I_3 and E_{in} from equations (5) to (8) and replacing R_a/A by $1/g_m$ (equation 3), we have

$$0 = \begin{vmatrix} 1 - a/A & 1/g_m + R_0 & k/j\omega C & R/n + k/j\omega C \\ 1 + a & R + R_0 + 1/j\omega nC & R & -1/j\omega nC \\ 0 & R & (1 + k)(R + 1/j\omega C) & k(R + 1/j\omega C) \\ 0 & -1/j\omega nC & k(R + 1/j\omega C) & (k + 1/n)(R + 1/j\omega C) \end{vmatrix}$$

$$\therefore \left[1 - a/A\right]\left[(R + R_0 + 1/j\omega nC)\left(\frac{1 + k}{n} + k\right)\right.$$
$$(R + 1/j\omega C) - R(kR + R/n + k/j\omega nC)$$
$$\left.-\frac{1}{j\omega nC}\left(kR + \frac{1 + k}{j\omega nC}\right)\right]$$
$$-\left[1 + a\right]\left[(R_0 + 1/g_m)\left(\frac{1 + k}{n} + k\right)\right.$$

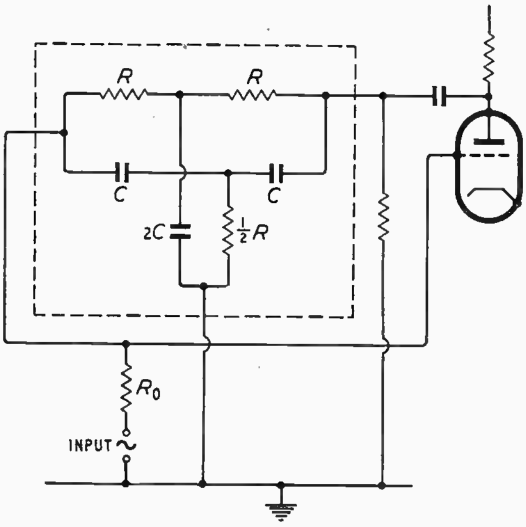


Fig. 1. *Parallel-T selective amplifier*

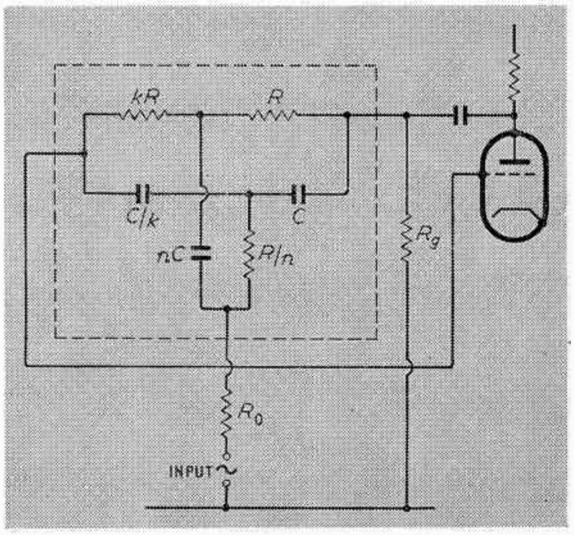


Fig. 2. *Parallel-T amplifier with signal input at the null point*

$$(R + 1/j\omega C) - \frac{k}{j\omega C}(kR + R/n + k/j\omega nC)$$
$$\left.+\left(R/n + k/j\omega C\right)\left(kR + \frac{1 + k}{j\omega nC}\right)\right] = 0 \quad .. \quad (9)$$

If we make a real and equal to a_0, say, then equating the real part of this to zero we obtain the value of ω which makes the output exactly out of phase with the input. Thus

$$\left[1 - a_0/A\right]\left[RR_0\left(\frac{1 + k}{n} + k\right) + kR^2/n - k/\omega^2{}_0C^2n\right]$$
$$-\left[1 + a_0\right]\left[R(1/g_m + R_0)\left(\frac{1 + k}{n} + k\right) + kR^2/n\right.$$
$$\left.- k/\omega^2{}_0C^2n\right] = 0$$

or $\omega_0 =$

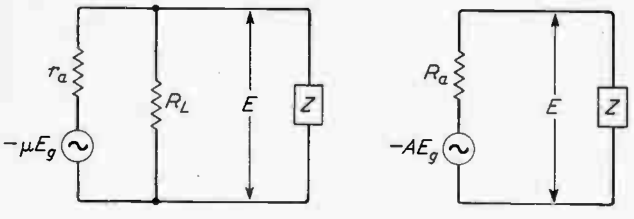
$$\frac{1}{CR}\sqrt{\frac{1}{1 + \frac{1}{R}\left(\frac{1 + k}{k} + n\right)\left(R_0 + 1/g_m \cdot \frac{1 + a_0}{a_0(1 + 1/A)}\right)}} \quad .. \quad .. \quad (10)$$

where ω_0 is the corresponding value of ω. If, as is usually the case, $1/g_m$ is negligible compared with R this is also the frequency at which maximum voltage gain occurs. Since A and a_0 are, in general, large compared with unity, we may write approximately

$$\omega_0 = \frac{1}{CR}\sqrt{\frac{1}{1 + h(R_0 + 1/g_m)/R}} \quad .. \quad .. \quad (10a)$$

where $h = \dfrac{1 + k}{k} + n$. Note that this is, in fact, the exact value of ω_0 when $a_0 = A$.

Now equating the imaginary part of Equ. (9) to zero, we obtain



(a) (b)

Fig. 3. *(a) Conventional equivalent circuit; (b) modified equivalent circuit*

$$\left[1 - a_0/A\right]\left[\left(\frac{1+k}{n} + k\right)\left(R_0 + R + R/n\right) - 2kR/n\right]$$
$$- \left[1 + a_0\right]\left[\left(\frac{1+k}{n} + k\right)\left(R_0 + R/n + 1/g_m\right)\right.$$
$$\left. - 2kR/n\right] = 0$$

This gives a quadratic equation from which we may calculate the value of n required to obtain a given a_0:

$$n^2k[a_0(1 + 1/A)R_0 + (1 + a_0)1/g_m - R(1 - a_0/A)]$$
$$- n[(1 - a_0/A)\{(R + R_0)(1 + k) - kR\} - (1 + a_0)$$
$$\{(R_0 + 1/g_m)(1 + k) - kR\}] + a_0(1 + 1/A)(1 + k)R$$
$$= 0 \quad .. \quad .. \quad (11)$$

To achieve a high Q the highest possible value of a_0 consistent with stability is required. Although a limited degree of regeneration may be employed with little risk to raise a_0, it is generally more advisable to make the circuit neither regenerative nor degenerative, that is put $a_0 = A$. Equation (11) then becomes

$$n^2k(R_0 + 1/g_m) - n[kR - (R_0 + 1/g_m)(1 + k)]$$
$$+ (1 + k)R = 0 \quad .. \quad .. \quad .. \quad .. \quad (11a)$$

This equation does not have a real, positive root for n unless R_0 and $1/g_m$ lie below a certain limit. In fact for n to be real

$$[k - (1 + k)(R_0/R + 1/g_mR)]^2$$
$$- 4k(1 + k)(R_0/R + 1/g_mR) \geqslant 0$$

or

$$\left[R_0/R + 1/g_mR - (3 + \sqrt{8})\frac{k}{1+k}\right]$$
$$\left[R_0/R + 1/g_mR - (3 - \sqrt{8})\frac{k}{1+k}\right] \geqslant 0$$

Since the product of the roots of Equ. (11a) is positive, the two roots must be of the same sign. Hence for n to be positive, their sum must be positive, or

$$kR - (R_0 + 1/g_m)(1 + k) \geqslant 0$$

Combining the two inequalities we have

$$R_0 \leqslant \frac{3 - \sqrt{8}}{1 + k}kR - \frac{1}{g_m} \quad .. \quad .. \quad .. \quad (12)$$

The *Q* of the Circuit

We now derive a formula for the Q of the circuit, which will be approximate in that we shall assume that $1/g_m$ is negligible compared with R so that the frequency of maximum gain coincides with the frequency of zero power factor as given in Equ. (10).
We write

$$x = RR_0\left(\frac{1+k}{n} + k\right) + kR^2/n - k/\omega^2C^2n$$
$$t = R(R_0 + 1/g_m)\left(\frac{1+k}{n} + k\right) + kR^2/n - k/\omega^2C^2n$$
$$y = \left(\frac{1+k}{n} + k\right)(R + R_0 + R/n) - 2kR/n$$
$$z = \left(\frac{1+k}{n} + k\right)(R_0 + 1/g_m + R/n) - 2kR/n$$

Then Equ. (9) is

$$(1 - a/A)(x + y/j\omega C) - (1 + a)(t + z/j\omega C) = 0$$
$$.. \quad .. \quad (9a)$$

and the conditions obtained by putting $a = a_0$ and equating the real and imaginary parts of (9) to zero are

$$a_0(x_0/A + t_0) = x_0 - t_0 = x - t$$

and

$$a_0(y/A + z) = y - z,$$

where x_0 and t_0 are the corresponding values of x and t, y and z being independent of frequency.

Now $(x/A + t) = b - (1 + 1/A)k/\omega^2C^2n$, where b is independent of ω and is, in fact, equal to

$$[1 + 1/A]\left[RR_0\left(\frac{1+k}{n} + k\right) + kR^2/n\right]$$
$$+ \frac{R}{g_m}\left(\frac{1+k}{n} + k\right),$$

and so

$$(x/A+t)-(x_0/A+t_0)=[(1+1/A)k/C^2n](1/\omega^2{}_0-1/\omega^2)$$

or

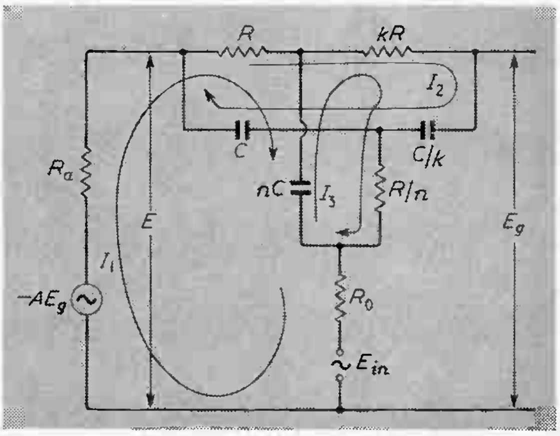
$$(x/A+t)=[(1+1/A)k/C^2n][1/\omega^2{}_0-1/\omega^2)]+(x-t)/a_0$$

Also $x - t = khR/ng_m$ and $y - z = (R - 1/g_m)kh/n$, where $h = \left(\frac{1+k}{k}\right) + n$ as before.

Thus

$$|a|^2 = \frac{\omega^2C^2(x-t)^2 + (y-z)^2}{\omega^2C^2(x/A + t)^2 + (y/A + z)^2} \text{ from } (9)$$

Fig. 4. *Complete equivalent circuit used in the analysis*



$$= \frac{[\omega^2C^2/g^2_m + (1 - 1/g_mR)^2]k^2h^2R^2/n^2}{\omega^2C^2[(1 + 1/A)(1/\omega^2_0 - 1/\omega^2)k/nC^2 + khR/ng_ma_0]^2 + \dfrac{k^2h^2(R - 1/g_m)^2}{n^2a^2_0}}$$

Now, if we assume that $1/g_m$ is negligible compared with R and $R\omega C$ is approximately unity, then $\omega C/g_m$ is negligible and this expression becomes

$$|a|^2 = \frac{h^2R^2}{(1 + 1/A)^2(1/\omega^2_0 - 1/\omega^2)^2\omega^2/C^2 + h^2R^2/a^2_0}$$

We see at once that this has a maximum value when $\omega = \omega_0$, justifying our earlier statements.

We now find an equation for ω corresponding to a gain of 3 dB below that at resonance. Putting $|a|^2 = \frac{1}{2}a^2_0$ and taking the square root we obtain the two quadratic equations

$$\omega^2(1 + 1/A)/\omega^2_0 \pm \omega hCR/a_0 - (1 + 1/A) = 0$$

Each of these two equations has one positive and one negative root. If the two positive roots are ω_1 and ω_2, the 3-dB bandwidth of the amplifier is

$$\omega_1 - \omega_2 = \frac{RC\omega^2_0h}{a_0(1 + 1/A)}$$

and so using Equu. (10a)

$$Q = \frac{\omega_0}{(\omega_1 - \omega_2)} = \frac{a_0}{h}(1 + 1/A)\sqrt{1 + h(1/g_m + R_0)/R} \quad \ldots \quad \ldots \quad (13)$$

Hence slightly more approximately

$$Q = a_0/h \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13a)$$

with an error of no more than about 10 per cent in most practical cases. This formula indicates that, to give a reasonable selectivity, the lower of the two roots for n given by Equ. (11) must be used. Also for given a_0 a small increase in Q is obtained by making k greater than one, which is the value usually used in RC networks of this type. Making k less than one provides a useful method for reducing the Q without lowering the gain. Now

$$a_0 = \frac{y - z}{y/A + z}$$

$$= \frac{h(R - 1/g_m)}{h[1/g_m + R/A + (1 + 1/A)(R_0 + R/n)] - 2R(1 + 1/A)} \quad \ldots \quad \ldots \quad (14)$$

from which we find that increasing n above the value given by the lower root of Equ. (11) increases a_0 and so also Q.

The Circuit Used as an Oscillator

In certain applications it may be useful to make a_0 somewhat greater than A by making the circuit slightly regenerative. The useful limit to this increase in a_0 occurs before the amplifier becomes unstable and both a_0 and Q tend to infinity. Dividing Equ. (11) by a_0 and letting it tend to infinity we have, in fact, an equation giving the value of n for which oscillation sets in:

$$n^2k\left(R_0 + 1/g_m + \frac{R_0 + R}{A}\right)$$
$$+n[R_0(1+1/A)(1+k)+(1+k)/g_m-R(k-1/A)]$$
$$+ R(1 + 1/A)(1 + k) = 0 \quad \ldots \quad \ldots \quad (15)$$

The circuit makes a reliable and stable oscillator from which excellent waveforms can be obtained. It does, however, in common with the amplifier, suffer from the disadvantage that the operating frequency cannot be conveniently varied over a wide range. When the circuit is intended for use as an oscillator, R_0 is omitted. If $1/g_m$ is negligible compared with R, Equ. (15) then becomes $n^2k(1/g_m + R/A) - nkR(1 + 1/A) + R(1 + k)(1 + 1/A) = 0$. For this to give a real value of n

$$A \geqslant \frac{4 + 3k}{k - 4(1 + k)/g_mR} \approx \frac{4}{k} + 3.$$

Thus, even a small general-purpose triode can be made to oscillate quite vigorously. The frequency of oscillation is found by letting a_0 tend to infinity in Equ. (10):

$$\omega_0 = \frac{1}{CR}\sqrt{\frac{1}{1 + \dfrac{h}{R}\left[R_0 + \dfrac{1}{g_m(1 + 1/A)}\right]}}$$

If A is much greater than unity this is approximately the same as Equ. (10a).

The Design of a Typical Amplifier

In order to demonstrate the use of the equations in the design of a selective amplifier we shall apply them to a special case. Suppose that it is desired to derive a 50-c/s sine wave of at least 100 volts peak-to-peak by selecting the fifth harmonic from a 10-c/s square wave. This square wave will usually come from the anode of a valve and then it will be convenient to attenuate the input signal to the selective amplifier to a suitable degree by dividing the anode load of this valve into two parts and connecting the junction directly to the null point of the filter. If the upper and lower components of the anode load are R_1 and R_2 respectively and the anode impedance of the valve is r_a, we must then put R_0 equal to $R_1(R_2 + r_a)/(R_1 + R_2 + r_a)$.

The Q of the selective amplifier should be made as high as possible, since as much rejection as possible is required at all frequencies but the fifth harmonic. To achieve this, a high-slope pentode should be used for the amplifier; such a valve could well have a gain of around 300 and a g_m of, say, 3·5 mA/V. If we make a_0 equal to A, to obtain the required output we require a fifth harmonic input of one-third of a volt peak-to-peak, which means a fundamental of 1·3 volts. It is reasonable to assume a current of half a milliamp in the valve producing the square wave, so that an R_0 of 3 kΩ will give an ample input.

We next choose the value of k. We must consider the fact that the grid leak must be sufficiently large to avoid loading the anode of the pentode and so reducing its gain, but that the total resistance in the grid, which must be below a certain value, is equal to the grid leak together with R and kR. This consideration is particularly important at very low frequencies when R is large. At high frequencies k is also limited, now because the component C/k must not become small enough to be swamped by stray capacitances. It is necessary to reach a compromise between these considerations and the fact that, as noted above, the Q is increased when k is taken up to about five.

In our example we shall make the grid leak 470 kΩ, $R = 180$ kΩ and $k = 2$. We could have made k larger and R smaller, but this would have made C larger, which is inadvisable, since close-tolerance components are required. With the values we have chosen we find from Equ. (11a) that

$$n = 1{\cdot}589 \text{ or } 51{\cdot}69.$$

As explained above, we discard the larger root, and then we find from Equ. (10a) that the value of C required is 0·01721μF. The corresponding values of the components C/k and nC are 0·00860μF and 0·02735μF respectively, and R/n is 113·3 kΩ. Equ. (13a) tells us that Q is about 100.

The Accuracy of the Calculations

As may be seen from Equ. (10) the operating frequency of the amplifier is hardly affected by small changes in the gain or in the g_m of the valve. Hence, if high-stability components are used it will generally be found unnecessary to make provision for adjusting the resonant frequency of the circuit. This has been verified experimentally on a large number of amplifiers and oscillators whose operating frequency varied between one-tenth of a cycle per second and thirty kilocycles per second. Design calculations were made using five-figure logarithms and careful bridging and selection of components ensured that they were within one-fifth per cent of the required value. On measurement it was found that the circuit is extremely predictable, errors in the resonant frequency always being less than one-third per cent. Should it be desired to make a small variation in frequency, however, probably the best method at higher frequencies is to include a trimmer in the component C/k; when the latter is large, as it will be at low frequencies, part of the component R should be made variable instead.

As may be predicted from Equ. (14) errors due to inaccuracies in the shunt components nC and R/n were found to have a slightly more serious effect, particularly for values of a_0 somewhat greater than A. In order to obtain the exact value of a_0 desired it is probably best to provide a means of varying A, such as a variable resistance in the cathode of the valve. This will also compensate for any inaccuracy in the value of R_0.

Subjective Sharpness of Television Pictures

By W. N. Sproson, M.A.*

SUMMARY. *The subjective sharpness of television pictures has been measured using a comparison technique and a multi-criterion scale for assessment. Two types of degrading network were used and the subjective sensitivity to changes in equivalent rectangular bandwidth has been evaluated for both static and moving pictures.*

This investigation was conducted in order to determine the subjective effect of reducing the definition of a 405-line television picture. The reduction of definition was achieved by causing the response of the picture-originating source to diminish with increasing frequency. This was done with either a sine-squared network[1] or a simple resistance-capacitance circuit. In the former case, the effect is to cause a reduction of amplitude of the higher-frequency components of the video waveform, without phase distortion. In the second case, the amplitude and phase characteristics of the signal spectrum are both modified. The reason for selecting circuits which give a gradual loss of amplitude with increase of frequency is that this simulates the performance of a lens in respect of resolution, and the ultimate aim of these experiments is to produce a 'figure of merit' for lenses when used as part of a television channel.

Apparatus

The picture-producing device was a 35-mm flying-spot film scanner[2], which was used for producing moving pictures and also single frames of still pictures. The resolution of this picture source is very good in that the 3-Mc/s bars on Test Card 'C' are reproduced with no visible loss of modulation. A laboratory monitor using a 15-in. (38-cm) cathode-ray tube was used as the display device. The response of the monitor has been measured and is within 0·5 dB of uniformity from zero frequency to 6 Mc/s. The picture-degrading networks were introduced before the gamma-corrector circuits in the film scanner, so that their effect should be analogous to that produced by a lens, except that an electrical low-pass

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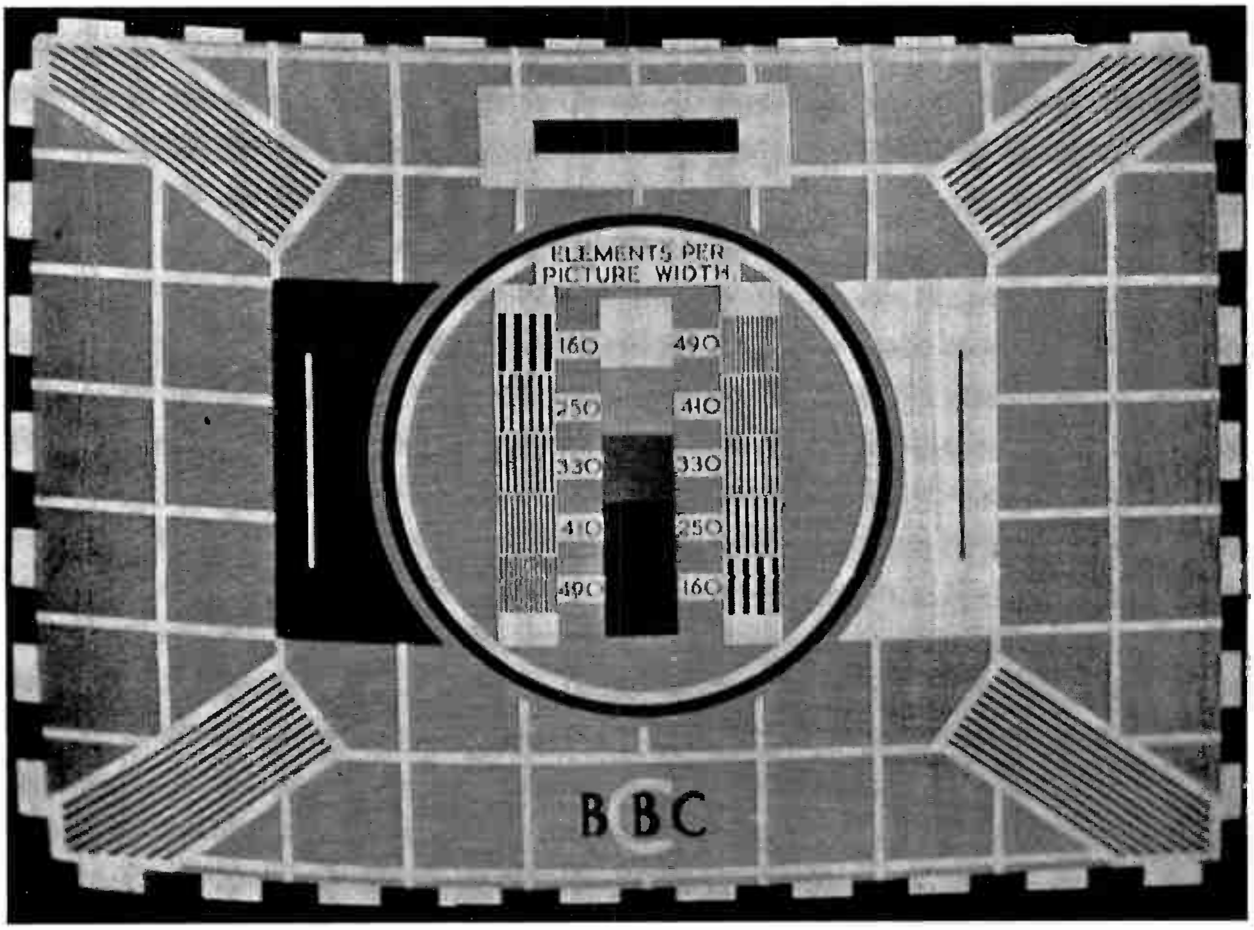
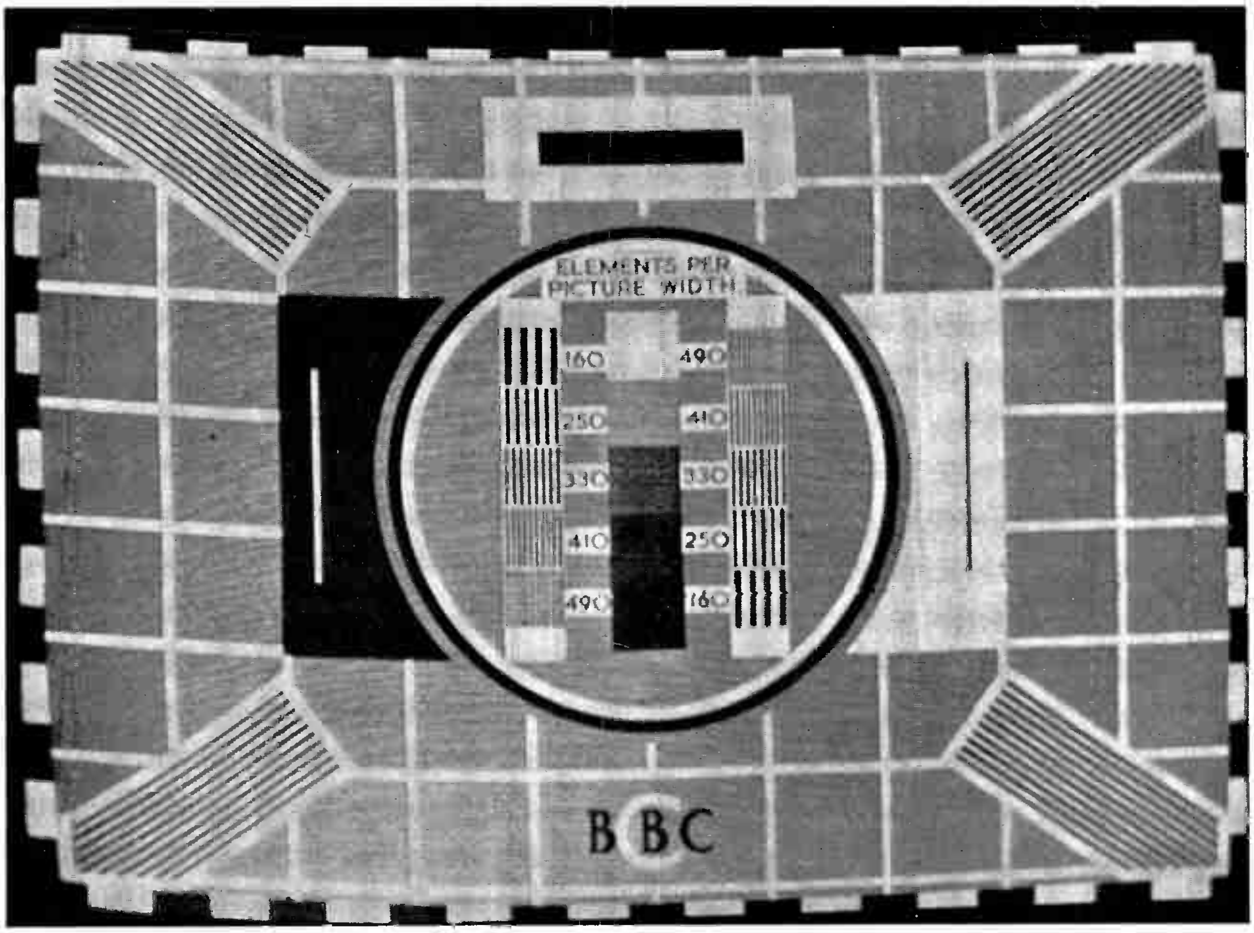


Fig. 1 (*above*). *Test card 'C' without degrading network* (*i.e., Network No.* 0). Fig. 2 (*below*). *With sine-squared network No.* 4



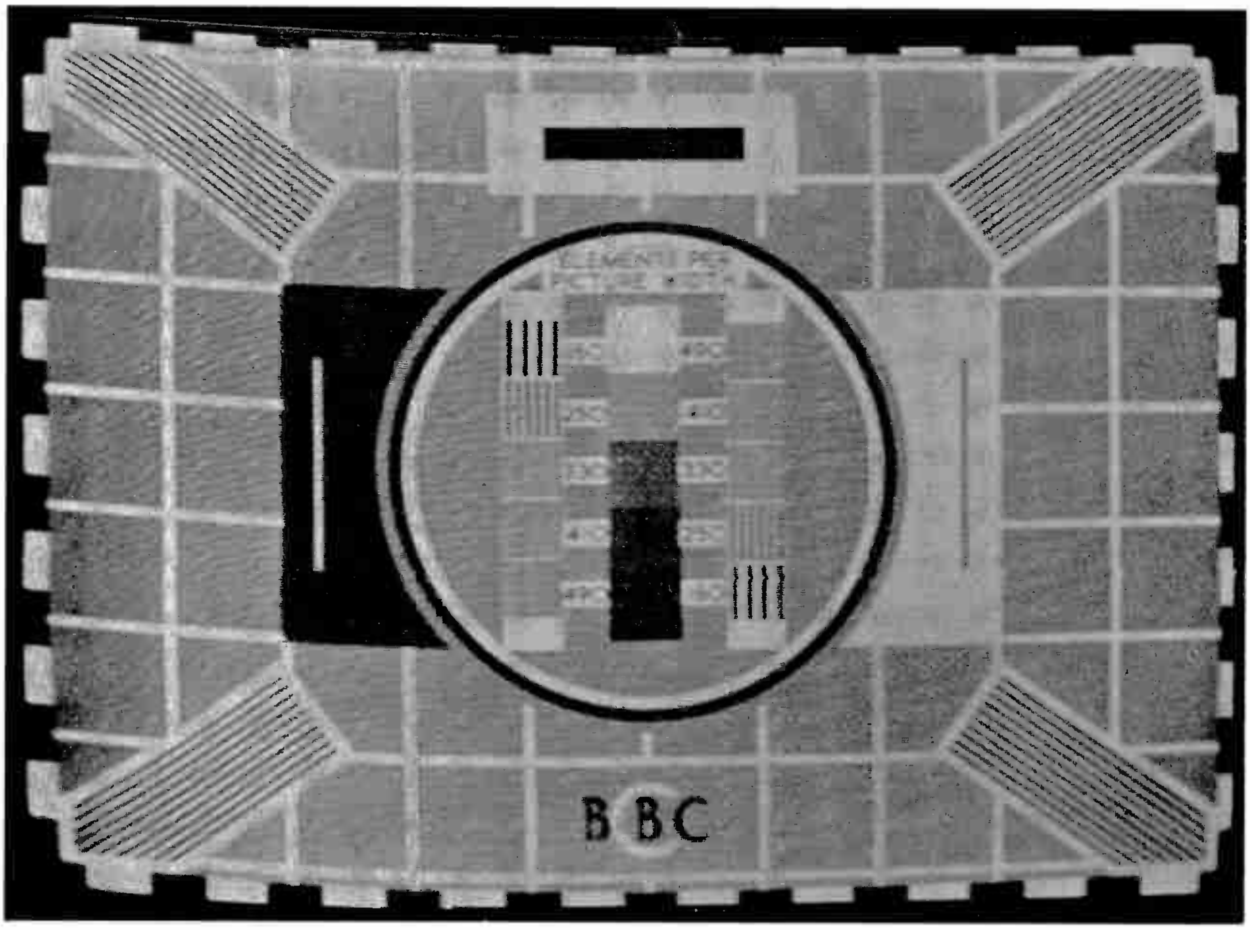
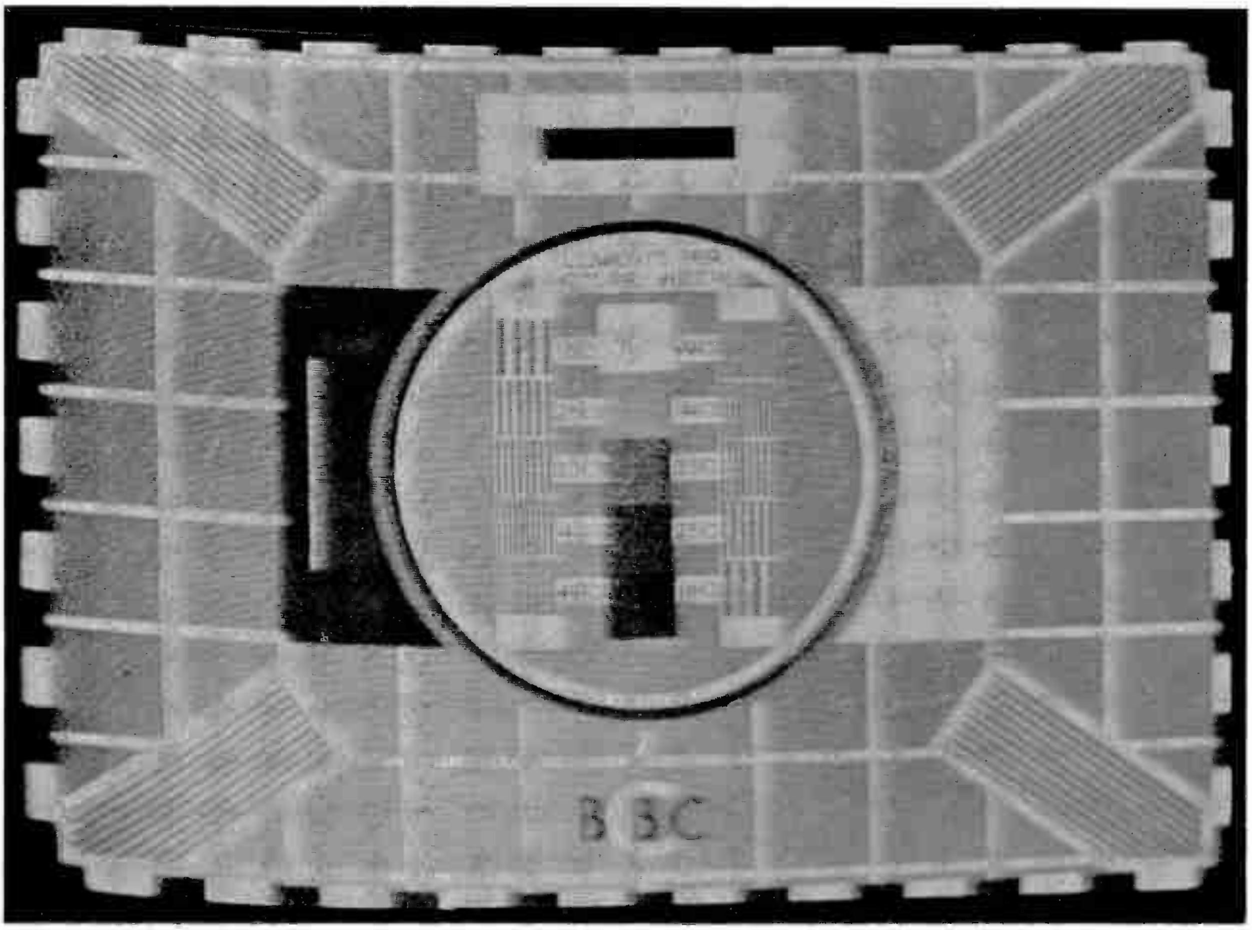


Fig. 3 (*above*). *Test Card 'C' with sine-squared network No.* 7. Fig. 4 (*below*). *With* RC *network No.* 9



filter uniformly degrades the resolution in the direction of the scanning lines, whereas a lens affects the resolution also in the vertical direction, and varies in its effect over the field. Figs. 1 to 4 illustrate the effects of these degrading networks on Test Card 'C'. The amplitude/frequency characteristics of the sine-squared networks are shown in Fig. 5 and the characteristics of the *RC* networks are given in Fig. 6.

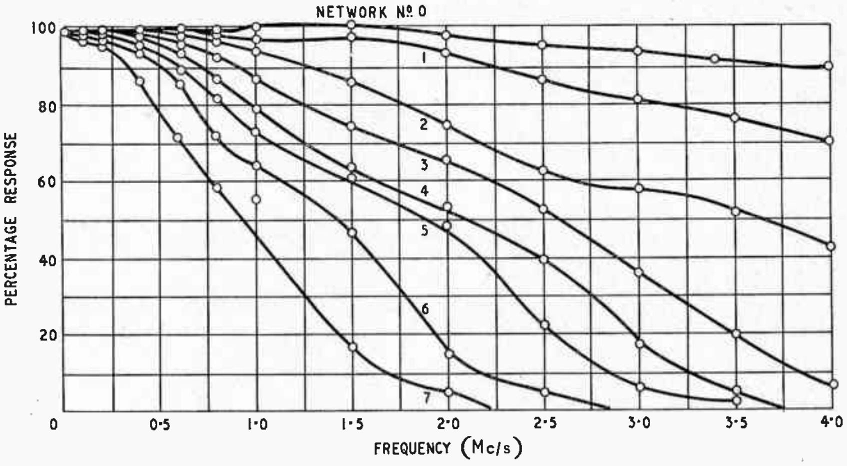


Fig. 5. *Amplitude/frequency responses of sine-squared networks*

It must be emphasized, however, that the illustrations can give only an approximate representation of the effect seen on the monitor screen, because of the errors of reproduction in the photographic and printing processes. Further, the material used in the tests was always a picture and never a test card. Nevertheless, the photographs will serve to give some idea of the effect on picture quality by the introduction of three of the sixteen networks used in the experiments.

Method

Five observers were seated in the arc of a circle at a distance of approximately four times picture height from the display monitor. The judgements were made in terms of a standard presentation of a picture which was not necessarily the best or the worst that could be produced. In every case the standard presentation of the picture was shown first and then an unknown presentation of the same picture, the observers being asked to state their opinion of the unknown in terms of the standard. Seven categories were allowed, including the category 'about the same as'. A copy of the questionnaire is given in Appendix 1. Seven different presentations were shown for a given standard presentation which was repeated each time; the observers were not expected to remember the sharpness of the standard. One session with five observers included four sets of comparisons (seven presentations in each) and thus involved twenty-eight judgements for each observer. The pictures were presented in random order and a typical schedule is shown in Table 1.

TABLE 2

Category	Much Sharper	Sharper	Slightly Sharper	About the Same	Slightly Less Sharp	Less Sharp	Much Less Sharp
Score	3	2	1	0	−1	−2	−3

The results were evaluated by allocating the scores given in Table 2.

Results

Static Pictures

A typical result for one picture and three sessions (i.e., fifteen observers in all) is shown in Fig. 7. This relates to the assessment of a static picture (street scene with building) in which the definition of the picture was changed by the use of sine-squared networks. The abscissa of the graph is the area under the response curve from zero frequency to 3 Mc/s and is proportional to the steepness at the point of 50% response with which a phase-linear system responds to a unit step. Alternatively, we can regard the abscissa as the equivalent rectangular bandwidth,* normalized to a 3 Mc/s-wide rectangle. During each set of seven judgements the standard picture was also shown as an unknown and this should give rise to zero score. Curve 1 of Fig. 7 crosses the axis at 0·83, which is, in fact, the sharpness factor of the standard picture. Curve 2 of Fig. 7 does not quite satisfy the condition, as there is a small negative score for the standard picture when used as an unknown. Both curves

TABLE 1

Presentation Number	Set 1 *RC* Networks		Presentation Number	Set 2 Sin^2 Networks		Presentation Number	Set 3 *RC* Networks		Presentation Number	Set 4 Sin^2 Networks	
	Standard	Unknown		Standard	Unknown		Standard	Unknown		Standard	Unknown
1	3	6	1	5	4	1	6	8	1	2	6
2	3	0	2	5	7	2	6	5	2	2	3
3	3	5	3	5	5	3	6	6	3	2	1
4	3	4	4	5	1	4	6	3	4	2	4
5	3	1	5	5	3	5	6	9	5	2	5
6	3	2	6	5	2	6	6	7	6	2	0
7	3	3	7	5	6	7	6	4	7	2	2

The numbering of the networks is the same as shown in Figs. 5 and 6.

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* This is not the same as Schade's equivalent pass-band. Schade takes the area under the curve of squared ordinates. See National Bureau of Standards Circular 526, pp. 231–49.

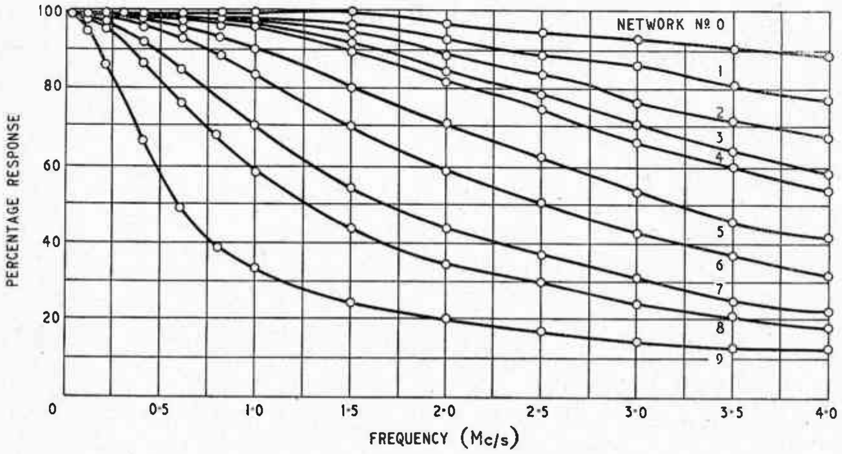

Fig. 6. *Amplitude/frequency responses of resistance-capacitance networks*

in Fig. 7 show that the arithmetical sum of the scores corresponding to the judgements of the observers gave a close approximation to a straight line when assessed in this manner. This is not entirely surprising, as previous workers[3] have found that there is good correlation between the subjective sharpness of a picture and the gradient of the response to a unit step at the mid-point of the transition. The correlation shown in Fig. 7 shows that television images are in the same category as optically formed images. The data from Fig. 7 enable us to quote a loss of bandwidth which is liminal in its effect; i.e., a change which 50% of the observers can perceive and of which 50% are unaware. Another effect which is apparent from Fig. 7 is that the slope of the curve relating score to equivalent bandwidth is a function of the sharpness of the standard comparison picture. Observers can more easily perceive a given change of equivalent bandwidth when the comparison picture is a relatively poor one; the slope of curve 2 is higher than that of curve 1. This effect was substantiated in all the tests undertaken. The gradients of the best* straight lines from several tests are shown in Fig. 8. Each point on the graph is the slope of a straight line from the results of three sessions (i.e., fifteen observers in all). The picture used for these tests was the street scene with buildings. In the case of a sine-squared degradation the gradient for a fully-resolved 3-Mc/s picture (unit sharpness factor) is 9·2 liminal units per unit change in sharpness factor or, more significantly, 0·109 change in sharpness factor per liminal unit. This is caused by a change in equivalent rectangular bandwidth from 3·0 Mc/s to 2·67 Mc/s.

In the case of a resistance-capacitance network, the gradient at all points is greater and for a fully-resolved 3-Mc/s picture its value is 13·5. This means that one liminal unit is produced when a 3-Mc/s picture is degraded to one with an equivalent rectangular bandwidth of 2·78 Mc/s. The human eye is thus seen to be more sensitive to losses which involve phase than to those which are phaseless.

It should be mentioned that the straight-line correlation which is illustrated in Fig. 7 was found to exist for both sine-squared networks and *RC* networks with one exception; namely, when the standard comparison picture was of low definition (sharpness factor = 0·45). The judgements of the observers plotted in the usual way gave a curved line for this case. This is illustrated in Fig. 9, which includes the information given in Fig. 7, and shows that although a straight-line relationship holds for four of the sets of points, the set relating to a comparison picture of sharpness factor 0·45 is quite definitely curved. It would be pointless to try a change of parameter for the abscissa, as the excellent correlation which holds for all the other experimental results would then be vitiated. The straight-line relationship which exists for almost all the experimental results cannot exist over a very large range of values of equivalent bandwidth, as there will be a standard of picture in the high-definition class for which the comment 'much sharper' holds and, beyond that, no increase in score can take place however excellent the definition of the picture. Likewise in the opposite direction, a certain picture will be classed as 'much less sharp' and any further reduction in equivalent bandwidth cannot alter the grading. Thus, the curve must become asymptotic to its maximum and minimum scores (±45 in the case of fifteen observers). In spite of this, there does seem to be a considerable range of values for which a straight-line relationship holds.

The question of picture astigmatism is of interest. The degrading networks affect resolution in the line direction only, so that the fully-resolved 3-Mc/s picture (unity

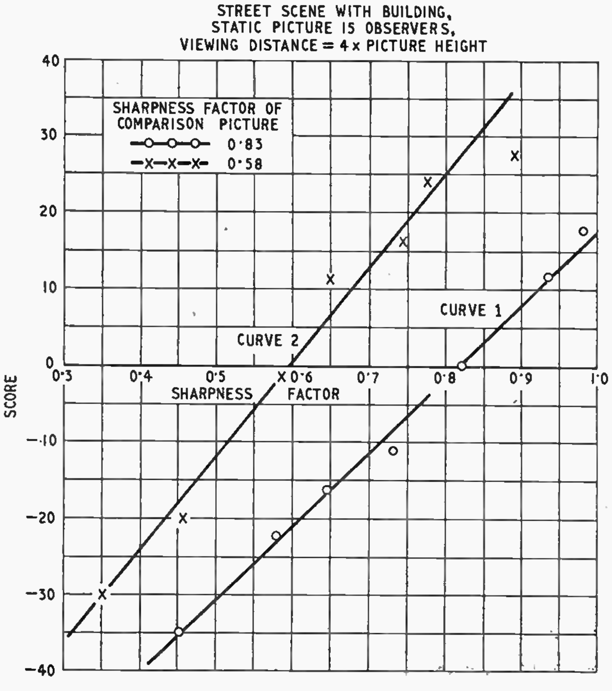

Fig. 7. *Subjective assessment using sine-squared networks*

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* Best in the statistical sense of minimum sum of squares of deviations.

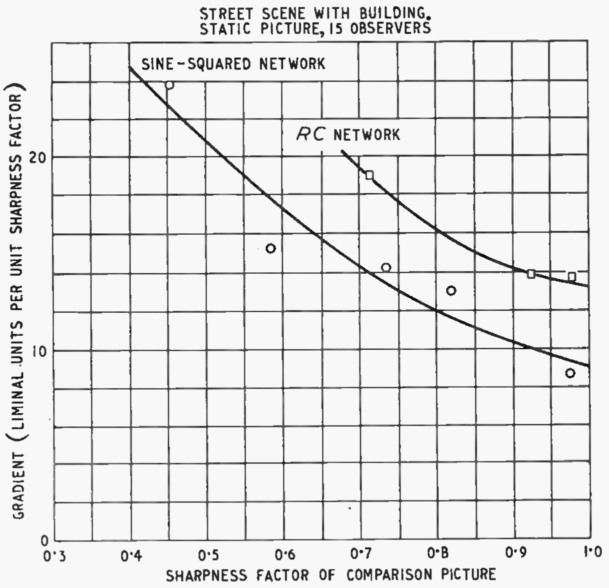


Fig. 8. *The effect of the sharpness factor of the comparison picture upon the subjective response to changes in sharpness factor*

sharpness factor) is the only one in the United Kingdom 405-line system which is non-astigmatic. As the equivalent bandwidth is reduced, the picture must of necessity become astigmatic. Baldwin[4] has shown that over fairly wide limits the eye is not sensitive to picture astigmatism, and as our limits are within those found by Baldwin, we shall make the assumption that our method of reducing the bandwidth does not seriously restrict the application of our results to cases where there is no astigmatism. This is certainly true in the vicinity of a fully-resolved 3-Mc/s picture and most of the interest centres around this value, since we are concerned with the production of good 3-Mc/s pictures.

Moving Pictures

The type of scene and the presence or absence of motion might considerably affect the subjective judgement of sharpness. The scene used for all the tests described so far was a static frame from a 35-mm motion-picture film and showed a considerable amount of detail. To extend the scope of the tests, a loop of 35-mm film was made up which had rapid movement in it; two beach scenes involving ball games were used for this purpose. The results from these tests were similar to those already described except that the gradients of the straight lines connecting score with sharpness factor were somewhat less. The correlation was still good (correlation coefficients greater than 0·9) and the case of comparison with a standard of sharpness 0·45 once more produced a curve and not a straight line. The gradients for a range of values of the sharpness of the standard comparison picture are shown in Fig. 10. The resistance-capacitance networks once more give higher gradients than the sine-squared networks. For a fully-resolved 3-Mc/s picture the gradient for a phaseless degradation is 6·7 liminal units per unit change of sharpness factor; this means that 50% of the population can perceive a change in sharpness when the equivalent bandwidth is reduced from 3 Mc/s to 2·55 Mc/s. For a resistance-capacitance loss the gradient is 9·7: one liminal unit corresponds to a change from a fully-resolved 3-Mc/s picture to one with an equivalent rectangular bandwidth of 2·69 Mc/s.

Movement in a television picture is not always rapid. Another loop of film was therefore made up depicting Richard Dimbleby making an announcement. The movement in this loop was very restricted. The gradients for this subject were found to be intermediate between the static case when the subject matter was detailed and the moving case where the movement was at times rapid. It is considered that almost all practical cases will lie somewhere between the two cases just quoted, the results of which are given in Figs. 8 and 10.

Discussion on the Use of a Multi-Criterion Series of Judgements

In describing the method it was stated without explanation or justification that seven gradings were permitted for the observers to classify their reactions to the sharpness of an unknown picture in terms of a given comparison picture. Fundamentally there are only three categories in which an observer may place a picture, namely 'sharper', 'the same as', 'less sharp'. Nevertheless, degrees of comparative sharpness must exist, although there is no guarantee that different observers will mean the same thing when a given description is used. Hopkinson[5] has been using the multi-criterion technique in subjective studies for some considerable time and has shown that observers can maintain consistency and constancy over a period of time. His work was concerned with the assessment of glare conditions but the

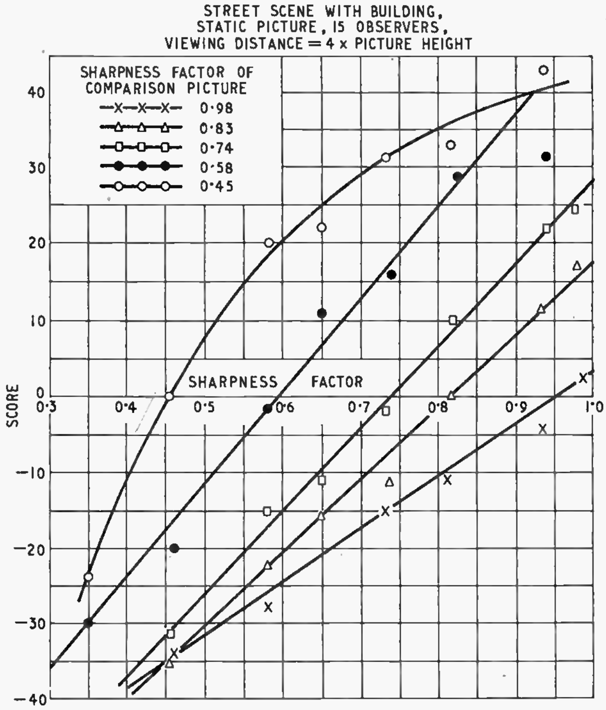


Fig. 9. *Subjective assessment using sine-squared networks*

principles of multi-criterion assessment are probably of wide validity. Clearly, if too many categories are used, the attempt to secure more detailed information will be spoilt by the spread of opinions amongst the observers as to the precise meaning to be attached to a given phrase. On the other hand, it may be possible to obtain more information than is possible by restricting the categories to 'better than' and 'worse than'. In the present instance the 'sharper' category has been subdivided into three—likewise the 'less sharp' category. This measure of subdivision would not appear to be excessive and the good correlation between score and sharpness factor in the results is some evidence in favour of this. If the question asked is difficult or meaningless, a subjective experiment will give very low correlation between the score and the objective variable.

As a more precise check of this method of assessing results the original data were classified into two categories only. The 'about the same' judgements were divided equally into the two categories, 'sharper' and 'less sharp'. The results were then plotted on arithmetic probability paper. The points so obtained in many cases (although not all) were capable of a straight-line interpretation, and from the slopes of these straight lines the value of a liminal unit could be deduced. Fig. 11 shows one such plot, and in Table 3 the gradients of the best straight lines are compared on the seven-criterion and two-criterion bases.

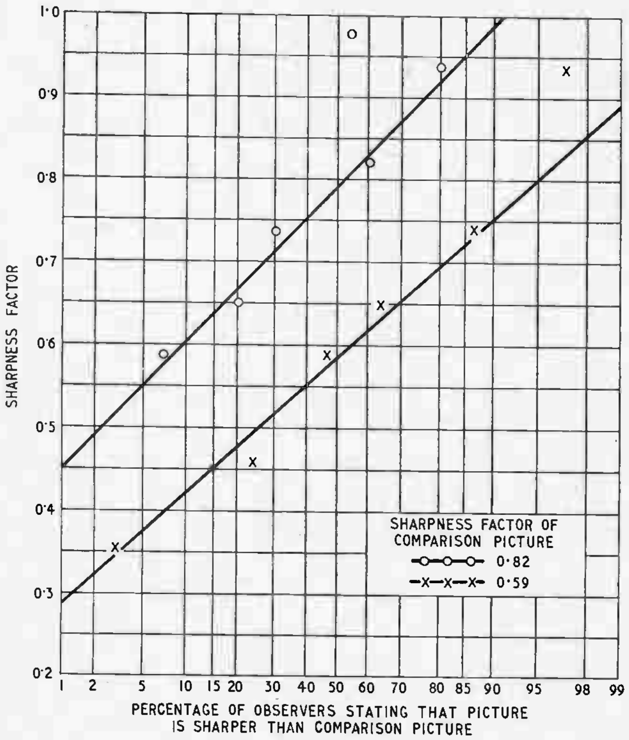


Fig. 11. *Assessment of sharpness using a two-criterion scale*

Apart from one or two exceptions, the two methods of assessment give very similar answers. One point of interest is that a probability scale cannot deal with complete certainty, and in one case (viz., sessions Nos. 4, 5 and 6, comparison with picture through sine-squared network No. 6) we have this situation if the results are assessed on a two-criterion basis. Thus, when the picture was shown with sine-squared network No. 7, all fifteen observers stated that the unknown was less sharp than the comparison picture. When the picture was shown through network No. 5 all the observers agreed that the unknown was sharper than the comparison picture. Since 0 and 100% lie at minus and plus infinity on a probability scale, the slope of the line cannot be deduced from this experimental evidence. The subjective change in sharpness between the pictures produced by networks Nos. 5 and 7 using No. 6 as a comparison is too great for this method of assessment. The seven-criterion

TABLE 3

Comparison of Gradients

Session No.	Standard Picture No.	7-Criterion Assessment Gradient (limens/unit change of sharpness factor)	2-Criterion Assessment Gradient (limens/unit change of sharpness factor)	Type of Degrading Network
1, 2, 3	2	13·1	13·8	$\sin^2$
	5	15·2	17·0	,,
	2	13·7	15·5	*RC*
	6	19·0	20·6	,,
4, 5, 6	0	8·6	11·8	$\sin^2$
	3	14·4	14·9	,,
	6	23·7	—	,,
	0	13·9	22·0	*RC*
7, 8, 9	2	9·5	9·8	$\sin^2$
	5	9·6	11·5	,,
	2	10·1	12·7	*RC*
	6	14·7	10·2	,,
10, 11, 12	0	6·7	5·5	$\sin^2$
	3	8·9	7·8	,,
	6	13·3	7·8	,,
	0	9·8	7·3	*RC*
13, 14, 15	2	10·7	9·8	$\sin^2$
	5	12·6	13·8	,,
	2	11·0	14·2	*RC*
	6	16·4	19·7	,,

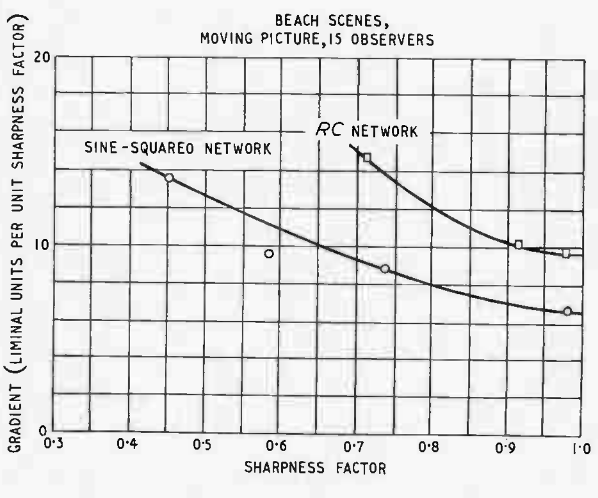


Fig. 10. *The effect of the sharpness factor of the comparison picture upon the subjective response to changes in sharpness factor*

method is in no such difficulty, since its multiple criteria permit various levels of certainty before the maximum or minimum score is reached.

The similarity of the gradients would appear to show that either method of assessment gives substantially the same answer, but there are conditions where the multi-criterion method gives more information.

Application to Lens Characteristics

Under the condition of a static picture containing much detail a phaseless degradation has a gradient of 9·2 limens per unit of sharpness factor. This means that a 3-Mc/s picture will be degraded to the extent of one limen when the equivalent bandwidth is 2·67 Mc/s. It is not possible to quote a modulation (sine-wave or square-wave) at 3 Mc/s which is exactly related to this equivalent bandwidth, since the shape of the frequency/response curve must be known. It is possible, however, on the basis of many measurements of lens characteristics to say that, on the average, an equivalent bandwidth of 2·67 Mc/s is approximately equivalent to 89% square-wave modulation at 3 Mc/s. If we have a lens giving 89% modulation over the whole of the field, its image will be one liminal unit down from the image produced by a lens which gives 100% modulation at 3 Mc/s. Certain figures have been suggested[6] for the performance of lenses for television. This appraisal of observer reaction to loss of definition shows that there is not a great deal to be gained by making lenses better than this figure, but the manufacturers have still some way to go before lenses in general have as good a performance as this. The axial modulation can (and often does) exceed 89%, but the off-axis performance produces an integrated figure which is appreciably lower than this.

The case of the *RC* circuit is somewhat analogous to that of a lens having comatic aberrations. The response of an *RC* circuit to unit-time impulse is markedly asymmetrical and the spatial response of a lens with coma is also asymmetrical. The degree of asymmetry from an *RC* network is, in general, greater than that of the residual comatic error in a lens and, hence, the gradients quoted for *RC* circuits will, in general, be greater than is likely to be found with any high-grade lens. For general purposes the gradient appropriate to the phaseless degradation can be applied to the whole field of a lens.

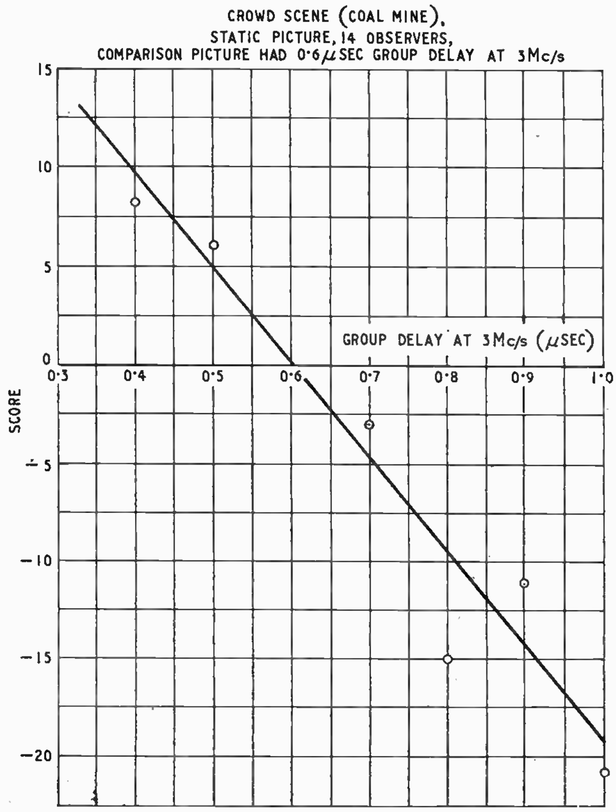
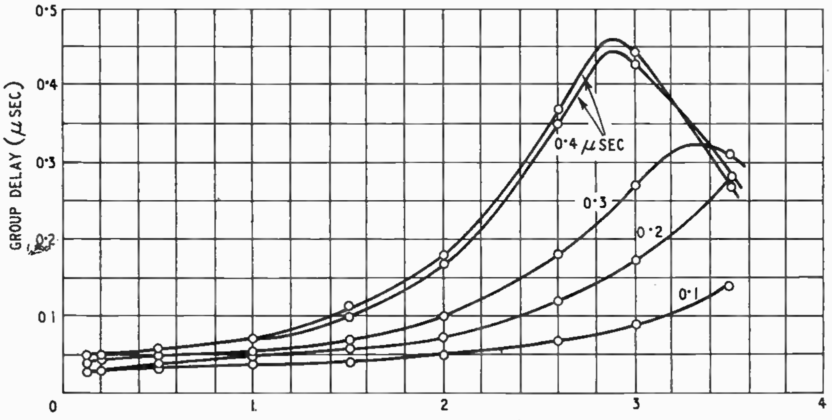


Fig. 13. *Subjective assessment of picture quality using phase-distorting networks*

Fig. 12. *Group delay characteristics of all-pass networks*



Conclusions

(1) The subjective assessment of sharpness is found to correlate well with the gradient of the response to unit step.

(2) The sensitivity to changes of equivalent bandwidth becomes less as the picture quality improves; i.e., for a liminal unit of improvement of picture quality a progressively greater increment of bandwidth is required as the picture sharpness increases.

(3) The eye is less sensitive to loss of amplitude which is phaseless (or phase-linear) than to a loss of amplitude associated with a non-linear phase characteristic.

(4) Under the optimum conditions of viewing and with a scene containing much detail one liminal unit corresponds to a phase-free change in equivalent bandwidth from 3 Mc/s to 2·67 Mc/s.

Acknowledgement

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

APPENDIX 1

Picture Sharpness Tests

Assess the unknown in terms of the standard comparison picture.

	Sharper				Less Sharp		
	Much Sharper	Sharper	Slightly Sharper	About the Same	Slightly Less Sharp	Less Sharp	Much Less Sharp
Set 1 1							
2							
3							
4							
5							
6							
7							
Set 2 1	–7 (as	Set 1)					
Set 3 1	–7 (as	Set 1)					
Set 4 1	–7 (as	Set 1)					

The complete form has four rows labelled Set 1, Set 2, Set 3 and Set 4. Only the first is reproduced here

APPENDIX 2

Subjective Effect of Phase Change without Loss of Amplitude

Although lenses produce some loss of amplitude (with or without phase distortion) it is an interesting case to consider an electrical network of the all-pass type, such as to give a constant-amplitude response but group delay increasing with frequency (up to about 3 Mc/s). A series of such all-pass networks was constructed with group delays of approximately 0·1, 0·2, 0·3 and 0·4-μsec delay at 3 Mc/s. The group delay characteristics are shown in Fig. 12. By adding several networks, group delays up to 1·4 μsec may be obtained. The effect on the picture is to produce more-or-less pronounced 'rings'—although the amplitude/frequency characteristic remains unaffected. Hence, one cannot question observers in this instance about the sharpness of the picture. The wording of the questionnaire was altered to 'better' and 'worse' and, as before, each of these two categories was subdivided into three. A very similar experimental technique was used and a typical result is shown in Fig. 13. The score is plotted against the group delay at 3 Mc/s. The 'goodness-of-fit' to a straight line was quite as good as in the previous experiment and correlation coefficients of values greater than 0·9 were nearly always obtained. The effect of the group delay of the standard comparison picture on the gradient is shown in Fig. 14. This graph shows a maximum sensitivity for a comparison picture with a group delay of approximately 0·4 μsec when the viewing distance is approximately four times picture height. Thus, from 0 to 0·4 μsec the effect on the gradient is not unlike that found for change of bandwidth; viz., the eye becomes less sensitive to a given change as the quality of the picture improves. For a group delay exceeding 0·4 μsec, the gradient becomes less and this may be due to the fact that we are considering pictures containing many 'rings' so that a slight change in the 'ringing' is not as noticeable as when the picture is fairly free from 'rings'. In this respect phase errors do not produce the same effect as loss of bandwidth, where the gradient continues to increase as the equivalent bandwidth is reduced (Figs. 8 and 10). The position of the maximum shown in Fig. 14 would appear to depend on viewing distance. A repeat of the experiment with the observers placed at eight times picture height gave gradients of very similar magnitude (in the range five to seven liminal units per μsec) but the peak value now occurred for a group delay exceeding 0·7 μsec if a peak still existed. The results of this experiment are shown in Fig. 15.

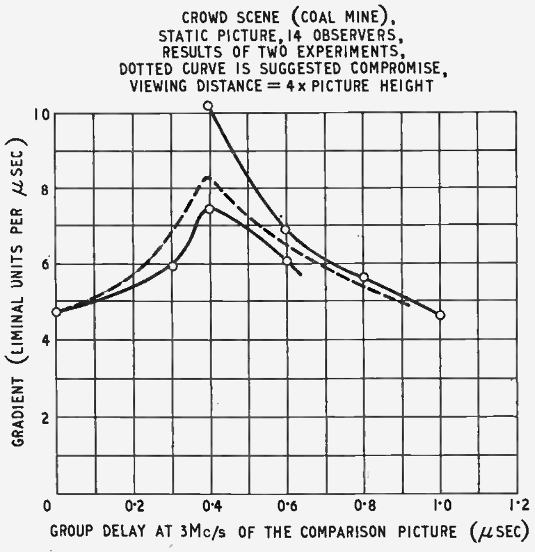


Fig. 14. *The effect of the group delay of the comparison picture on the subjective response to changes in group delay*

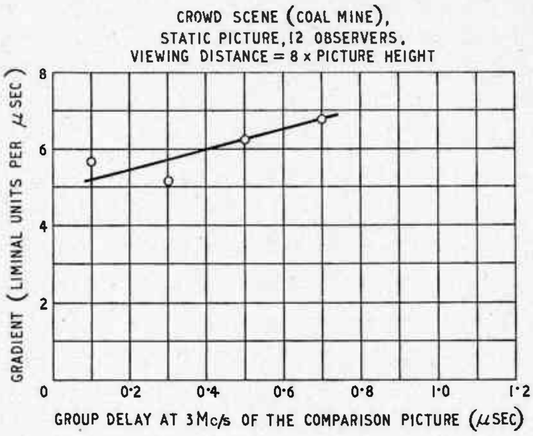


Fig. 15. *The effect of the group delay of the comparison picture on the subjective response to changes in group delay*

Restricting our interest to the region of good pictures, we can summarize the results by saying that a group delay of about 0·2 μsec at 3 Mc/s produces the subjective effect of one liminal unit; i.e., 50% of the population can just perceive a change and 50% regard the picture as unchanged.

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SOLID-STATE IMAGE CONVERTER

An infra-red picture converted to a visible one by means of a composite solid-state device utilizing semiconducting materials was shown by B.T-H. at the Physical Society Exhibition.

It consists of large-area photo-conductive and electroluminescent layers sandwiched between two transparent conducting electrodes. An alternating voltage is applied between the electrodes and the intensity of incident radiation on the photo-conducting layer controls the brightness of the corresponding electroluminescent area.

The Fringe of the Field

By Quantum

THE 21-CENTIMETRE LINE

In one sense this article is not getting very far away from its two predecessors, in which Lamb and Retherford's experiment on the splitting of the $2^2S_{\frac{1}{2}}$ level of the hydrogen atom was described. For, only a few megacycles away from the frequency there involved, less than a hand's-breadth shorter in wavelength, and only one step down in quantum number (to $1^2S_{\frac{1}{2}}$) for the level concerned, there is a single and singularly important emission line in the hydrogen spectrum—the 21-centimetre line. I must also add a postscript to the others for, while hunting up some 21-cm information, I consulted G. W. Series' monograph on 'The Spectrum of Atomic Hydrogen' (Oxford University Press, 1957). You can almost guess what would happen. Nothing about this particular line at all is there—but a very clear and authoritative account of the Lamb Shift and recent work on it. Indeed, if I had come across it sooner I could probably have been much more helpful on that subject; and can only, very much after the event now, commend it to you for further information.

In another sense, there is all the difference in the universe between the earlier articles and this one. They dealt with the problems of very short-range interactions; the 21-cm radiation from neutral atomic hydrogen has given the radio-astronomer a new and very powerful tool for exploring the contents of *remotest outer space*! This does not refer to the emission from the radio stars, which are discrete and concentrated sources of radio waves of various wavelengths, but to a single individual spectral line which is emitted from *everywhere* and comes to us from all over the place, even when the place is what we call the emptiest of 'empty space'.

It comes from the interstellar space of our own galaxy, from the galaxies beyond our own, and even probably from the intergalactic space between them. Although observations have only been made for seven years or so, the general results are becoming well known; and the use of the 21-cm line is mentioned, for example, in the astronomical section of Whitaker's *Almanack*. The main source of information that I have relied on is the excellent chapter on the subject in 'The Exploration of Space by Radio', by R. Hanbury Brown and Prof. A. C. B. Lovell (Chapman and Hall, 1957). This book presupposes a good knowledge of radio, and little of modern astronomy, so that the radio engineer should extract very full value from it.

Source of the 21-cm Line

The energy levels (or states) concerned with this line are those of the ground state $1^2S_{\frac{1}{2}}$ of the hydrogen atom. This is a hyperfine-structure doublet, in which $m = +\frac{1}{2}$ (the upper of the two) is separated from $m = -\frac{1}{2}$ by an energy difference of 0·048 cm^{-1}, which corresponds to a wavelength of about 21 cm, and a frequency of about 1,400 Mc/s; the exact figures are 21·105 cm and 1,420·4056 Mc/s.

The separation was measured as long ago as 1934 by Rabi, Kellogg and Zacharias. The agent responsible for the hyperfine structure is the magnetic moment of the proton, and the combination of this with $m = +\frac{1}{2}$ and $m = -\frac{1}{2}$ gives two possible magnetic moments to the H atom as a whole. So a beam of neutral atomic hydrogen (an 'atomic ray') passing through an inhomogeneous magnetic field is divided into two, each of which leaves its trace on a target. Working backwards from the trace separation to the difference of magnetic moment, the level separation was obtained. There are several other ways of determining it; but I probably said more than enough about hyperfine-structure last month, and need not repeat myself.

The upper of the two levels is metastable, with a transition probability $2{\cdot}85 \times 10^{-15}$ sec^{-1}, and an average lifetime for spontaneous emission of about 11 million years. The product of the transition probability for emission and the quantum energy $h\nu$ of the radiation should give the average power radiated by a single atom. This is

$2{\cdot}85 \times 10^{-15} \times 6{\cdot}5 \times 10^{-27} \times 1{,}420 \times 10^6$ erg/sec,

which works out to about 2×10^{-39} watt. Comparing it with the H_α line, with probability about 4×10^7 sec^{-1} and $h\nu = 3 \times 10^{-12}$ erg, giving an average power of $1{\cdot}2 \times 10^{-11}$ watt, this does not look very formidable. What the individual atom lacks in power, however, is compensated in exceedingly full measure by the huge numbers of atoms available in space.

The 21-cm line should be very sharp, but for one broadening agency—the Doppler effect. Both radiative damping and collision broadening should be entirely negligible. As for thermal-agitation collisions, these do indeed happen on an average once every hundred years or so. This does not put the above calculation, consistent with an 11-million-year lifetime, out of court; for the metastable state seems to be proof against this sort of rude interruption. The original excitation, indeed, possibly happens once for all in a big way with some such incident as a collision between two clouds of fairly 'concentrated' gas; but this part of the story seems rather obscure. Random motion of the individual sources gives a Doppler line *broadening* of about 75 kc/s; but the main value of the 21-cm line is the information that bodily Doppler *shifts of the whole line* give us about the velocity of whole gas clouds relative to the observer.

Distribution of Neutral H Atoms in Space

W. H. McCrea, in an article on cosmology written in 1953, gives the approximate extent of the observable universe as $1{\cdot}7 \times 10^{27}$ cm; its mass, $1{\cdot}1 \times 10^{55}$ gm; average density, 5×10^{-28} gm/cc; contents, some 10^8 galaxies; with the further information that, according to the Bondi-Gold-Hoyle theory, matter is being continuously created at 3×10^{-44} gm/cc/sec. 'Observable' means with the best optical resources available; and the figures really refer, of course, to estimates according to different 'models' of the universe.

Our own galaxy is a spiral Catherine-wheel-like structure which can be regarded as a disc about 10^{23} cm in diameter and 10^{21} cm thick, of total volume roughly 10^{67} cc. The matter distribution is densest in the 10^{11} stars, of which the sun is a typical example situated well off the centre of the galaxy, which is itself rotating about its own centre. Ionized hydrogen (H II) appears in the neighbourhood of stars, but the rest of the space is populated with a thin and unevenly distributed cloud (if that is the term) of neutral hydrogen atoms (H I), which are concentrated most densely in the spiral whirls but average about one atom per cc over the whole galaxy. This is almost unimaginably tenuous; comparing it with the conditions within a laboratory source of H_α, for example, the contents of an ordinary small discharge tube at this density would occupy about 10^{18} cc, or fill a cube of side 10 km. But, thin as it may be, there is enough of it present! The mass of a hydrogen atom is $1{\cdot}67 \times 10^{-21}$ gm; the total mass of 10^{67} hydrogen atoms is then $1{\cdot}67 \times 10^{46}$ gm. Now the mass of the sun, complete with the whole solar system, is only some 2×10^{33} gm; the interstellar hydrogen outweighs it by a factor of several million million. It even overtops the total mass of all the 10^{11} stars in the galaxy by more than a hundred times. The same sort of thing occurs in galaxies outside our own; the neutral hydrogen atoms in the Clouds of Magellan, for example, have about 10^8 times the mass of the sun; those in the Coma cluster some 8×10^{13} solar masses.

Power Radiated throughout the Galaxy

Take 10^{67} atoms, radiating with an average power of 2×10^{-39} watt apiece; the result is some 2×10^{28} watts. The total radiation from the sun, worked out from Stefan's law for a black body at 6,000°K, is of the order of 8×10^{25} watts. The interstellar hydrogen dispersed throughout the galaxy thus concentrates in this one spectral line about 3,000 times as much power as that radiated by the sun over its whole black-body spectrum.

When it comes to receiving the line, however, simple photometry shows there is not so much power to play with after all. An aerial which picks up a beam of solid angle Ω collects $\Omega/4\pi$ of the power reaching the earth from all directions. Now, the individual sources can be imagined to be situated on the surface of a sphere centred on the observer, at an average distance somewhere about half-way between us and the edge of the galaxy. This is not just a guess for, if ρ is the concentration of sources per unit volume, R the radius of a sphere which they fill uniformly, and x the radius of the spherical surface that we imagine them spread over instead, then using the inverse-square law and the usual symbols,

$$\frac{4\pi R^3\rho}{3x^2} = \int_0^{4\pi}\int_0^{R}\frac{\rho^2 d\Omega r dr}{r^2} = 4\pi R\rho, \text{ whence } x^2 = \tfrac{1}{3}R^2$$

So it appears that the equivalent distance x is $R/\sqrt{3}$. Of course, the galaxy is not spherical. But for this sort of argument it looks as if the right kind of distance is about 10^{22} cm. If so, then using the inverse-square law, the estimate for the resulting power collected is $2 \times 10^{28}/10^{44}$ (that is, 2×10^{-16} watts) to be multiplied by the $\Omega/4\pi$ factor. This does indeed turn out to be the sort of power level available.

Information from the 21-cm Line

Probably the most valuable single piece of information given by the 21-cm emission line is the distance of each of its more intense sources from the observer. These intense sources are clouds, or concentrations, of neutral hydrogen atoms, in which the population density is appreciably above the average—perhaps ten or more atoms per cc.

Just now I was saying, for the purpose of a rough power-order estimate, that the emitters were to be imagined as being uniformly distributed. Really, the more intense sources are concentrated in the spiral spokes of the Catherine wheel. But whichever way the 'spectroscope' is pointed, the 21-cm line is picked up. Relative motion between the source and the observer gives the line a considerable Doppler shift, and indeed the record as a rule shows several peaks, each corresponding to different concentrations along the line of sight that are moving with different velocities.

Now, the general way in which the galaxy rotates has been worked out, and the formula connecting the angular velocity ω of any part of it with the distance R from the centre of the galaxy is known. Starting with the values ω_0 and R_0 for the solar system, then ω for a source is found after calculating its relative velocity from the Doppler shift. Hence, as the relation between ω and R is known, R can be determined.

Secondly, the absorption of the 21-cm line from the emission of a radio star by an intervening cloud or spiral arm can be used to estimate the distance of the star; for it must at least be beyond the absorbing gas cloud.

Other problems in which it has given, or may give, valuable information are the 'red-shift', which seems to fit in with the optical red-shift for spectral lines, and to support the earlier interpretation of this as a straightforward recession effect; the possible measurement (by means of the Zeeman effect) of interstellar magnetic fields; and, with more sensitive receivers, the exploration of intergalactic space.

I do not propose to drop the topic for good, because Hanbury-Brown and Lovell give a number of references which it may be profitable to follow up. So, all in good time, I shall be coming back to spacemanship for you again.

CHANGE OF ADDRESS

The Professional Engineers Appointments Bureau has moved from 9 Victoria Street, London, S.W.1, to 39 Victoria Street, London, S.W.1. Telephone number: ABBey 1737.

CHANGE OF NAME

The name of the Radio Communication & Electronic Engineering Association (R.C.E.E.A.) has been changed to "The Electronic Engineering Association" (E.E.A.). The change took effect from 1st March.

Propagation Through a Dielectric Slab

EFFECT ON POLAR DIAGRAM OF SOURCE

By T. B. A. Senior*

The object of this note is to consider the apparent change in the nature of an electromagnetic source when viewed through a dielectric slab. Even when actually located at a point, the source appears to be diffused over a finite area, but for observation at large distances the effect is the same as that produced by a source of modified polar diagram situated at the original point. If the dielectric constant of the slab differs appreciably from that of the surrounding medium, the change may be important and could be used to provide an intentional modification to the polar diagram of an aerial system.

The Problem

An infinite slab of dielectric lies parallel to the xz-plane of some Cartesian co-ordinate system (x, y, z) and separates the surrounding medium (regarded for convenience as free space) into two regions (see Fig. 1) relative to which the refractive index of the dielectric is n. The surfaces of the slab are defined by the equations $y = a$ and $y = a + d$, where d is the thickness of the dielectric.

The m.k.s. system of units is employed and the time factor $e^{j\omega t}$ suppressed.

In the region below the slab there is a source of electromagnetic radiation and, for simplicity, we take this to be a line current coincident with the z-axis and propagating the cylindrical wave

$$H_0^{(2)}(kr) \approx \sqrt{\frac{2}{\pi kr}}\,\exp.\{-j(kr - \pi/4)\},$$

where r is the distance from the source. Our aim is to find the effect of the dielectric upon the field observed in the region above the slab, particular attention being paid to any apparent change in the nature of the source.

Analysis

In the form stated above the problem is two-dimensional and capable of expression in terms of the field component H_z alone. The field incident upon the slab is given by

$$H_z = H_0^{(2)}(kr) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

and may be analysed into plane waves using Sommerfeld's integral representation of the Hankel function

$$H_0^{(2)}(kr) = \frac{1}{\pi}\int_{S(\frac{1}{2}\pi)} \exp.\{-jkr\cos(\theta - \alpha)\}\,d\alpha \quad (2)$$

valid for $y>0$, where $x = r\cos\theta$, $y = r\sin\theta$ and $S(\frac{1}{2}\pi)$ is the path of 'steepest descent' through the angle $\frac{1}{2}\pi$. This result shows that Equ. (1), the solution for the incident field, can be obtained by appropriate integration, from that in which the plane wave

$$H_z = \exp.\{-jkr\cos(\theta - \alpha)\} \quad \ldots \quad \ldots \quad (3)$$

is incident upon the slab and suggests that a convenient approach to the problem would be to consider first the effect of the incident field (3). This we now proceed to do.

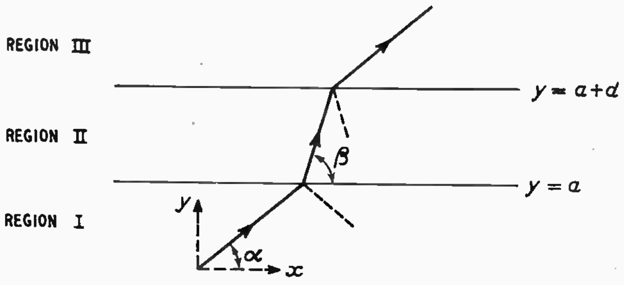

Fig. 1. *A typical ray path through the dielectric*

When the incident field is a plane wave it is clear that the total field in each of the three regions I–III (see Fig. 1) will consist of a superposition of plane waves. The following expressions for H_z are therefore assumed:

$$\begin{aligned} H_z^{\mathrm{I}} &= \exp.\{-jk(x\cos\alpha + y\sin\alpha)\} \\ &\quad + A\exp.\{-jk(x\cos\alpha - y\sin\alpha)\} \\ H_z^{\mathrm{II}} &= B\exp.\{-jnk(x\cos\beta + y\sin\beta)\} \\ &\quad + C\exp.\{-jnk(x\cos\beta - y\sin\beta)\} \\ H_z^{\mathrm{III}} &= D\exp.\{-jk(x\cos\alpha + y\sin\alpha)\}, \end{aligned}$$

where the angle β is given by Snell's law

$$\cos\beta = \frac{1}{n}\cos\alpha \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

The amplitude factors A, B, C, D are to be determined from the conditions that the tangential components of E and H are continuous across the surfaces $y = a$ and $y = a + d$. From the continuity of H_z at $y = a$ we have

$$\begin{aligned} &\exp.\{-jka\sin\alpha\} + A\exp.\{jka\sin\alpha\} \\ &\quad = B\exp.\{-jnka\sin\beta\} + C\exp.\{jnka\sin\beta\} \end{aligned}$$

and from the continuity of E_x,

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* *Royal Radar Establishment, Malvern.*

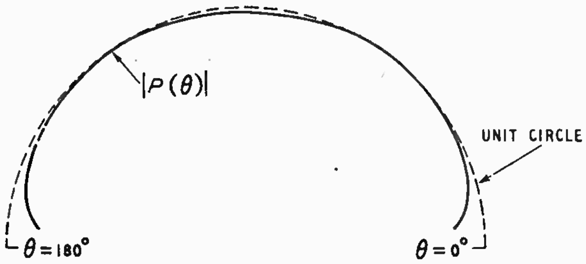

Fig. 2. *A typical polar diagram when the difference in refractive index is small:* $n = 1{\cdot}02$, $kd = 4$

$$\exp.\{-jka\sin\alpha\} - A\exp.\{jka\sin\alpha\} = \frac{1}{n}\frac{\sin\beta}{\sin\alpha}$$
$$[B\exp.\{-jnka\sin\beta\} - C\exp.\{jnka\sin\beta\}].$$

Similarly, the conditions at $y = a + d$ give

$$B\exp.\{-jnk(a+d)\sin\beta\} + C\exp.\{jnk(a+d)\sin\beta\}$$
$$= D\exp.\{-jk(a+d)\sin\alpha\}$$
$$B\exp.\{-jnk(a+d)\sin\beta\} - C\exp.\{jnk(a+d)\sin\beta\}$$
$$= n\frac{\sin\alpha}{\sin\beta}D\exp.\{-jk(a+d)\sin\alpha\}$$

and by eliminating A, B and C we find

$$D = \frac{\exp.(jkd\sin\alpha)}{\cos(nkd\sin\beta) + \frac{1}{2}j\left(\frac{\sin\beta}{n\sin\alpha} + \frac{n\sin\alpha}{\sin\beta}\right)\sin(nkd\sin\beta)} \quad .. \quad (5)$$

We now insert the value of β from Equ. (4) and write

$$\frac{1}{2}\frac{1 - 1/n^2\cos^2\alpha + n^2\sin^2\alpha}{\sin\alpha\sqrt{n^2-\cos^2\alpha}} = 1 + f(n, \alpha),$$

where we observe that $f(n, \alpha) \to 0$ as $n \to 1$. Then

$$D = \frac{\exp.\{jkd(\sin\alpha - \sqrt{n^2-\cos^2\alpha})\}}{1 + jf(n,\alpha)\exp.\{-jkd\sqrt{n^2-\cos^2\alpha}\}\sin(kd\sqrt{n^2-\cos^2\alpha})} \quad .. \quad .. \quad (6)$$

and satisfies the physical requirements that $D \to 1$ as either $n \to 1$ or $d \to 0$. The field appearing in region III is therefore

$$H_z = \frac{\exp.\{jkd(\sin\alpha - \sqrt{n^2-\cos^2\alpha}) - jk(x\cos\alpha + y\sin\alpha)\}}{1 + jf(n,\alpha)\exp.\{-jkd\sqrt{n^2-\cos^2\alpha}\}\sin(kd\sqrt{n^2-\cos^2\alpha})} \quad .. \quad .. \quad (7)$$

To obtain the transmitted field arising from the line source (1) it is only necessary to multiply (7) by π and integrate with respect to α over the steepest descent path $S(\frac{1}{2}\pi)$. Although the integration is over all real values of $\cos\alpha$ and hence includes complex angles of incidence, this introduces no difficulty, Equ. (7) being valid for all α. In this way we arrive at the following expression for the field produced in region III by the line source:

$$H_z = \frac{1}{\pi}\int_{S(\frac{1}{2}\pi)} P(\alpha)\exp.\{-jkr\cos(\theta - \alpha)\}d\alpha, \quad (8)$$

where $P(\alpha)$ is equal to the value of D given by Equ. (6) and represents the angular spectrum of the transmitted field.

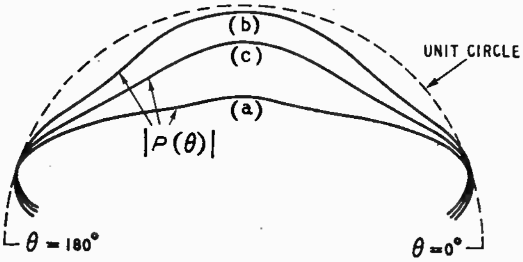

Fig. 3. *Polar diagrams for* $n = 3$; (*a*) $kd = 1/2$, (*b*) $kd = 2$, (*c*) $kd = 4$.

The Source

Explicit evaluation of the integral contained in Equ. (8) is a difficult, if not impossible, task, and we shall therefore content ourselves with an approximate treatment. A certain amount of information, however, can be obtained from a study of the general form of the integral.

The field component H_z has been represented as an angular spectrum of plane waves and if the dimensions of the receiver are supposed to be small, it is possible to regard the waves as rays radiating from some source. Each of these suffers a lateral displacement on passing through the dielectric slab, the magnitude of the displacement depending upon the angle at which the ray is incident. If the receiver is now moved an infinitesimal amount parallel to the xz-plane, the apparent position of the source may be determined as the point of intersection of the two rays. This position, however, is a function of the angle α and will change with further movement of the receiver; indeed, if the receiver is allowed to roam around in region III we shall be led to the conclusion that the source of radiation is diffused over a finite area.

When the receiver is at a large distance from the source (strictly speaking, we require that kr be large compared with unity), the effect of the diffused source is indistinguishable from that of a line source whose position is that of the true source but whose polar diagram is no longer uniform. To determine the polar diagram we return to Equ. (8) and apply the method of steepest descent. The saddle point is clearly $\alpha = \theta$, and since $P(\alpha)$ is a slowly varying function for all α not near to zero or π, we may remove this factor from beneath the integral sign at the predominant value $\alpha = \theta$. It follows that for large kr,

$$H_z \approx \frac{1}{\pi}P(\theta)\int_{S(\frac{1}{2}\pi)} \exp.\{-jkr\cos(\theta - \alpha)\}d\alpha$$
$$= P(\theta)H_0^{(2)}(kr), \quad .. \quad .. \quad .. \quad .. \quad (9)$$

and the radiation now appears to emanate from a single line source of polar diagram $|P(\theta)|$.

Let us consider $|P(\theta)|$. From Equ. (6) we have

$$\frac{1}{|P(\theta)|^2} = \cos^2 (kd\sqrt{n^2 - \cos^2 \theta}) + \{1 + f(n, \theta)\}^2 \sin^2 (kd\sqrt{n^2 - \cos^2 \theta})$$

$$= 1 + \frac{1}{4}\left\{\frac{\sqrt{n^2 - \cos^2 \theta}}{n^2 \sin \theta} - \frac{n^2 \sin \theta}{\sqrt{n^2 - \cos^2 \theta}}\right\}\sin^2 (kd\sqrt{n^2 - \cos^2 \theta}) \quad .. (10)$$

from which it is clear that $|P(\theta)|$ never exceeds unity. Nevertheless, $|P(\theta)|$ may equal unity, the condition for this being

$$n^2 - \cos^2 \theta - n^4 \sin^2 \theta = 0$$

or

$$\sin (kd\sqrt{n^2 - \cos^2 \theta}) = 0.$$

The former yields a real value of θ,

$$\theta = \sin^{-1} 1/\sqrt{1 + n^2}, \quad .. \quad .. \quad .. \quad .. (11)$$

depending only on the refractive index; the latter, however, provides a real θ only if there exists an integer m such that

$$n^2 - 1 \leqslant \left(\frac{m\pi}{kd}\right)^2 \leqslant n^2 \quad .. \quad .. \quad .. (12)$$

and the corresponding value of θ in this case is given by

$$\cos \theta = \pm\sqrt{n^2 - \left(\frac{m\pi}{kd}\right)^2} \quad .. \quad .. \quad .. (13)$$

The form of the polar diagram depends critically upon whether or not the condition (12) is satisfied. For reasonable values of n and dielectric thicknesses of the order of a wavelength (or less), the condition is, in general, unfulfilled, and all the examples which we shall give to illustrate the polar diagram fall into this category.

The removal of the factor $P(\alpha)$ from beneath the integral sign in Equ. (8) is valid only if $P(\alpha)$ is a slowly varying function of α near to the saddle point and this in turn requires that $\sin \theta$ be not small. We shall therefore confine our attention* to values of θ in the range $5° \leqslant \theta \leqslant 175°$.

If $(n - 1)$ is small compared with unity we write $n = 1 + \delta$ and, from Equ. (10), deduce that

$$|P(\theta)| = 1 - \tfrac{1}{2}\delta^2 \frac{\cos 2\theta}{\sin^4 \theta} \sin^2 (kd \sin \theta) + O(\delta^3).$$

This is relatively independent of the dielectric thickness and a typical polar diagram (actually drawn for $n = 1{\cdot}02$ and $kd = 4$) is shown in Fig. 2. The maximum change in $|P(\theta)|$ due to the presence of the dielectric is only about 2%.

When $(n - 1)$ is not small it is necessary to use the full expression for $|P(\theta)|$, and reference to Equ. (10) then shows that, in the main, $|P(\theta)|$ decreases with increasing n; indeed, for given refractive index, the minimum central value of $|P(\theta)|$ is

$$\frac{2n}{\sqrt{n^4 + 2n^2 - 1}},$$

a result which may be obtained by differentiating Equ. (10) with respect to kd.

For large values of n, $|P(\theta)|$ may differ appreciably from unity. To demonstrate this we have chosen the case $n = 3$, and in Fig. 3 polar plots are given for several thicknesses of dielectric. The shape of these curves is rather unusual and it is therefore of interest to trace their development with change of slab thickness. If $kd < \pi/3$ $(= 1{\cdot}047)$, no integer m can be found to satisfy the condition (12), and the resulting curves for values of kd much less than unity resemble thin ellipses touching the unit circle only at $\theta = \theta_0$ and $\theta = 180° - \theta_0$, where $\theta_0 = 18° \, 26'$. The thinnest of these corresponds to $kd = \pi/6$, the central value then being 0·606. As kd increases, so does the overall level of the curve; a hump begins to appear in the middle and finally, when $kd = \pi/3$, the hump touches the unit circle at $\theta = 90°$. With further increase of kd this tac point splits into two, one moving away towards $\theta = 0°$ and the other towards $\theta = 180°$, the limiting positions being attained when $kd = \pi/(2\sqrt{2})$ $(= 1{\cdot}111)$.

If we increase the value of kd still further, the curve reverts to the shape characteristic of small kd until, for $kd = 2\pi/3$ $(= 2{\cdot}094)$ the tac point reappears at $\theta = 90°$; the above process is then repeated, and so on for all integers m satisfying Equ. (12). The critical dependence on kd is now obvious.

The problem which has been discussed is, admittedly, rather an idealized one as regards practical application, but the results show that an appreciable change of polar diagram can be produced by even a thin sheet of dielectric. When the refractive index differs little from that of the surrounding medium, the modification to the polar diagram is small, and while it is possible for a freak atmospheric disturbance to produce a limited region of the atmosphere in which the refractive index differs from that of the neighbouring air, the change in n is not likely to be sufficient to produce an observable effect upon the polar diagram of a transmitter radiating under such conditions.

For large n, however, changes in $|P(\theta)|$ of the order of 60% (and more) are possible, though these are critically dependent upon the thickness of the dielectric, and this effect could be used to provide an intentional modification to the polar diagram of an aerial system by the placing of a sheet of suitable material (e.g., loaded dielectric) across the aperture. Since only a thin sheet is required, the losses associated with any practical dielectric are not likely to be troublesome.

PREMIUMS FOR TECHNICAL WRITING

The Radio Industry Council has announced the award of five premiums of 25 guineas each for articles published in the technical press during 1957. They are to:—

E. N. Shaw, A.M.I.E.E., for his article on "Heat Control in Electronic Equipment", which appeared in *Electronic Engineering* for January, February and March 1957.

E. J. Gargini, for his article on "An Alternative Colour TV System", which appeared in *Wireless World*, August 1957.

L. Atherton, B.Sc., A.M.I.E.E., for his article on "Cleaning by Ultrasonics", which appeared in *British Communications and Electronics*, March 1957.

E. J. P. Long, B.Sc.(Eng.), A.M.I.E.E., A.M.I.Mech.E., for his article on "Speed Control in Industry", which appeared in *British Communications and Electronics* for November and December 1957 and January 1958.

L. W. D. Sharp, M.A., M.I.E.E., for his article on "Components for Printed Circuits", which appeared in *British Communications and Electronics*, April 1957.

The awards will be presented at a luncheon to be held by the Radio Industry Council on 10th April.

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* In passing we note that, according to Equ. (10), $|P(\theta)| \to 0$ as $\theta \to 0°, 180°$, although such θ are excluded from our consideration for the additional reason that the receiver does not then lie in region III.

MATHEMATICAL TOOLS

By Computer

Matrices. 3—Relations between Network Matrices

We have already considered the general way in which matrix (and particularly A-matrix) elements can give information about the behaviour of networks, and the algebra of matrices by means of which the behaviour of a compound network can be expressed in terms of that of its components. In this article, we shall be concerned with various different matrices which can be associated with the behaviour of a given four-terminal network, and the way in which all of the matrices can be derived when any one of them is known. Further examples of analysis of circuits (including coupled circuits) by means of matrices will then be considered.

We shall, as hitherto, use the sign convention of Fig. 1. Previously, we have assumed that the input voltage and current were expressed in terms of the output voltage and current in the form

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

and that the rule of matrix multiplication immediately enables us to express (1) in the equivalent equation form

$$V_1 = a_{11} V_2 + a_{12} I_2 \, ; \; I_1 = a_{21} V_2 + a_{22} I_2 \quad \ldots \quad (2)$$

Initially, we assumed that our data were in the equation form (2), and (1) became a mere shorthand for (2) until we had learnt matrix algebra. Henceforward, however, we shall regard (1) as the natural way to express a relation, and only use the equivalent form (2) when this is absolutely necessary for purposes of explanation.

In (1) and (2) we have tacitly regarded V_2 and I_2 as 'known' quantities, in terms of which we wish to express the 'unknown' quantities V_1 and I_1. But we might equally well regard any two of V_1, V_2, I_1, and I_2 as 'known', and the other two as 'unknown'. We can choose six pairs of quantities to regard as 'known', and there are thus six different matrices which can be used for expressing the same relation as (1). We shall derive some of these matrices on the assumption that the elements a_{11}, a_{12}, a_{21} and a_{22} of the A-matrix A in (1) are known. First suppose that we regard V_1 and I_1, as known. We must then solve (2) as simultaneous equations for V_2 and I_2, and the solution in equation form is

$$V_2 = (a_{22} V_1 - a_{12} I_1)/|A| \, ; \; I_2 = (a_{11} I_1 - a_{21} V_1)/|A| \quad (3)$$

where $$|A| = a_{11} a_{22} - a_{12} a_{21} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

For a passive reciprocal network, the determinant $|A|$ in (4) is unity, but the determinants of matrices other than the A-matrix may have any value. We have therefore retained the denominator $|A|$ in (3), which will be needed when we carry out the corresponding process for a network for which the given matrix is not the A-matrix.

TABLE 1. Matrix Conversion

	A	A^{-1}	D^*	D^{-1}	Y	Z
A	$\begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix}$	$\frac{1}{\lvert A\rvert}\begin{pmatrix} a_{22} & -a_{12} \\ -a_{21} & a_{11} \end{pmatrix}$	$\frac{1}{a_{22}}\begin{pmatrix} \lvert A\rvert & a_{12} \\ -a_{21} & 1 \end{pmatrix}$	$\frac{1}{a_{11}}\begin{pmatrix} 1 & -a_{12} \\ a_{21} & \lvert A\rvert \end{pmatrix}$	$\frac{1}{a_{12}}\begin{pmatrix} a_{22} & -\lvert A\rvert \\ 1 & -a_{11} \end{pmatrix}$	$\frac{1}{a_{21}}\begin{pmatrix} a_{11} & -\lvert A\rvert \\ 1 & -a_{22} \end{pmatrix}$
D^*	$\frac{1}{d_{22}}\begin{pmatrix} \lvert D\rvert & d_{21} \\ -d_{21} & 1 \end{pmatrix}$	$\frac{1}{d_{11}}\begin{pmatrix} 1 & -d_{12} \\ d_{21} & \lvert D\rvert \end{pmatrix}$	$\begin{pmatrix} d_{11} & d_{12} \\ d_{21} & d_{22} \end{pmatrix}$	$\frac{1}{\lvert D\rvert}\begin{pmatrix} d_{22} & -d_{12} \\ -d_{21} & d_{11} \end{pmatrix}$	$\frac{1}{d_{12}}\begin{pmatrix} 1 & -d_{11} \\ d_{22} & \lvert D\rvert \end{pmatrix}$	$\frac{1}{d_{21}}\begin{pmatrix} -\lvert D\rvert & d_{11} \\ -d_{22} & 1 \end{pmatrix}$
Y	$\frac{1}{y_{21}}\begin{pmatrix} -y_{22} & 1 \\ -\lvert Y\rvert & y_{11} \end{pmatrix}$	$\frac{1}{y_{12}}\begin{pmatrix} -y_{11} & 1 \\ -\lvert Y\rvert & y_{22} \end{pmatrix}$	$\frac{1}{y_{11}}\begin{pmatrix} -y_{12} & 1 \\ \lvert Y\rvert & y_{21} \end{pmatrix}$	$\frac{1}{y_{22}}\begin{pmatrix} -y_{21} & 1 \\ \lvert Y\rvert & y_{12} \end{pmatrix}$	$\begin{pmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{pmatrix}$	$\frac{1}{\lvert Y\rvert}\begin{pmatrix} y_{22} & -y_{12} \\ -y_{21} & y_{11} \end{pmatrix}$
Z	$\frac{1}{z_{21}}\begin{pmatrix} z_{11} & -\lvert Z\rvert \\ 1 & -z_{22} \end{pmatrix}$	$\frac{1}{z_{12}}\begin{pmatrix} z_{22} & -\lvert Z\rvert \\ 1 & -z_{11} \end{pmatrix}$	$\frac{1}{z_{22}}\begin{pmatrix} z_{12} & \lvert Z\rvert \\ 1 & -z_{21} \end{pmatrix}$	$\frac{1}{z_{12}}\begin{pmatrix} z_{21} & \lvert Z\rvert \\ 1 & -z_{12} \end{pmatrix}$	$\frac{1}{\lvert Z\rvert}\begin{pmatrix} z_{22}, & -z_{12} \\ -z_{21} & z_{11} \end{pmatrix}$	$\begin{pmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{pmatrix}$
Explanation	$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = A\begin{pmatrix} V_2 \\ I_2 \end{pmatrix}$	$\begin{pmatrix} V_2 \\ I_2 \end{pmatrix} = A^{-1}\begin{pmatrix} V_1 \\ I_1 \end{pmatrix}$	$\begin{pmatrix} V_1 \\ I_2 \end{pmatrix} = D\begin{pmatrix} V_2 \\ I_1 \end{pmatrix}$	$\begin{pmatrix} V_2 \\ I_1 \end{pmatrix} = D^{-1}\begin{pmatrix} V_1 \\ I_2 \end{pmatrix}$	$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = Y\begin{pmatrix} V_1 \\ V_2 \end{pmatrix}$	$\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = Z\begin{pmatrix} I_1 \\ I_2 \end{pmatrix}$

* In transistor work, the h-matrix is commonly used. This matrix, in our notation, could be written

$$\begin{pmatrix} V_1 \\ -I_2 \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix} \begin{pmatrix} V_2 \\ I_1 \end{pmatrix}$$

since the current convention has the output current I_2 in the opposite direction from that assumed in these articles. Since we have used the D-matrix notation

$$\begin{pmatrix} V_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} d_{11} & d_{12} \\ d_{21} & d_{22} \end{pmatrix} \begin{pmatrix} V_2 \\ I_1 \end{pmatrix}$$

it follows that

$$h_{11} = d_{11} \, ; \; h_{12} = d_{12} \, ; \; h_{21} = -d_{21} \, ; \; h_{22} = -d_{22}$$

When (3) is expressed in matrix form, we find that

$$\begin{pmatrix} V_2 \\ I_2 \end{pmatrix} = A^{-1} \begin{pmatrix} V_1 \\ I_1 \end{pmatrix} \quad \ldots \quad (5)$$

where A^{-1} is the reciprocal matrix of A as defined in an earlier article.

Next suppose that we regard the currents I_1 and I_2 as known, and the voltages V_1 and V_2 as unknown. The matrix

$$\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = Z \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{pmatrix} \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} \quad \ldots \quad (6)$$

which expresses the relation (1) in this form is known as the impedance or Z-matrix; all its elements have the nature of impedances. Re-arranging the second of (2) and dividing by a_{21}, we have

$$V_2 = (1/a_{21})\, I_1 - (a_{22}/a_{21})\, I_2 \quad \ldots \quad (7)$$

and substituting for V_2 from (7) into the first of (2),

$$V_1 = (a_{11}/a_{21}) I_1 - (|A|/a_{21}) I_2 \quad \ldots \quad (8)$$

which in matrix form becomes (6) where

$$z_{11} = a_{11}/a_{21};\ z_{12} = -|A|/a_{21};\ z_{21} = 1/a_{21};$$
$$z_{22} = -a_{22}/a_{21} \quad \ldots \quad (9)$$

and we notice in passing that for a passive network obeying the law of reciprocity, so that $|A| = 1$, we have

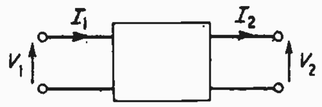
$$z_{12} = -z_{21} \quad \ldots \quad (10)$$

If we regard V_1 and V_2 as known and I_1 and I_2 as unknown, (6) can be expressed in the form

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = Y \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = \begin{pmatrix} y_{11} & y_{21} \\ y_{21} & y_{22} \end{pmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \quad \ldots \quad (11)$$

where Y, which is often called the admittance matrix, is the reciprocal Z^{-1} of the Z-matrix already found in (9). The values of y_{11}, y_{12}, y_{21} and y_{22} can be obtained in a very similar manner. The remaining pair of matrices is the D-matrix (when V_2 and I_1 are treated as known and V_1 and I_2 are treated as unknown) and its reciprocal D^{-1} (when V_1 and I_2 are treated as known and V_2 and I_1 as unknown). For these matrices also, the derivation of the elements in terms of a_{11}, a_{12}, a_{21} and a_{22} involves merely a repetition of the same kinds of algebraic manipulation. The situation is similar if the given matrix is other than the A-matrix; the formulae giving the elements of any matrix in terms of those of any other matrix can be worked out once for all, and the well-known results are given in Table 1. In this Table it is to be understood that $|A|$ means the determinant of the A-matrix, given by (4), and $|Y|$, $|Z|$, etc., are likewise the appropriate determinants of other matrices. Also, a factor outside any matrix must be multiplied into each element of that matrix; thus, the value of z_{11} when the A-matrix is given is, as already seen in (9), a_{11}/a_{21} and not a_{11}. If, for example, we know the A-matrix and require the Y-matrix, the required elements are in the A-row and the Y-column.

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



As we have seen, the A-matrix is very useful in enabling us to determine the behaviour of a compound network in terms of that of its components. It also has the advantage of having finite elements in the case of an ideal transformer, or a single series or shunt element; some of the other matrices contain infinite elements for these cases, as is seen from Table 1. However, the nature of the network may make it more convenient to determine initially the elements of any matrix, and then use Table 1 to deduce the different matrix most convenient for manipulation. If two networks are in series, as in Fig. 2, the Z-matrix of the combination is the sum of the Z-matrices of the separate networks. If two networks are in parallel, as in Fig. 3, the Y-matrix of the combination is the sum of the Y-matrices of the separate networks.

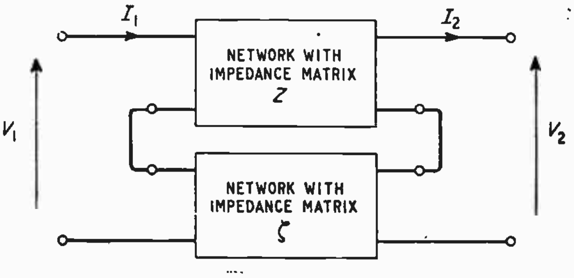


Fig. 2. *Netwo'ks in series*

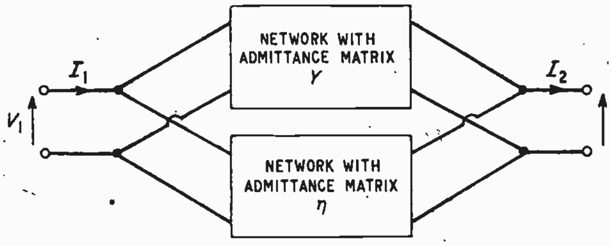


Fig. 3. *Networks in parallel*

Let us now apply the various techniques we have built up to the coupled circuit of Fig. 4.

If we consider the coils alone, we have

$$\left.\begin{aligned} V_1 &= pL_1I_1 + pMI_2 \\ V_2 &= pMI_1 + pL_2I_2 \end{aligned}\right\} \quad \ldots \quad (13)$$

so that, in matrix notation, it is the Z-matrix which arises naturally, and

$$\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = Z \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = p \begin{pmatrix} L_1 & M \\ M & L_2 \end{pmatrix} \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} \quad \ldots \quad (14)$$

As in Table 1, the factor p outside (14) multiplies every element of the matrix. As all the additional elements in Fig. 4 are parallel elements, we require (14) in Y-matrix form, not Z-matrix form. This means that we must express I_1 and I_2 in terms of V_1 and V_2; the result is given in the Z-row and the Y-column of Table 1, so that

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \frac{1}{p(L_1L_2 - M^2)} \begin{pmatrix} L_2 & -M \\ -M & L_1 \end{pmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \quad \ldots \quad (15)$$

To allow for the element C_1, we observe that the capacitance C_1 carries a current pC_1V_1 which increases I_1 only and has no effect on I_2. Likewise the conductance g_1 carries a current g_1V_1 which increases I_1 alone.

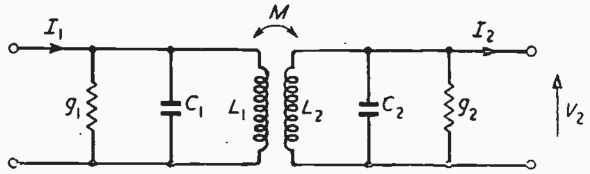


Fig. 4. *A coupled circuit*

Similarly, the capacitance C_2 carries a current pC_2V_2 and the conductance g_2 a current g_2V_2 which affect I_2 alone. The modification of (15) required when I_1, I_2, V_1 and V_2 are currents and voltages applicable to all the elements of Fig. 4 instead of merely the coils is therefore

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} pC_1+g_1+L_2/(pK) & -M/(pK) \\ -M/(pK) & pC_2+g_2+L_1/(pK) \end{pmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \quad .. \quad (16)$$

where $\quad K = L_1L_2-M^2 \quad .. \quad .. \quad .. \quad .. \quad (17)$

As there are no further elements in Fig. 4, we now set I_2 equal to zero. We can obtain V_1 and V_2 in terms of I_1, by re-converting (16) into Z-matrix form by means of the Y-row and the Z-column of Table 1. This gives

$$\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = \frac{1}{\Delta} \begin{pmatrix} pC_2+g_2+L_1/(pK) & M/(pK) \\ M/(pK) & pC_1+g_1+L_2/(pK) \end{pmatrix} \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} \quad .. \quad (18)$$

where

$$\Delta = \{pC_1 + g_1 + L_2/(pK)\}\{pC_2 + g_2 + L_1/(pK)\} - M^2/(p^2K^2)$$

$$= \left\{g_1 + \frac{L_2(1-\omega^2c_1K/L_2)}{j\omega K}\right\} \times \left\{g_2 + \frac{L_1(1-\omega^2c_2K/L_1)}{j\omega K}\right\} + \frac{M^2}{\omega^2K^2} \quad .. \quad (19)$$

when $p = j\omega$.

Hence, since $I_2 = 0$

$$V_1 = I_1\{pC_2+g_2 + L_1/(pK)\}/\Delta \; ; \; V_2 = I_1M/\{pK\Delta\} \quad .. \quad (20)$$

Now suppose that we are at liberty to choose C_1 and C_2, and that they are chosen so that

$$\omega_0{}^2 = L_2/C_1K = L_1/C_2K \quad .. \quad .. \quad .. \quad (21)$$

Then $\Delta = g_1g_2 + M^2/\omega_0{}^2K^2$

Suppose also that M is small compared to $(L_1L_2)^{1/2}$ and that g_1 and g_2 are small compared to $M/(\omega_0K)$. In this case, at frequencies sufficiently near $\omega_0/2\pi$, the circuit of Fig. 4 behaves like an impedance inverter, since the matrix (16) reduces for $p = j\omega_0$ approximately to

$$\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = \begin{pmatrix} 0 & jM/(\omega_0K) \\ jM/(\omega_0K) & 0 \end{pmatrix} \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \quad .. \quad (22)$$

and $|V_2/I_1|$ in (20) has approximately the large value $\omega_0L_1L_2/M$, indicating high sensitivity.

For our present purpose, however, what matters is not so much the particular conditions under which the circuit of Fig. 4 behaves like an impedance inverter and has high sensitivity as that, by means of matrices, an explicit formula (20) can be derived for the input and output voltages whatever the nature of the given input current I_1. As in many recent examples discussed in "Mathematical Tools", (20) is an explicit formula for the "p-world" counterparts of V_1 and V_2, and the complete time-expressions for these voltages can if necessary be deduced by operational calculus; under steady-state conditions at frequency $\omega/(2\pi)$, however, p can be replaced by $j\omega$.

MEETINGS

I.E.E.

9th April. "U.H.F. Test Transmissions". Informal evening.

10th April. "Radio Observations on Artificial Satellites", by J. A. Ratcliffe, O.B.E., M.A., F.R.S.

15th April. "A Train Performance Computer", by Prof. E. Bradshaw, M.B.E., M.Sc.Tech., Ph.D., M. Wagstaff, M.Sc.Tech., and F. Cooke. "The Simulation of Distributed-Parameter Systems with particular reference to Process Control Problems", by J. F. Meredith, B.Sc., and E. A. Freeman, B.Sc. "A Magnetic-Drum Store for Analogue Computing", by J. L. Douce, M.Sc., Ph.D., and J. C. West, B.Sc., Ph.D.

23rd April. "Survey of Performance Criteria and Design Considerations in High-Quality Monitoring Loudspeakers", by D. E. L. Shorter, B.Sc.(Eng.).

28th April. "Economic Usage of Broad-Band Transmission Systems", by R. J. Halsey, C.M.G., B.Sc.(Eng.). Informal evening.

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

Brit. I.R.E.

15th April. "Factors in the Design of Airborne Doppler Navigation Equipment", by E. G. Walker.

23rd April. "Measurement of the Frequency Response of a Nuclear Reactor", by R. J. Cox, B.Sc., D. Harrison and R. B. Stevens, B.Sc.

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

The Television Society

18th April. "Transistors in Television Receivers", by B. Overton, B.Sc., to be held at 7 o'clock at the Cinematograph Exhibitors' Association theatre, 164 Shaftesbury Avenue, London, W.C.2.

The Institute of Physics

18th April. "High Speed Radiography: 1. Tubes and Mechanisms", by R. F. Thumwood; 2. "Circuits and Evaluation", by R. Meakin, to be held at 6 o'clock at the Institute of Physics, 47 Belgrave Square, London, S.W.1.

Society of Instrument Technology

16th April. Annual General Meeting followed by "Electronically Operated Power-Plant Instrumentation and Control", by R. E. J. Putman, to be held at 6 o'clock at Manson House, Portland Place, London, W.1.

The Institution of Electronics

24th April. "Radio-Astronomy and the Sputniks", by Dr. F. G. Smith, at 7 o'clock in the Assembly Hall, University of London Institute of Education, Malet Street, London, W.C.1.

British Kinematograph Society

30th April. "An Instantaneous Electronic Colour Film Analyser", by the Hazeltine Research Corporation and Pathe Laboratories Inc., to be held at 7.15 at the Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.

Limited-Gain Operational Amplifiers

EQUIVALENT NETWORKS FOR SIMPLIFYING COMPUTATION

By A. W. Keen, M.I.R.E., A.M.I.E.E.*

SUMMARY. *The effect of finite gain in an operational amplifier can be allowed for by assuming that the amplifier gain is infinite and then adding fictitious circuit elements to the feedback network to reduce the gain to the value actually obtained. This procedure reduces the labour involved in calculating the operational error. The equivalent networks for some practical single-stage amplifiers are given.*

Frequent use is made, notably in analogue computers, of feedback amplifiers of a type whose external gain tends, with increase of internal gain, to a limiting value which is simply the ratio of two impedances comprising the feedback path. These circuits are called operational amplifiers, or operator networks, because they are well suited to the performance of prescribed mathematical operations on an input signal. Generally, an amplifier having three stages is employed to obtain sufficient internal gain to ensure that the voltage transfer actually achieved differs negligibly from the limiting value set by the feedback impedance ratio, but cases arise in practice where such a high gain may not be desired; in such cases, a single-stage amplifier would be preferred. For example, maximum computing accuracy may not be mandatory, or it may be difficult to secure an adequate margin against Nyquist instability, or economy of circuitry may be necessary, and so on. A familiar example of a single-stage operator network is the Blumlein-Miller integrator.

When the operational error due to inadequate internal gain is not negligible it is, of course, a straightforward matter to calculate its amount by conventional methods, but the labour involved may be considerable, even for a relatively simple circuit. In this article equivalent networks will be derived which illustrate clearly the effect on performance of the gain deficiency. It will be assumed that the internal amplifier has unlimited gain; the operational error will then be introduced by adding fictitious elements to the feedback network to reduce the external gain from the limiting value to that which is actually realized. Moreover, these elements will be inserted in such a manner that the input impedance is displayed as an independent group of elements, consisting of a principal component, equal to the value the input impedance would have with unlimited amplifier gain, together with additional secondary elements which modify its effective value to that obtained with limited gain.

Shunt-Shunt Feedback Form

In the most common form of operational amplifier, the shunt-shunt feedback configuration is employed. Its basic form is shown in Fig. 1. An inverting amplifier of very high voltage gain is fed from a source of signal voltage e_i through an impedance Z_1, and is bridged for feedback purposes by an impedance Z_2. It is convenient to add a third impedance Z_3 across the amplifier input so that, assuming the system to be terminated by a load impedance, all three terminal-pairs of the amplifier are shunted by external impedances, into which the amplifier impedances may be absorbed where they are not so large as to be negligible.

It is well known† that if the amplifier gain A were unlimited, the external gain of the system would exactly equal the ratio of Z_2 to Z_1, and the input impedance would be equal to Z_1 because the signal voltage at the amplifier input would be vanishingly small. In practice the limited value of A results in a gain less than Z_2/Z_1, which is an upper limiting ratio, and the input impedance exceeds Z_1, an 'error' signal voltage ϵ existing at the amplifier input. A comparison of the two cases suggests that the practical circuit may be represented by an equivalent network based on a hypothetical amplifier of unlimited gain, with the feedback network modified to obtain the values of voltage gain and input impedance achieved in practice. It would be necessary to augment Z_1 by a fictitious series impedance and to reduce the effective value of Z_2 by a fictitious shunt impedance.

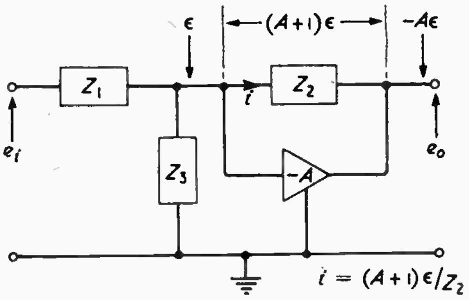


Fig. 1. *The basic form of the shunt-shunt feedback type of operational amplifier*

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* *Coventry Technical College.*

† See, for example: A. W. Keen, "Automatic Stabilization of Amplifier Gain", *J. Brit. Inst. Radio Engrs*, Vol. 10, No. 6, pp. 198–207 (June 1950).

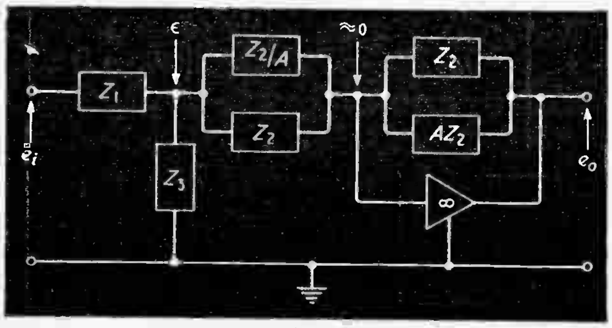


Fig. 2. *An equivalent network to Fig. 1, using an amplifier of unlimited gain together with a modified feedback network*

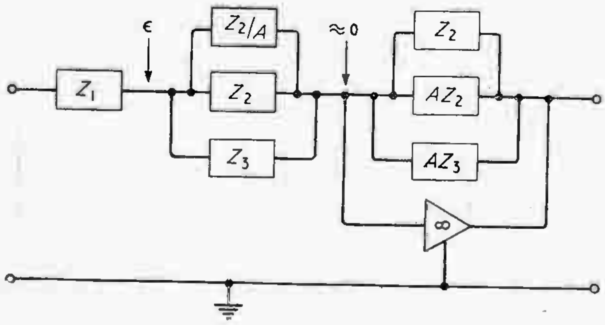
Both impedances would be functions of the actual amplifier gain and would vanish if the latter were indefinitely large.

In order to determine the values of these additional elements it is only necessary to consider that part of the system which lies to the right of Z_3 in Fig. 1. The signal current which flows into this section passes entirely through Z_2 because this is the only available path. The impedance presented by this path may be written by inspection in terms of the amplifier gain A as $Z_2/(A+1)$. It should be noted that the value of A is that which is obtained with Z_2 and any load impedance *in situ*. If, therefore, the amplifier is replaced by one of unlimited gain, as in Fig. 2, an impedance of value $Z_2/(A+1)$ must be inserted between the junction Z_1, Z_3 and the ideal-amplifier input to preserve the input impedance of the system.

In Fig. 2 this impedance has been resolved into two components Z_2 and Z_2/A in parallel; it will be noted that with increase of A the resultant decreases and ultimately becomes zero, as required. To restore the value of the gain to A it is necessary to reduce its effective value of Z_2 to A times $Z_2/(A+1)$. Since the latter consists of Z_2 and Z_2/A in parallel, the bridge impedance must consist of AZ_2 and Z_2 in parallel; an impedance of value AZ_2 must therefore be added in parallel with the existing impedance Z_2. Again, with increase of A this additional element becomes less effective and ultimately vanishes.

It will be seen from Fig. 2 that the signal voltages occurring across Z_3 and $Z_2/(A+1)$ are equal. This observation suggests connecting Z_3 directly across $Z_2/(A+1)$, with the advantage of reducing the group of elements which represent the input impedance to a two-terminal one. If this is done, it will be necessary to add a path of impedance AZ_3 over the amplifier to allow the current in Z_3 to flow over the amplifier bridge without increasing the forward gain. The alternative equivalent circuit so obtained is shown in Fig. 3.

Fig. 3. *An alternative form of the equivalent network given in Fig. 2*



It will be understood that these equivalent networks apply only to forward transmission, but they may be elaborated to account for output impedance (which in the form given is zero) if desired.

As a simple example of the foregoing, the basic Blumlein-Miller integrator circuit shown in Fig. 4 has the equivalent circuit shown in Fig. 5. In the latter the valve is assumed to have infinite values of μ and r_a; the finite μ and r_a values of the actual valve of Fig. 4 are included in the factors A and $1/A$ occurring in the values of the fictitious feedback elements of Fig. 5.

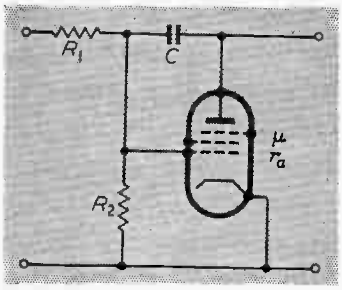


Fig. 4. *The basic Blumlein-Miller integrator circuit*

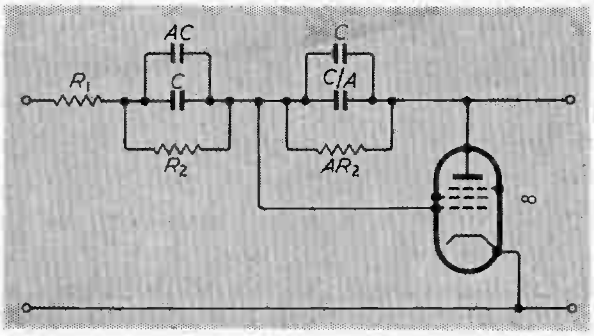


Fig. 5. *The equivalent network of the Blumlein-Miller integrator, using a valve of infinite μ and r_a*

Series-Series Feedback Form

In a less frequently used form of operational system, feedback is introduced by inserting an impedance (Z_1 in Fig. 6) 'underneath' the amplifier, where it is in series with both the input and the output circuits. It is convenient to add an impedance Z_3 in shunt with the amplifier input and to assume that the three external impedances Z_1, Z_2, Z_3 include the amplifier impedances where these are significant, as before.

If the amplifier gain were unlimited, the external gain would equal the ratio Z_2/Z_1 and the input impedance would be infinite. With limited internal gain, however, the external gain is less than the impedance ratio, and the input impedance, although possibly very

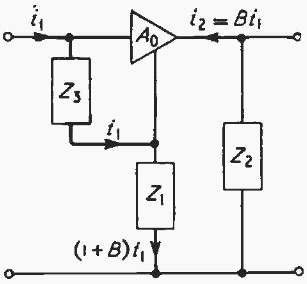


Fig. 6. *The series-series feedback form of operational amplifier*

great, is not infinite. In seeking an equivalent network for forward transmission of the type developed for the shunt-shunt system it is appropriate to make use of the duality between the two systems and to work in terms of current gain (B say) rather than voltage gain.

In terms of the current gain B $(= i_2/i_1)$* the input impedance of the system given in Fig. 5 may be written by inspection as $(1 + B)Z_1 + Z_3$, so that, if the amplifier gain is supposed to be infinite, an impedance of this value must be connected directly across the input terminal-pair of the equivalent network to give the correct value of input impedance since that of the hypothetical amplifier itself will be infinite. In Fig. 7, the input impedance has been resolved into a series group comprising Z_1, BZ_1 and Z_3. It will be noted that as B ranges from 0 to ∞ the sum of these elements ranges from $Z_1 + Z_3$ to ∞, as required. To reduce the gain to its proper value, keeping Z_2 fixed, Z_1 must be increased by the factor $(1 + B)/B$ because the current through it has been reduced by the inverse of this factor. In addition, an element of value Z_3/B must be added in series with the modified feedback impedance; otherwise the entire input signal voltage will appear across the modified feedback impedance. These modifications correspond exactly to those made in deriving Fig. 3 from Fig. 1.

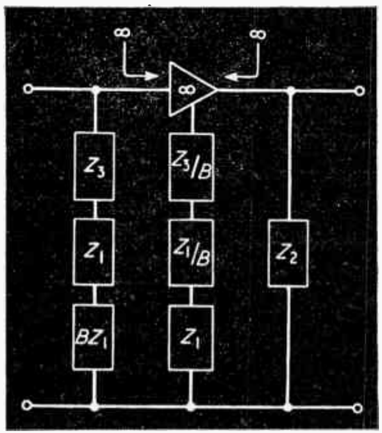


Fig. 7. *An equivalent network to Fig. 6, using an amplifier of unlimited gain and a modified feedback impedance*

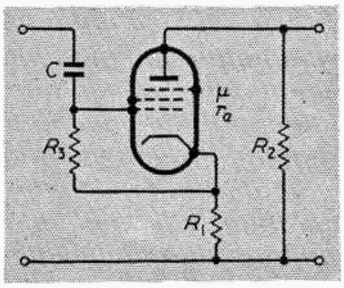


Fig. 8. *A simple series-series feedback amplifier*

The simplest example of a series-series single-stage amplifier is shown in Fig. 8; its equivalent forward transmission network, using a valve of infinite μ and r_a, is shown in Fig. 9.

Conclusion

It is believed that the given equivalent networks will prove useful to the circuit designer. They display clearly the dependence of input impedance and forward gain on the internal gain and allow the operational error incurred by a system of limited internal gain to be assessed more readily than conventional equivalent circuits or algebraic expressions.

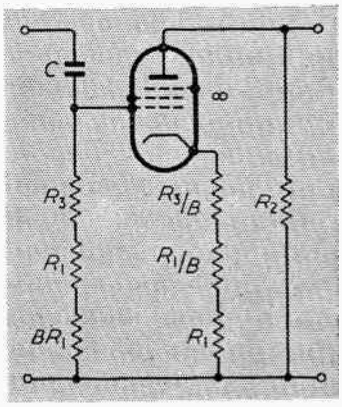


Fig. 9. *The equivalent network of Fig. 8 for forward transmission, using a valve of infinite μ and r_a*

APPENDIX

For an amplifier of open-circuit gain A_0 and output impedance Z_0 the values of A and B for the systems in Figs. 1–3 and 6–7 respectively are:

$$A = \frac{A_0 Z_2 (Z_L - Z_0)}{Z_0 Z_2 + Z_2 Z_L + Z_L Z_0},$$

where Z_L is the load impedance, and

$$B = \frac{A_0 Z_3 - Z_1}{Z_0 + Z_1 + Z_2},$$

where Z_2 includes the load.

For Figs. 4, 5:

$$A = \frac{\mu (Z_L - r_a)}{r_a + Z_L + j\omega r_a Z_L C} \approx g_m Z_L,$$

and for Figs. 8, 9:

$$B = \frac{\mu R_3 - R_1}{r_a + R_1 + R_2} \approx g_m R_3$$

TRANSLATION OF RUSSIAN PAPERS

An English translation of the *Bulletin of the Academy of Sciences of the U.S.S.R.*, Physical Series, Vol. 20, No. 12b, is available from Columbia Technical Translations, 5 Vermont Avenue, White Plains, New York, U.S.A., at the price of $20. It covers the Transactions of the 8th All-Union Conference on Semiconductors (Leningrad, 15th-21st November 1955), and includes approximately the first half of the papers listed in Abstract 2796 of 1957 (*Electronic & Radio Engineer*, September 1957).

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* Note that i_2 is taken inside Z_2 to allow the latter to include any load impedance

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Lawes Rabjohns Ltd.	940
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Livingston Laboratories Ltd.	412
Lloyds Bank Ltd.	400
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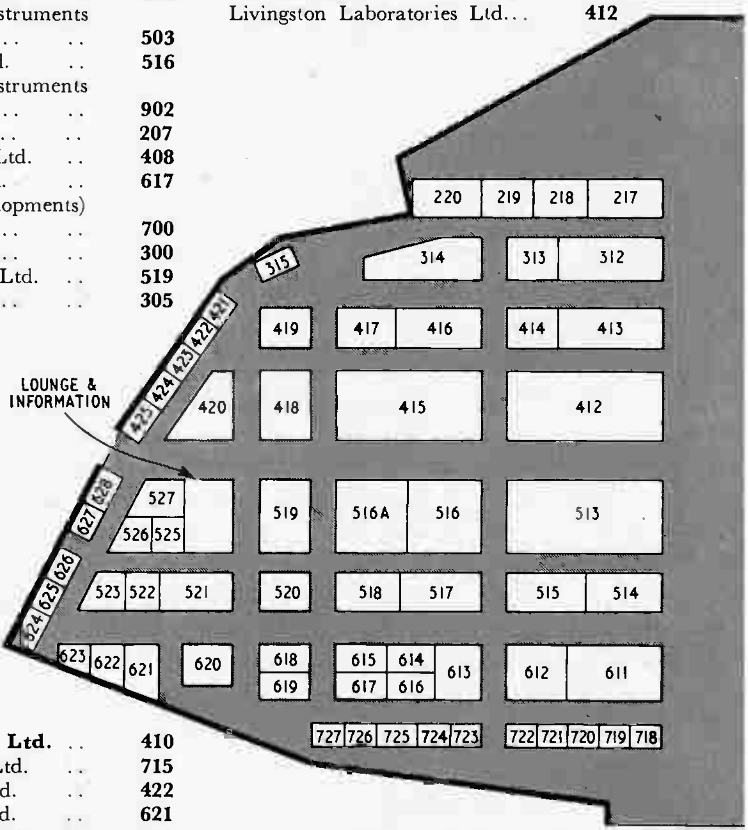

Continued on p. 147

Automation Exhibition

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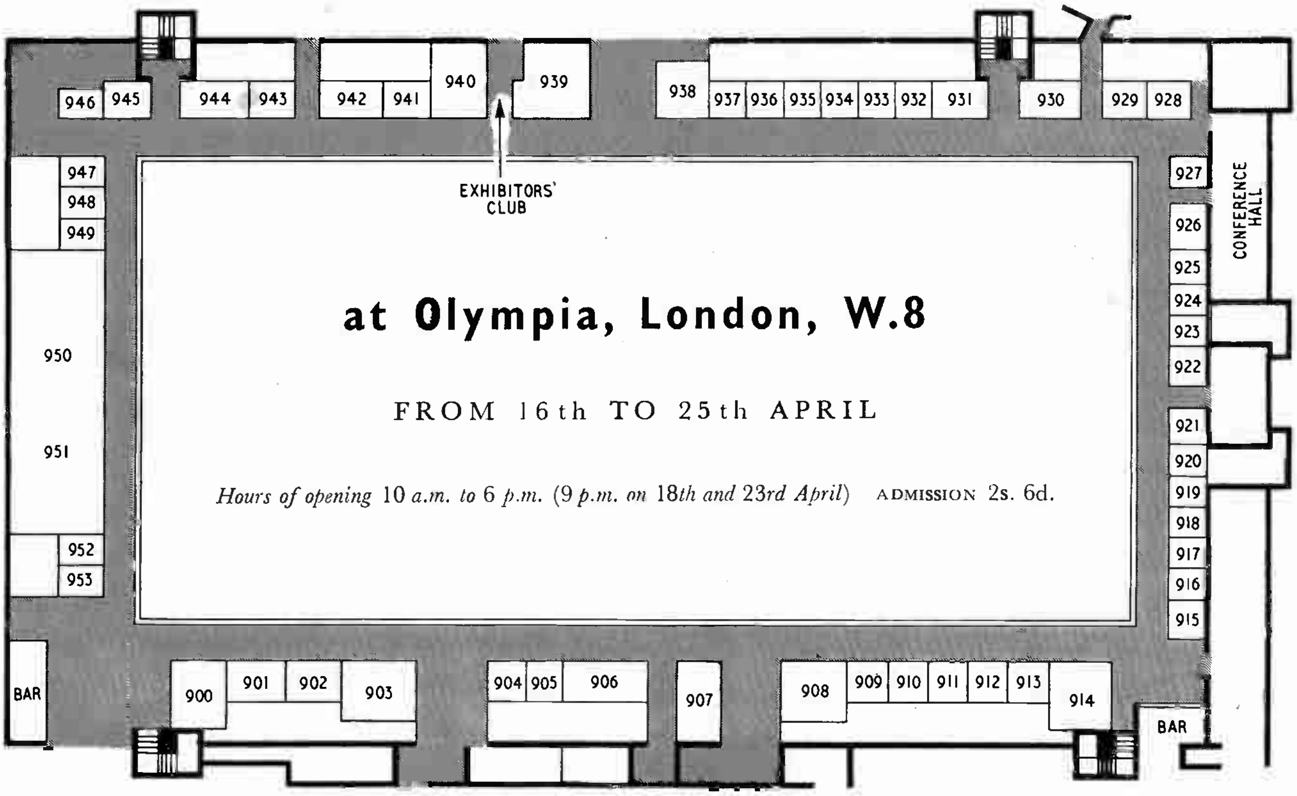

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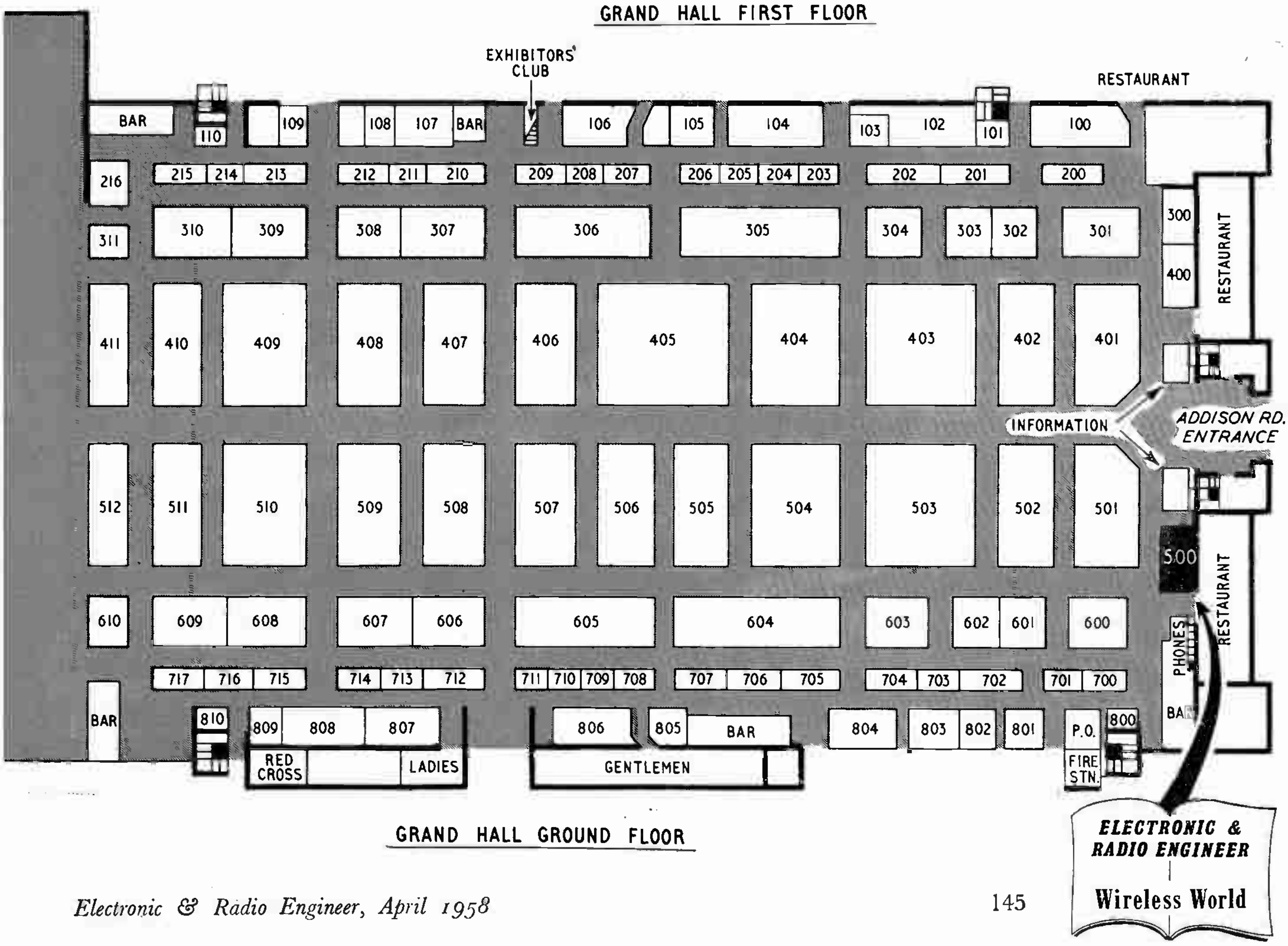
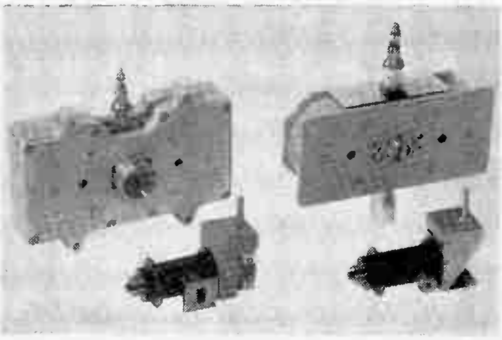
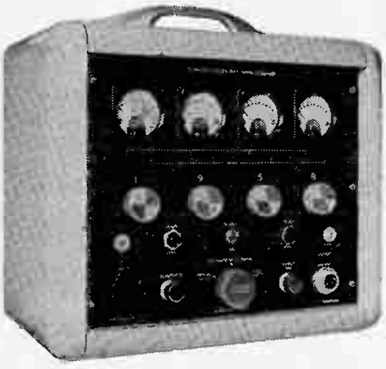
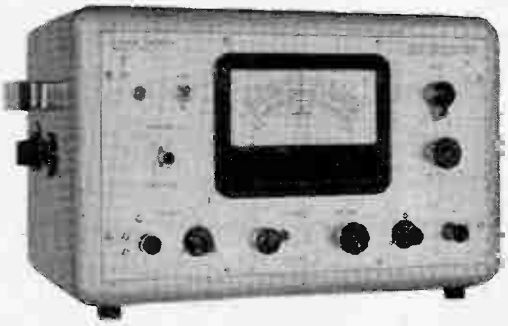




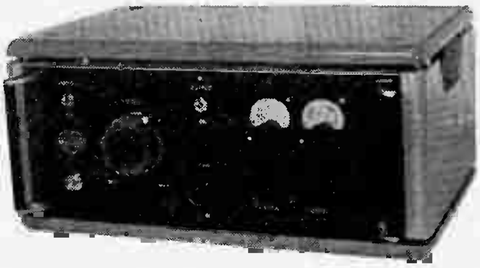
Top left: 'J' band magnetron VX5073; top right: 'Q' band magnetron VX5027; bottom left: reflex klystron R9520; bottom right: reflex klystron R5146. (E.M.I. Electronics Ltd.)



Transistorized batching counter for industrial counting. The instrument also provides precise delay times for laboratory use. Maximum counting speed 10 kc/s. (Venner Electronics Ltd.)

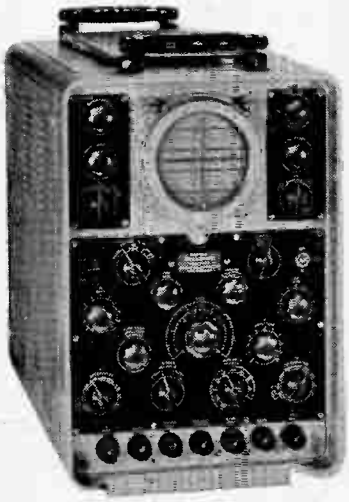


Component comparator for checking inductive, resistive or capacitive components against standards. Limits of ± 2%, ± 10% and + 25% to — 35% are indicated on a 5-in. meter. The bridge frequency is 1 kc/s. (British Physical Laboratories)

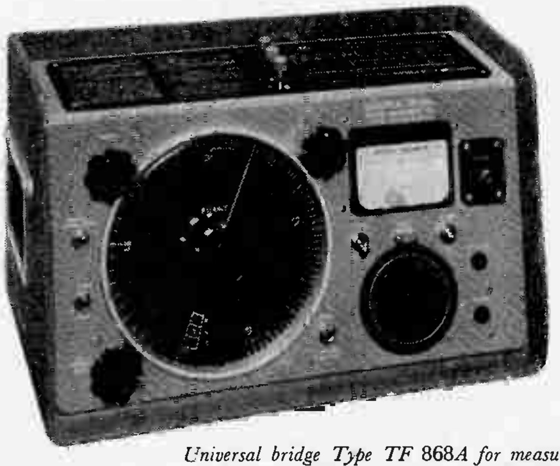


Stabilized power supply for transistors. A mains-operated power unit giving 0–30 V at currents up to 1A. The output resistance is less than 0·08 Ω, the ripple voltage less than 1 mV and the output voltage is stable within 200 mV at 20°–50° C. (Mullard Ltd.)

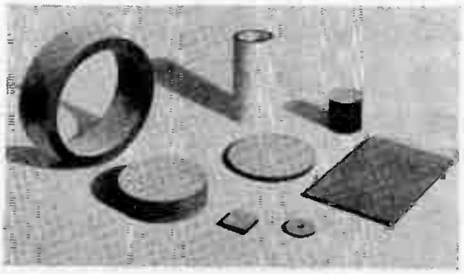
Low-frequency signal generator covering 15 c/s–200 kc/s. Output: 35 dBm to + 25 dBm in 600 ohms, frequency stability 0·1% at 1 kc/s, harmonics and hum—40 dB when the output is 1 mW. (Advance Components Ltd.)



Model 1059 double-beam wideband oscilloscope. The instrument has two similar amplifiers and a 4-in. p.d.a. tube. The sensitivity of the calibrated amplifiers is continuously-variable between 0·1 V/cm to 10 V/cm over a working bandwidth of 5 c/s to 10 Mc/s but additional pre-amplifiers covering the range 10 c/s–200 kc/s are provided. (Cossor Instruments Ltd.)



Universal bridge Type TF 868A for measuring inductance 1µH — 100H, capacitance 1 pF–100 µF, resistance 0·1 Ω — 10 MΩ. A bridge-voltage monitoring system is incorporated. (Marconi Instruments Ltd.)



Barium-titanite transducers for generating and detecting ultrasonic waves. (Technical Ceramics Ltd.)

High-speed rotary switch comprising 1-in. diameter 6-volt servomotor with 24-pole switch set. A life of 100 hours is obtained at 6,000 r.p.m., and the contact-position accuracy is 6 minutes per pole. (Vactric [Control Equipment] Ltd.)

Model 705 microammeter. This instrument has been developed to give enhanced presentation, greater readability and economy in panel space. (Ernest Turner Electrical Instruments Ltd.)

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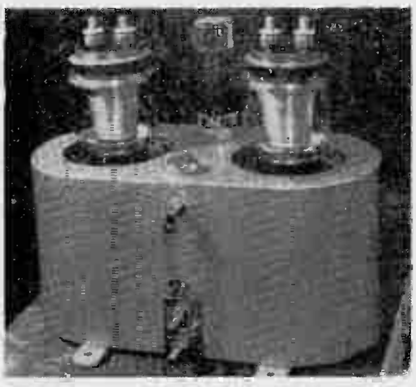
Stanton Instruments Ltd.	803
Sunvic Controls Ltd.	511
Taylor Controls Ltd.	602
Technical Ceramics Ltd.	949
Technical Sales	627
Technograph Electronic Products Ltd.	912
Teledictor Ltd.	724
Telegraph Construction & Maintenance Co. Ltd.	942
Telephone Manufacturing Co. Ltd.	302
Temple Press Ltd. (Nuclear Engineering)	913
Thermal Syndicate Ltd.	603
Tintometer Ltd.	947
Ronald Trist & Co. Ltd.	605
Ernest Turner Electrical Instruments Ltd.	910
20th Century Electronics Ltd.	904
Ultrasonoscope Co. (London) Ltd.	422
Unicam Instruments Ltd.	503
United Trade Press Ltd.	726
Vactric (Control Equipment) Ltd.	906
Venner Ltd.	953
Victoria Instruments Ltd.	312
Wade Couplings Ltd.	719
W. Watson & Sons Ltd.	503
Wayne Kerr Laboratories Ltd.	104
A. West & Partners Ltd.	108
West Instrument Ltd.	707
Western Manufacturing (Reading) Ltd.	423
Williams & James (Engineers) Ltd.	807
Wireless World	500
Worcester Royal Porcelain Co. Ltd.	705
W.S. Electronics (Production) Ltd.	206
G. H. Zeal Ltd.	801

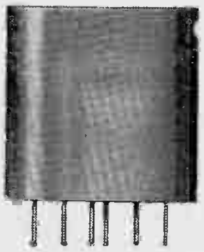
Overseas Exhibitors

(Firms represented by agents are shown in italics.)

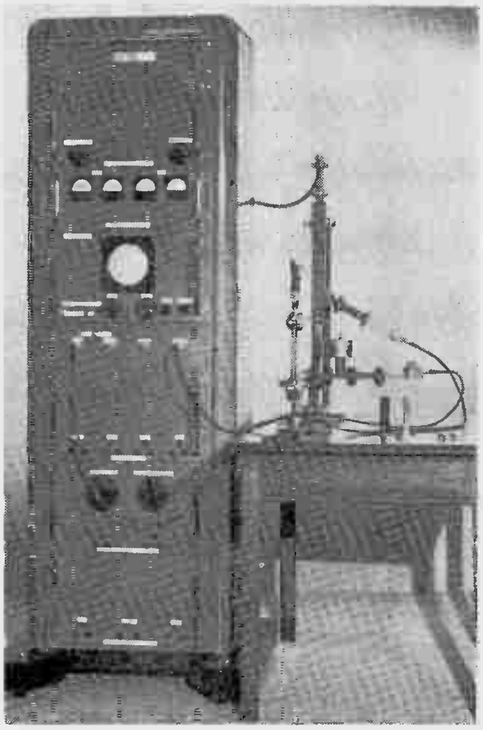
	Stand
Aveley Electric Ltd.	522
Rohde & Schwarz.	
Narda & Cascade.	
Gossen.	
Bayley, Clanahan & Co. Ltd.	420
Firma Degussa.	
B. & K. Laboratories Ltd.	628
Ampex Corporation.	
Bruel & Kjaer.	
DeMornay-Bonardi.	
Disa Electronik.	
Metrohm.	
Neutronics Research Co.	
Polarad Electronics.	
Polytechnic Research & Development.	
Sivers Lab.	
Varian Associates - Instrument Division.	
Wissenschaftlich Technische Werkstatten.	
Capitol Radio Engineering Inst.	626
Electro Mechanisms Ltd.	526
Kovo – Foreign Trade Corporation for the Import and Export of Precision Engineering Products	622
Dipl. Ing. P. Kretz	623
Livingston Laboratories Ltd.	415
Ballantine Laboratories.	
Bird Electronic Corporation.	
Max Funke KG.	
Gertsch Products Inc.	
Hewlett-Packard Co.	
Heinz-Gunther Neuwirth.	
Hoffman Electronics Corporation.	
Kay Electric Co.	
Kintel.	
Klemt Elektronische Messgeraete.	
Radiometer.	
S.I.D.E.R.	
Tektronix Inc.	
Telonic Industries Inc.	
Varian Associates.	
Perkin-Elmer A. G.	523
Bodenseewerk Perkin-Elmer & Co.	
Perkin-Elmer Corporation.	
Publi-Orga S. A.	950 & 951
Chauvin-Arnoux Et Cie.	
C.O.M.E.F.	
Constructions Electriques R.S.	
Ferisol.	
Intertechnique.	
Le Boeuf, Albert, et Fils.	
Legpa.	
L. Leiubray.	
Lemouzy.	
Nardeux.	
Photo Stress France.	
La Physiotechnie.	
S.A.M.A.	
S.E.F.R.A.M.	
S.N.R.S. France Automation.	
La Technique Electronique "Telec."	
Quality Bearings Ltd.	525
Barden Corporation.	
S.F.I.M. (Great Britain) Ltd.	526
Solartron Electronic Group Ltd.	527
Metrix Instruments Ltd.	
Wandel u. Goltermann.	
Schomandl K. G.	
Metrawatt A. G.	
Welmec Corporation Ltd.	521
Allgemeine Elektricitats Gesellschaft	
Dipl. Ing. Janovsky & Popp	
Gesellschaft Fur Elektrische Zugbeleuchtung M.B.H.	
W. Wykeham & Co. Ltd.	620
Siemens & Halske A. G.	



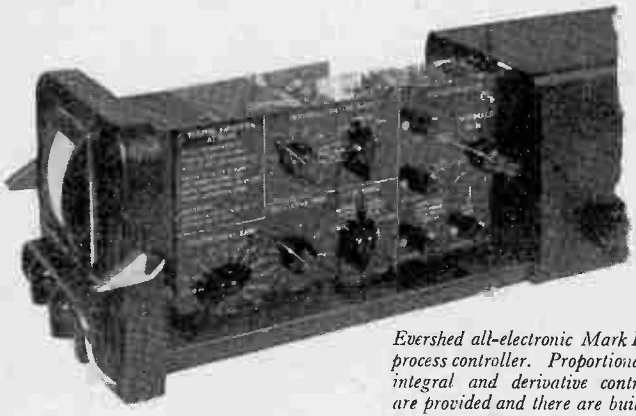
Two vapour-cooled triodes type BY 1102 in a boiler-condenser unit. Pure distilled water is used for cooling the anodes, heat being removed by a heat-exchanger supplied with a small quantity of ordinary water. (English Electric Valve Co. Ltd.)



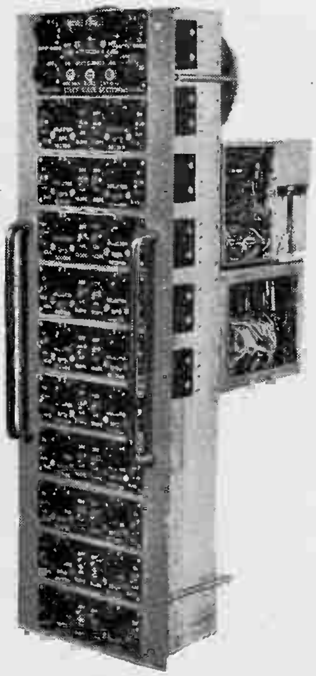
Series 335 sub-miniature relay. (Magnetic Devices Ltd.)



R. F. reflectometer. The use of this instrument substantially reduces the time needed to test wide-band waveguide and coaxial line components. The magnitude of the reflection coefficient of a circuit under test is displayed on an oscilloscope. (The General Electric Co. Ltd.)

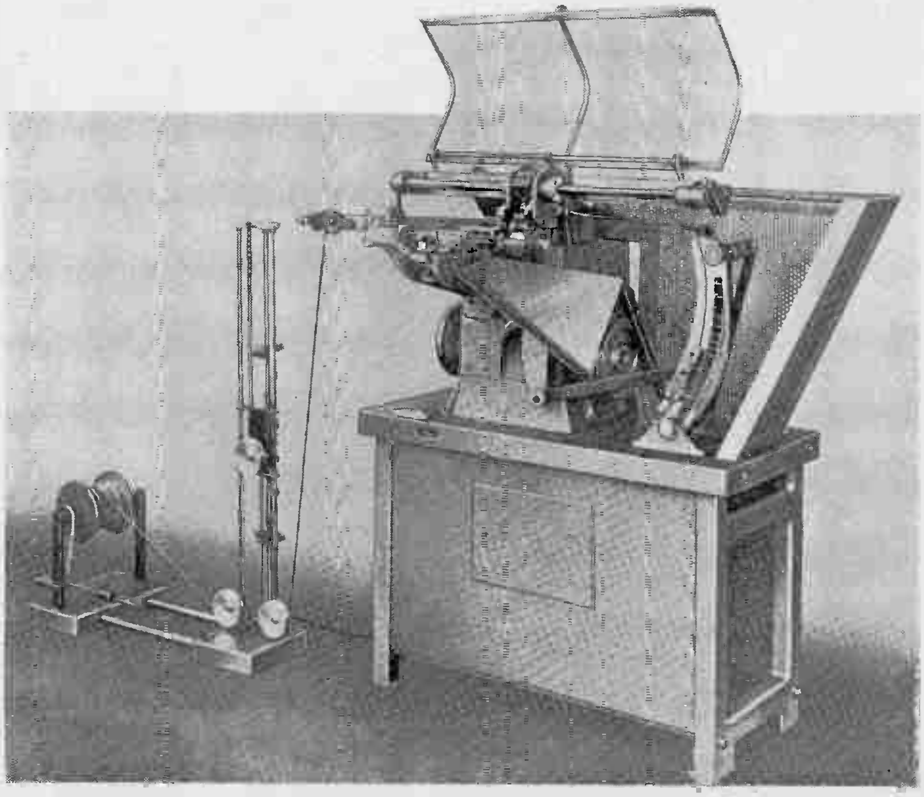


Evershed all-electronic Mark IV process controller. Proportional, integral and derivative control are provided and there are built-in test facilities. (Evershed & Vignoles Ltd.)



Continuous function unit as used in the Short analogue computer. May be set to generate an output signal which is any arbitrary continuous function of the input. (Short Bros. & Harland Ltd.)

Wire-cutting and stripping machine. This machine does the work of five trained girls. It only needs semi-skilled labour to set and start and requires no further attention till the supply reel is exhausted, when it automatically stops. It will cut and strip at both ends 3,000 pieces per hour up to 14½ in. long, or pieces up to 96 in. long at a reduced rate, and the machine also stops automatically on completion of a predetermined number of pieces. (Avo Ltd.)



Scattering of Electromagnetic Waves by Long Cylinders

By **Albert W. Adey, M.A.Sc., D.I.C., Ph.D.**

SUMMARY. *The field scattered by both a metal and a dielectric cylinder, when excited by an electromagnetic wave propagating in a direction normal to the cylinder axis, is discussed for both plane-wave and cylindrical-wave incidence. The radius of the cylinder is comparable with the free-space wavelength of the incident wave.*

The near-field behaviour is found to be more complex than that of the far field, very little similarity sometimes existing between the respective behaviours as a function of a parameter of the system.

It is shown that the dielectric cylinder is a resonant structure, its scattering properties varying widely with radius and material.

The angular distribution of the far scattered field is found to be predominantly forward, except for the smallest cylinders.

The internal field distribution for the dielectric cylinder is considered not to be amenable to the use of simple trial functions in approximate methods of treating the scattering.

In two previous papers[1, 2] the author discussed the total field external to the cylinder when a plane electromagnetic wave is diffracted by either a metal or a dielectric cylinder.

Since several recent papers[3, 4, 5, 6] have dealt with the scattered field itself, for both cylinders and spheres, it is felt to be of some interest to make some further observations about the scattered field of a cylinder and, in addition, the internal field in the case of the dielectric cylinder.

The effect on the scattering of a curvature of the incident wave-front is investigated. This factor is of practical interest when an array of cylinders is exposed to a plane wave, since the higher-order scattering arises from coupling among the cylinders, the scattered field from one cylinder exciting its neighbours.

Resonances in the scattering by a dielectric cylinder are studied. The scattering is found to be sensitive to the radius and material of the cylinder.

The angular distribution of the scattering is of interest and is found to be mainly forward, except for the smallest cylinders. The distribution determines the sensitivity of the scattering to the presence of other objects in the vicinity of the cylinder.

Since trial functions for the internal field of a dielectric cylinder are used in some approximate methods of determining the scattering, calculations of this field are made to determine whether or not the distribution appears amenable to the use of simple trial functions.

Only the case of incidence normal to the cylinder axis is considered here and, owing to the lack of suitable tables, only lossless cylinders are treated numerically.

Plane-Wave Scattering

Assume a plane wave of unit amplitude propagating in the direction $\phi = 0$ (of a cylindrical co-ordinate system r, ϕ, z) in a medium of permittivity, permeability and wave number ϵ, μ and k respectively. This plane wave is incident on an infinite cylinder whose axis coincides with the z axis (Fig. 1) and is parallel to the field electric vector. The cylinder is of circular cross section and is of radius a. It is either (a) perfectly conducting or (b) of a dielectric characterized by the parameters ϵ_1, μ_1, and k_1, defined as above.

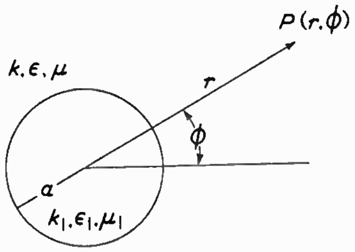


Fig. 1. *Section through the cylindrical co-ordinate system*

If the scattered field is taken to be of the form :

$$E_{sc}(r, \phi) = \sum_0^\infty e_n B_n H^{(1)}{}_n(kr) \cos n\phi \quad \text{..} \quad (1)$$

and the internal field of the dielectric cylinder as

$$E_{INT}(r, \phi) = \sum_0^\infty e_n A_n J_n(k_1 r) \cos n\phi \quad \text{..} \quad (2)$$

then, as has been shown previously[1, 2], the coefficients A_n and B_n are given by:

Conducting cylinder

$$B_n = -(i)^n J_n(ka)/H^{(1)}{}_n(ka) \quad \text{..} \quad \text{..} \quad (3)$$

Dielectric cylinder

$$B_n = -(i)^n \frac{\begin{vmatrix} J_n(k_1 a) \,,\, J_n(ka) \\ \frac{k_1}{\mu_1} J'_n(k_1 a) \,,\, \frac{k}{\mu} J'_n(ka) \end{vmatrix}}{\Delta} \quad \text{..} \quad (4)$$

$$A_n = \frac{k}{\mu} (i)^n \frac{\begin{vmatrix} J_n(k_1 a) , & H_n^{(1)}(ka) \\ J'_n(ka) , & H_n^{(1)\prime}(ka) \end{vmatrix}}{\Delta} \quad \ldots \quad (5)$$

where

$$\Delta = \begin{vmatrix} J_n(k_1 a) , & H_n^{(1)}(ka) \\ \frac{k_1}{\mu_1} J'_n(k_1 a) , & \frac{k}{\mu} H_n^{(1)\prime}(ka) \end{vmatrix} \quad \ldots \quad (6)$$

The B_n are seen to be of the form

$$B_n = -(i)^n \frac{a_n}{a_n + i\, b_n} \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

If $\mu = \mu_1$, which will be the case for all the scattered fields treated numerically, the above expressions for the dielectric cylinder simplify finally to

$$a_n = J_n(k_1 a)\, J_{n+1}(ka) - k_r J_{n+1}(k_1 a)\, J_n(ka) \quad (8)$$

$$b_n = J_n(k_1 a)\, Y_{n+1}(ka) - k_r J_{n+1}(k_1 a)\, Y_n(ka) \quad (9)$$

$$A_n = (i)^{n-1} \left(\frac{2}{\pi\, ka} \right) \Big/ (a_n + i\, b_n) \quad \ldots \quad \ldots \quad (10)$$

where

$$k_r = \frac{k_1}{k} = \sqrt{\frac{\epsilon_1 \mu_1}{\epsilon \mu}} \left(= \sqrt{\frac{\epsilon_1}{\epsilon}} \text{ when } \mu = \mu_1 \right)$$

The behaviour of the near, scattered electric field of a conducting cylinder, for $\phi = 0$, $\pi/2$ and π, is illustrated by the curves of Fig. 2.

The back-scattered field amplitude does not exceed the incident amplitude and, for a fixed value of kr, increases with the cylinder radius. In the shadow, however, the forward-scattered field sometimes exceeds the incident amplitude and oscillates about that value. Thus, contrary to what is often assumed, the scattered field in the cylinder shadow does not always have the property of a divergent wave in the near region, and the departure from this property extends further from the cylinder as the radius increases.

Curves of the side-scattered field show that, as ϕ increases to $\pi/2$, the scattered field pattern has already assumed the relatively simple form seen for the illuminated side.

As the cylinder radius decreases, the angular fluctuation of the scattered field gradually disappears, corresponding to the excitation of only the first scattering mode.

As in the case of the metal cylinder the scattered field of the dielectric cylinder appears to be more complex in the near region. Fig. 3 illustrates the field for a cylinder of $k_r = 1{\cdot}60$. However, as in the former case, the near-scattered field complexities occurred mainly on the shadow side, here they appear on the illuminated side. The field develops a deep depression which moves out and becomes broader with increasing ka. At about five wavelengths from the cylinder the

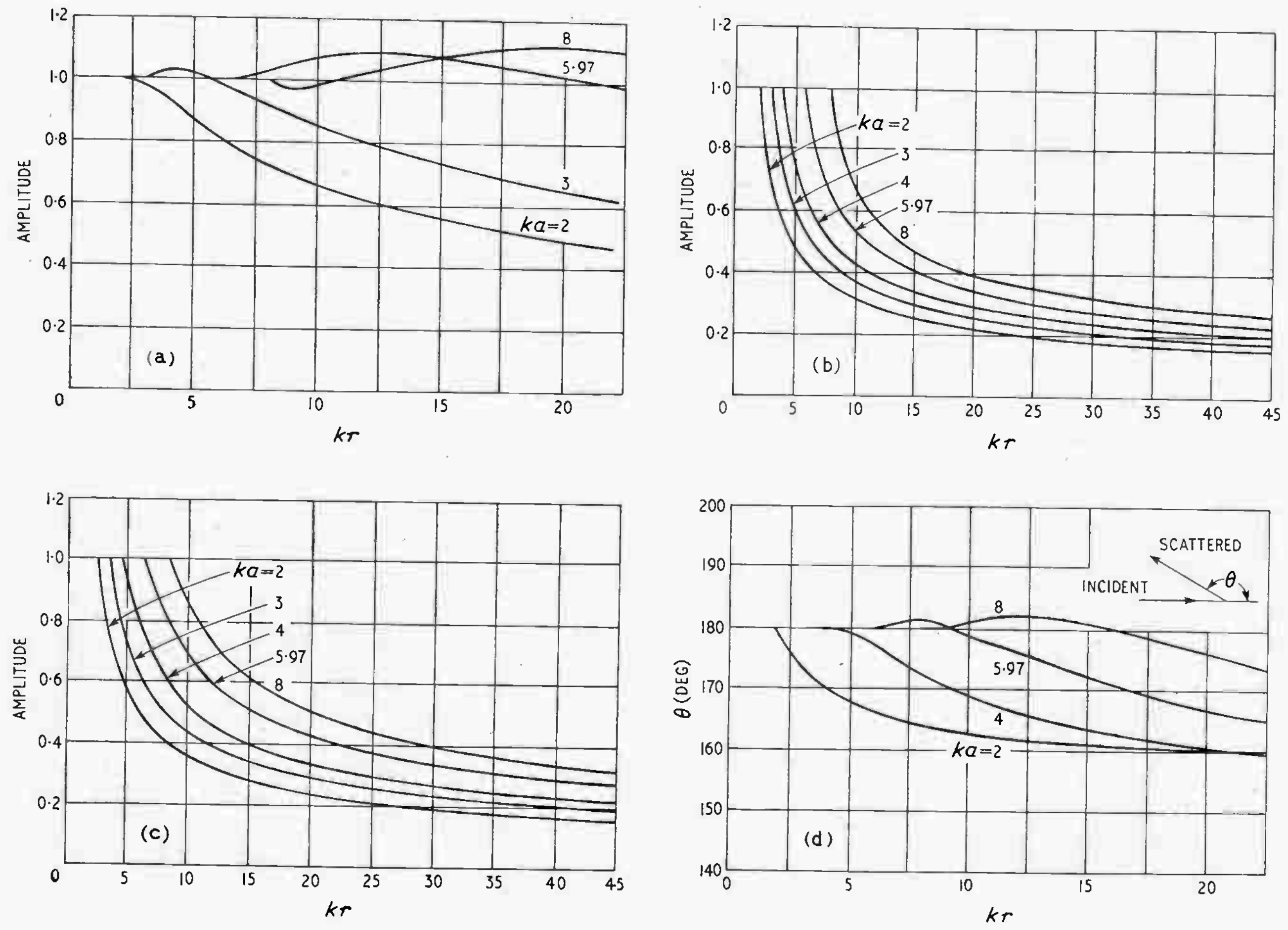
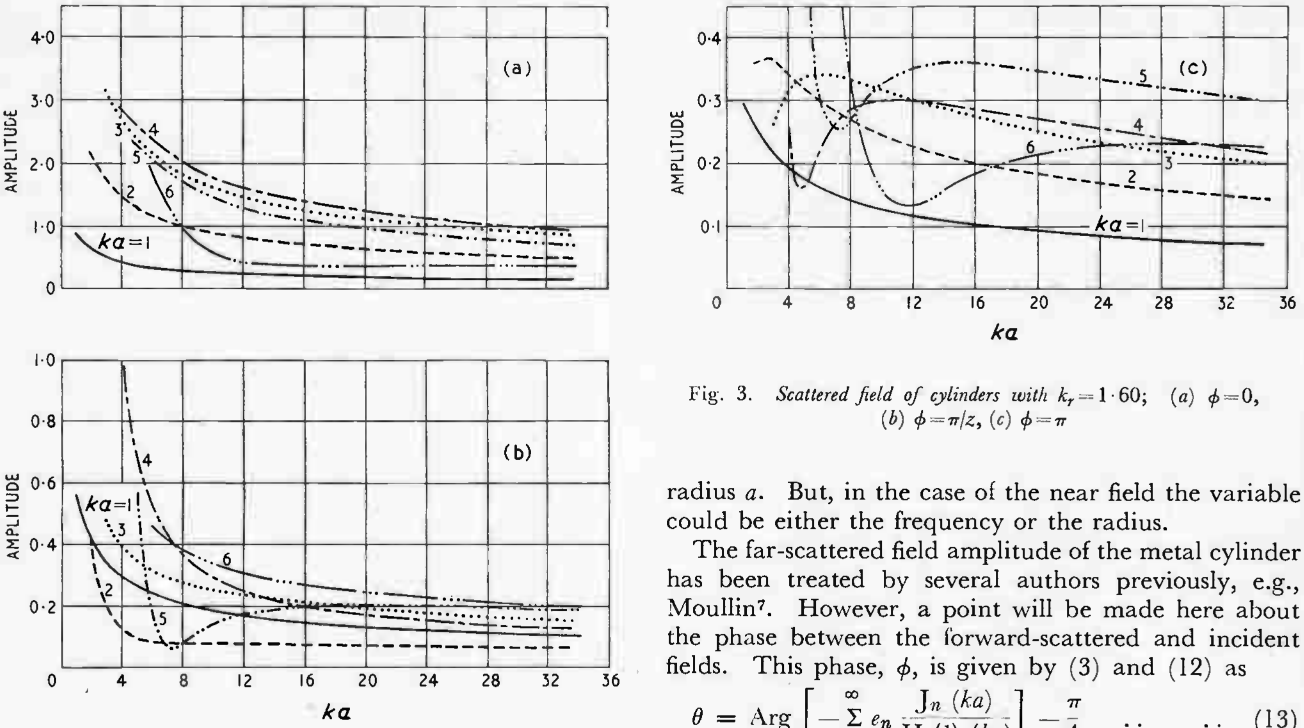


Fig. 2. *Scattered field of conducting cylinders;* (*a*) $\phi = 0$, (*b*) $\phi = \pi/2$, (*c*) $\phi = \pi$, (*d*) *phase for* $\phi = 0$



field appears to have settled down to a regular far-field pattern. The near-field behaviour extends for a smaller distance on the shadow side.

One observes the large values of the scattered field amplitude near the cylinder surface on the shadow side, as well as the large forward-to-back ratio of the scattered field. The latter feature has been noted by several workers in the case of scattering by a dielectric sphere and it has been attributed to the constructive interference in the forward direction, and vice versa, of the fields due to the elementary polarization currents produced in the dielectric material. The cylinder is thus essentially an end-fire radiator.

Far Fields

The number of terms in the series for the scattered field to be used for computational purposes is usually considered to be of the same order as ka. Thus when $kr \gg ka$ in (1), the first term of the asymptotic expansion of the $H_n^{(1)}(kr)$ is sufficient and (1) becomes

$$E_{sc}(r, \phi) \approx \left(\frac{2}{\pi kr}\right)^{\frac{1}{2}} \exp i\left(kr - \frac{\pi}{4}\right) \sum_{0}^{\infty} e_n B'_n \cos n\phi \quad (11)$$

where $B'n = \dfrac{B_n}{(i)^n}$

In what follows when the amplitude of the far-scattered field is referred to, it will be considered to be of the form

$$\text{Amplitude} = \left(\frac{\pi kr}{2}\right)^{\frac{1}{2}} E_{sc}(r, \phi) = \exp i\left(kr - \frac{\pi}{4}\right) \sum_{0}^{\infty} e_n B'_n \cos n\phi \quad .. \quad (12)$$

The far-field results will differ in interpretation from the near-field in that the amplitude will depend on the frequency. Thus, when the abscissa of a curve is given as ka (or k_1a), the variable will be considered to be the radius a. But, in the case of the near field the variable could be either the frequency or the radius.

Fig. 3. *Scattered field of cylinders with $k_r = 1\cdot60$; (a) $\phi = 0$, (b) $\phi = \pi/z$, (c) $\phi = \pi$*

The far-scattered field amplitude of the metal cylinder has been treated by several authors previously, e.g., Moullin[7]. However, a point will be made here about the phase between the forward-scattered and incident fields. This phase, ϕ, is given by (3) and (12) as

$$\theta = \text{Arg}\left[-\sum_{0}^{\infty} e_n \frac{J_n(ka)}{H_n^{(1)}(ka)}\right] - \frac{\pi}{4} \quad .. \quad .. \quad (13)$$

since $E_{INC}(r, 0) = \exp ikr$. For small ka this reduces to

$$\lim_{ka \to 0} \theta = \text{Arg}\left[-i\frac{J_0(ka)}{|Y_0(ka)|}\right] - \frac{\pi}{4} = -\frac{3\pi}{4} \quad .. \quad (14)$$

The limiting value for large ka can be determined from the paper of Papas[8] to be $3\pi/4$.

As is seen from Fig. 4, the antiphase value of θ occurs at $ka \approx 0\cdot3$. For this cylinder there is a departure from antiphase of less than one degree for the kr range $0\cdot3$–50.

A comparison of Fig. 4 with Fig. 2 (d) indicates that the behaviour of θ with increasing ka has been reversed in the near region.

For a cylinder with $k_r = 1\cdot60$ the behaviour of the forward, side and back-scattered fields is shown in Fig. 5, the computations having been made at intervals of $0\cdot25$ in ka.

One significant feature of these curves is the presence of well-defined ripples. Curves of the scattering cross-

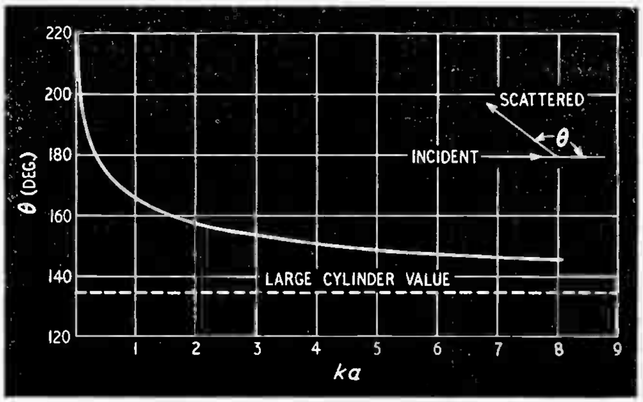


Fig. 4. *Phase of the far forward-scattered field of a conducting cylinder*

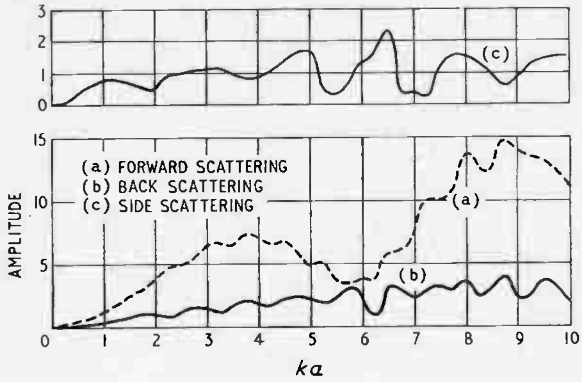


Fig. 5. *Far field scattered by a cylinder with* $k_r = 1\cdot60$; (*a*) $\phi = 0$, (*b*) $\phi = \pi/z$, (*c*) $\phi = \pi$

section for dielectric spheres—and those of Montroll and Hart[9] for cylinders—have usually been drawn as smooth curves. Since, as will be shown later, there is a simple relation between the cross section for scattering and the amplitude of the forward scattered field, it is probable that the latter would have appeared in the same form. Some departures from the smooth form are given by Van de Hulst[10], but only for the initially rising part of the curve. These he attributes to resonances in the sphere. This point will be referred to again later.

A large forward-to-back ratio is again evident, wherein the forward field fluctuates more widely than the back.

The phase of the far, forward-scattered field relative to the incident field has been plotted in Fig. 6. The small and large *ka* limits in this case are $\pi/4$ and $3\pi/4$ radians respectively.

Auxiliary Angles

The concept of auxiliary angles, or phase-shift analysis, was introduced from the field of particle scattering into that of the scattering of light waves by a sphere, by Van de Hulst[10], and by Lax and Feshbach[11] into the problem of scattering by cylinders and spheres. Van de Hulst, however, utilized the concept as a check on his computations and this application was adapted to the present problem of scattering by cylinders.

From (7)

$$T_n = -\frac{B_n}{(i)^n} = \frac{1}{1 + i\,.\,b_n/a_n} \quad \dots \quad \dots \quad (15)$$

and by the transformation

$$\tan\frac{\theta_n}{2} = \frac{a_n}{b_n} \quad \dots \quad \dots \quad \dots \quad \dots \quad \dots \quad (16)$$

(15) becomes

$$T_n = \frac{1}{2}\,(1 - \exp i\,\theta_n) \dots \quad \dots \quad \dots \quad \dots \quad (17)$$

i.e., for real *ka* all the vectors T_n terminate on a circle of radius 1/2 and centre at the point (1/2, 0). This circle property, and the fairly regular behaviour of the angles as compared with the irregular behaviour of the Bessel coefficients, can be utilized as an aid in detecting computational errors.

For the metal cylinder the behaviour of the θ_n, as a function of *ka*, is shown in Fig. 7, for $n = 0$–4. As *ka* increases the curves tend to become linear with a slope of -2, for the angles in radians. This follows since, for large $ka > n$, it may be shown that

$$\theta_n = c\pi - 2\,ka \quad \dots \quad \dots \quad \dots \quad \dots \quad \dots \quad (18)$$

where *c* is not a function of *ka*.

For the dielectric cylinder the curves for the auxiliary angles are not as simple as for the metal cylinder, but still exhibit a pattern which is sufficiently regular to aid in detecting errors. Fig. 8 illustrates their behaviour for $n = 0$ and 1 for a cylinder with $k_r = 1\cdot60$. They occur in pairs, interleaving with each other, because of the relation connecting a Bessel function of one order with the corresponding function of the next higher order. The locations of the rapid rises and the flat parts of the curves are associated with the behaviour of the functions a_n and b_n.

It is profitable to predict the cross-over points of a pair of curves such as those of Fig. 8. In general, for a pair of curves of θ_n and θ_{n+1}, the cross-over values of *ka* are given by the roots of $J_n(k_1a) = 0$ and $J_{n+1}(k_1a) = 0$. Thus the *ka* interval between cross-over points tends towards the value $\phi/2k_r$ for the higher roots.

Scattering Resonances for the Dielectric Cylinder

From the form of (7) it is apparent that the amplitude of the scattering by a dielectric cylinder is governed by the relative values of the a_n and b_n. Thus $-B_n/(i)^n$ can fluctuate between a maximum amplitude of unity, when $b_n = 0$, and zero, when $a_n = 0$. The condition $a_n = b_n = 0$ simultaneously will not occur, because of the Bessel-function Wronskian condition.

For illustration, the functions a_n and b_n have been plotted in Fig. 9 for the first four orders for $k_r = 1\cdot60$. They have been plotted in pairs, to indicate the interleaving effect, mentioned previously with regard to the auxiliary angles. It is seen that, for each function, all the orders tend to fall and rise in step, with the maximum of one function tending to coincide with the zeros of the other. For the near field the process is not simply additive, however, for the several orders, because of the presence of the $(i)^n$ in (7) for the scattered amplitude, but the trend is apparent. When the far field is considered, the $(i)^n$ disappears [see (11)] and the situation

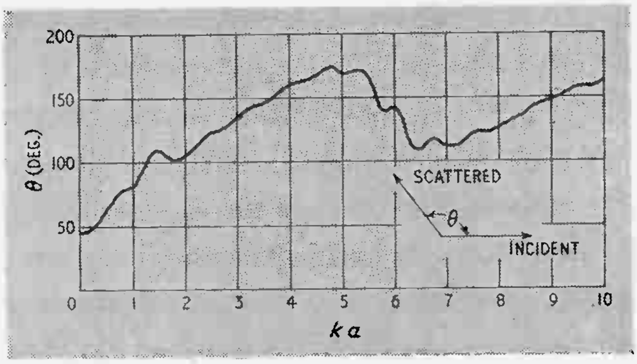


Fig. 6. *Phase of the far forward-scattered field of a cylinder with* $k_r = 1\cdot60$

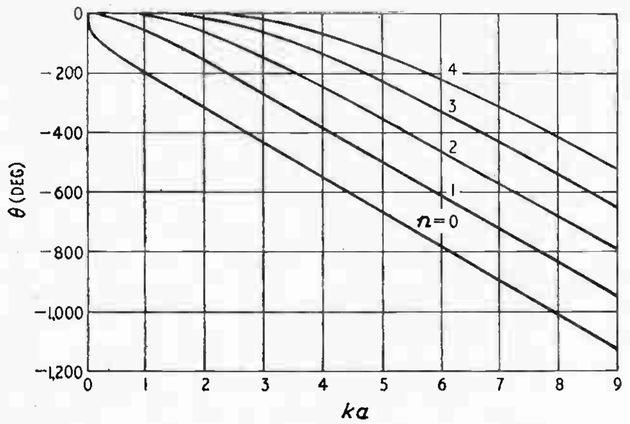


Fig. 7. *Auxiliary angles for conducting cylinders*

is a little more straightforward. Thus, one notes the tendency for the resonance peaks in the forward-field curve of Fig. 5 to coincide with the maxima of a_n and the vanishing of b_n in Fig. 9.

The scattering resonance condition can be derived from a consideration of the fields inside and outside the cylinder under the condition that the incident field has been removed, from the field-continuity relations across the cylinder boundary. This is equivalent to the usual procedure of matching the impedances looking in both directions across the cylinder boundary.

From the previously used series for the scattered and internal fields the boundary conditions yield at the cylinder surface—

$$A_n \mathrm{J}_n(k_1a) = B_n \mathrm{H}_n^{(1)}(ka) \qquad \ldots \quad \ldots \quad (19)$$

$$\frac{k_1}{\mu_1} A_n \mathrm{J}'_n(k_1a) = \frac{k}{\mu} B_n \mathrm{H}_n^{(1)\prime}(ka) \qquad \ldots \quad \ldots \quad (20)$$

The resonance condition can be obtained by division—

$$\frac{k_1}{\mu_1} \mathrm{J}'_n(k_1a)/\mathrm{J}_n(k_1a) = \frac{k}{\mu} \mathrm{H}_n^{(1)\prime}(ka)/\mathrm{H}_n^{(1)}(ka) \qquad (21)$$

which is identical with the vanishing of the expression for Δ in (6). This has been shown to be impossible, at least for real values of the arguments. There can thus be no undamped resonances in the cylinder for, even if the internal and external media were lossless, there would still be radiation damping.

The resonance condition has been discussed by Schaefer and Grossman[12] and by Thilo[13]. The latter arrives at a resonance condition corresponding to the observation previously made—that the resonances (or peaks in the scattering amplitude) are related to the vanishing of b_n—by supposing that the lossless cylinder is surrounded by a metal sleeve at a great distance from the axis to eliminate the radiation damping. The functions for the scattered field are then the standing-wave Y_n instead of the travelling-wave $\mathrm{H}_n^{(1)}$ and (21) becomes equivalent to $b_n = 0$.

Scattering Cross-Section

Interest sometimes lies more in the scattering cross-section of an object than in the scattering amplitude itself. This has been well treated in the case of the metal cylinder but, in view of the scattering resonances found for the dielectric cylinder, the cross-section of the latter will be considered.

This is defined as the ratio of the power scattered per unit length of cylinder to the incident power density. One obtains for the cross-section section σ of the cylinder

$$\frac{\sigma}{2a} = \left(\frac{2}{ka}\right) \sum_{0}^{\infty} e_n \left| B'_n \right|^2 \qquad \ldots \quad \ldots \quad \ldots \quad (22)$$

For a lossless cylinder, this may also be expressed in terms of the auxiliary angles

$$\frac{\sigma}{2a} = \left(\frac{1}{ka}\right) \sum_{0}^{\infty} e_n (1 - \cos \theta_n) \qquad \ldots \quad \ldots \quad (23)$$

Another expression[14, 15, 16] gives a relation between the cross section and the forward-scattered amplitude. For a cylinder this becomes

$$\frac{\sigma}{2a} = -\frac{2}{ka} \mathrm{Re} \sum_{0}^{\infty} e_n B'_n \qquad \ldots \quad \ldots \quad \ldots \quad (24)$$

where Re means "the real part of". If the cylinder is lossy, expression (24) represents the sum of the scattering and absorption cross-sections. If it is lossless, (24) gives the scattering cross-section.

Computations, based on (24), have been made for cylinders with k_r of 1·60 and 2·00. The results are given in Fig. 10. The two curves are similar in that each oscillates about a value of 2 and has the characteristic ripples found on the far-field curves of Fig. 4. The positions of the main maxima and minima correspond closely to those of the far, forward-scattered field.

The secondary ripples still require explaining. Recently, the results of a programme of computations of the scattering cross section for dielectric spheres were announced by Goldberg[4]. These showed the appearance of a systematic ripple on the curves. They had been computed at ka intervals of 0·1. The appearance of the ripple was attributed to this small interval, previous smooth curves having been based on too few computations. Another way of stating it might be that the ripple is due to the manner in which the various orders of the function B'_n make their contribution to the total scattering cross-section. This is equivalent to the resonance explanation of Van de Hulst since, in addition to the main fluctuations in the cross-section curve, there can be minor resonances corresponding to the

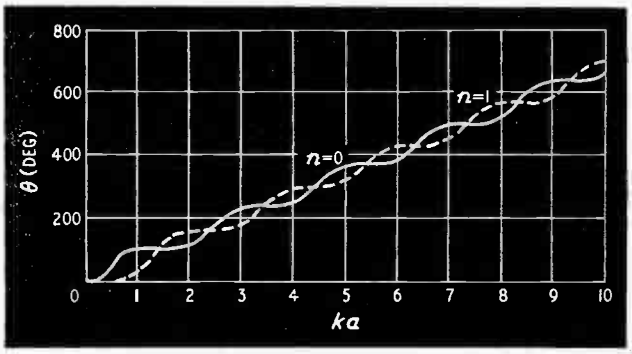


Fig. 8. *Auxiliary angles for cylinders with $k_r = 1 \cdot 60$*

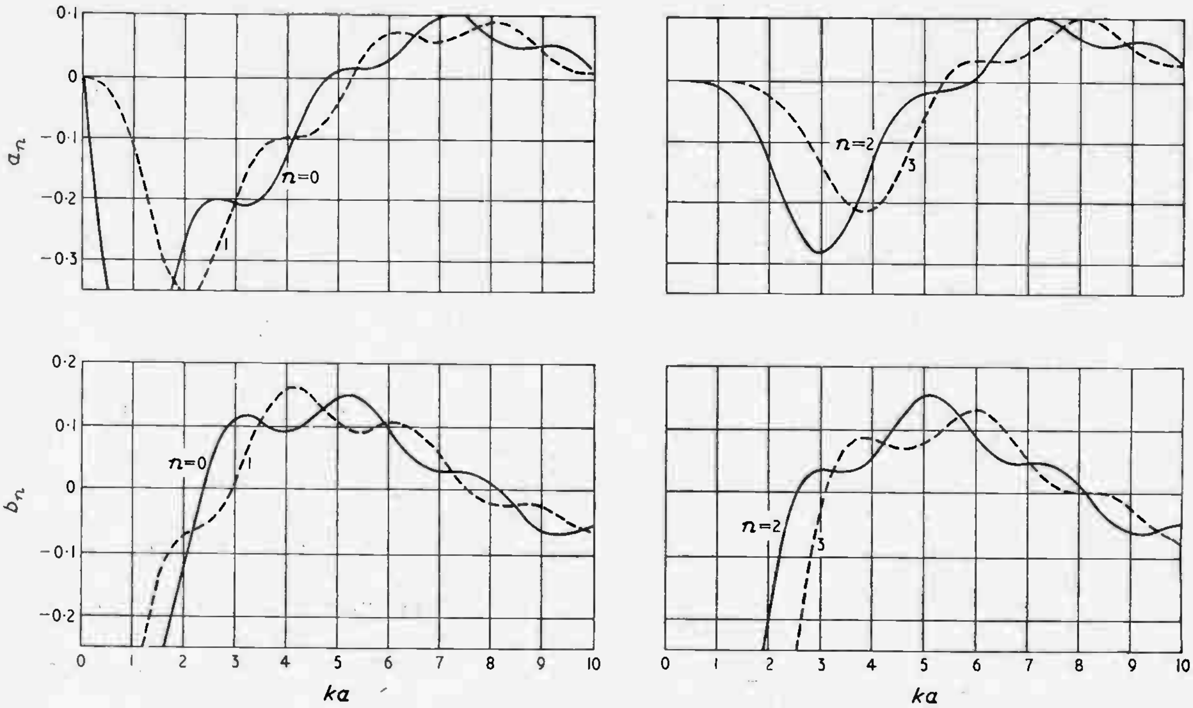


Fig. 9. *Behaviour of the functions* a_n *and* b_n *of Equ.* (7) *for* $k_r = 1\cdot60$

individual behaviour of the various terms of the series.

From the definition of B'_n

$$-\text{Re}\,B'_n = \text{Re}\,\frac{a_n}{a_n + ib_n} = \frac{a_n^2}{a_n^2 + b_n^2} \quad .. \quad (25)$$

This function has been drawn in Fig. 11 for $k_r = 1\cdot60$ for the first four terms of the series, for the values of ka which involve the first main peak of Fig. 10. One notes from (24) that an increase in the contribution of a series term with increasing ka is partially neutralized by the $1/ka$ factor, while a decrease is accentuated. Also, the contributions for $n > 0$ have to be doubled because of the e_n. Then it is possible to account for the ripple in the corresponding curve of Fig. 10 by the relative distributions of the various orders of Fig. 11.

Angular Distribution of the Far-Scattered Field

From (12) computations have been made of the angular distribution of the far-scattered field for cylinders with $k_r = 1\cdot60$. Some of these amplitude results are shown in Fig. 12. The complex lobe structure for the larger cylinders is understandable from the cos $n\phi$ factor, each term in the series being able to give rise to a pair of lobes. The number of lobes on each side of the pattern thus tends to equal the ka value of the cylinder. The high concentration of the scattered field in the forward direction is apparent.

The general lobe structure of the two curves shown was established experimentally in the course of some preliminary development work on a parallel-plate spectrometer operating at a wavelength of 1·25 cm.

Fig. 13 indicates the behaviour of the phase of the series part of (12). The rapid changes of phase are associated with the amplitude depressions.

The corresponding near-field calculations were not attempted, because of the work involved. It would not be sufficient to calculate for only one value of k_r, since it has been seen that the scattered field can vary rapidly with distance from the cylinder. But the information would be useful in enabling one to predict

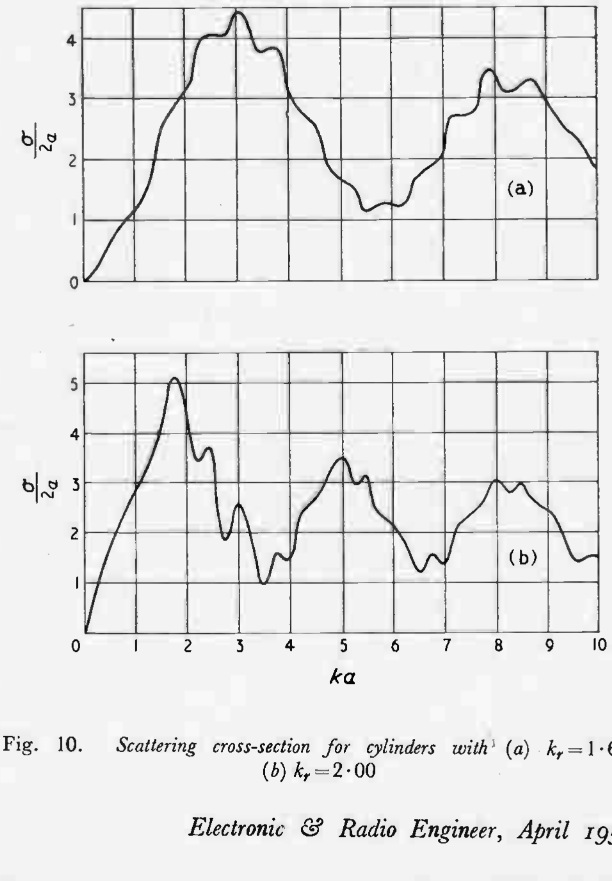


Fig. 10. *Scattering cross-section for cylinders with* (*a*) $k_r = 1\cdot60$, (*b*) $k_r = 2\cdot00$

the interaction between cylinders closely-spaced in an array, at least with respect to their first order or primary scattering.

Cylindrical-Wave Scattering

It is of interest to determine under what conditions departure of the incident wave-front from plane can be neglected and, in addition, the extent to which a curvature of the wave-front adds to the complexity of the scattering calculations.

The configuration of the problem is indicated by Fig. 14. A line source, of as yet unspecified strength A, is situated parallel to the axis of a dielectric cylinder and at a distance b from it. The cylinder parameters and those of the external region are as for the plane-wave case.

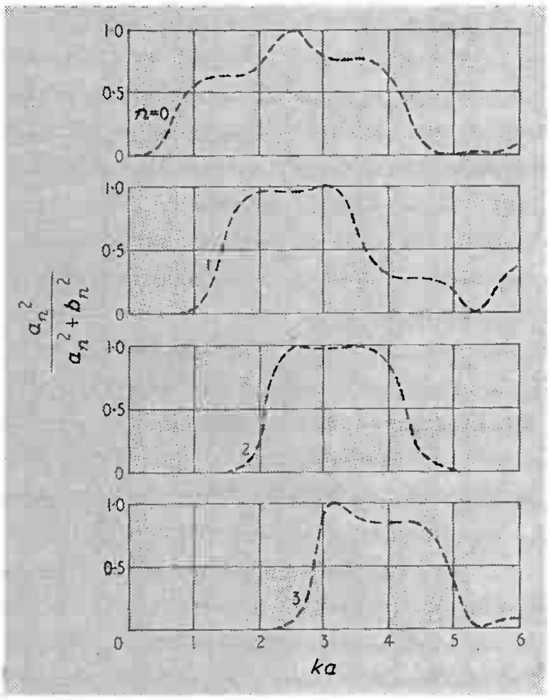


Fig. 11. *Behaviour of the function* $\frac{a_n^2}{a_n^2 + b_n^2}$ *for cylinders with* $k_r = 1{\cdot}60$

The incident wave can be written as

$$E_{INC}(r, \phi) = A\,\mathrm{H}_0^{(1)}(KB)$$

$$= A\sum_0^{\infty} e_n (-i)^n \mathrm{H}_n^{(1)}(kb)\,\mathrm{J}_n(kr)\cos n\phi \qquad (r < b) \qquad (26)$$

$$= A\sum_0^{\infty} e_n (-i)^n \mathrm{H}_n^{(1)}(kr)\,\mathrm{J}_n(kb)\cos n\phi \qquad (r < b) \qquad (27)$$

from Stratton[17], p. 374.

With the scattered and internal fields as before and using the same method, the scattered field amplitude is given by

$$B_n^c = R\,B_n^p \quad \ldots \quad (28)$$

which holds for all r.

The notation c and p refers to curved and plane incident fields respectively. B_n^p is given by (4) and R, which will be called the 'curvature factor', by

$$R = A\,(i)^n\,\mathrm{H}_n^{(1)}(kb) \quad \ldots \quad (29)$$

Since the effect of wave-front curvature, rather than amplitude, is being studied, it is now specified that, in (26), the amplitude always be unity at the origin of co-ordinates when the cylinder is removed. The factor A then becomes $1/\mathrm{H}_0^{(1)}(kb)$ and the curvature factor R:

$$R = (i)^n\,\mathrm{H}_n^{(1)}(kb)/\mathrm{H}_0^{(1)}(kb) \quad \ldots \quad (30)$$

This approaches unity as b approaches infinity (increases indefinitely).

It is thus simple, if the plane-wave results are available, to determine the scattering with a curved incident wave-front.

The factor (30) has been applied to the case of the far-field scattering by a cylinder with $kr = 1{\cdot}60$ and $ka = 3$, already computed for plane-wave incidence, for $kb = 6$ and 15. From the results of Fig. 15 one can note the following effects of the curvature on the angular distribution of the scattering—

(a) For certain values of ϕ the curvature seems to have little effect.

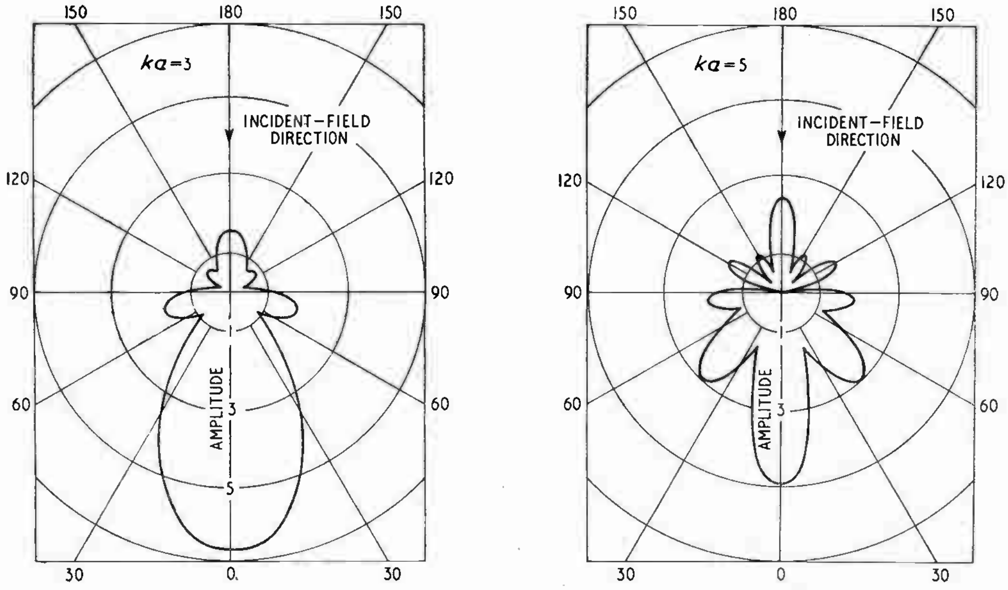


Fig. 12. *Angular distribution of the far-scattered field of cylinders with* $k_r = 1{\cdot}60$

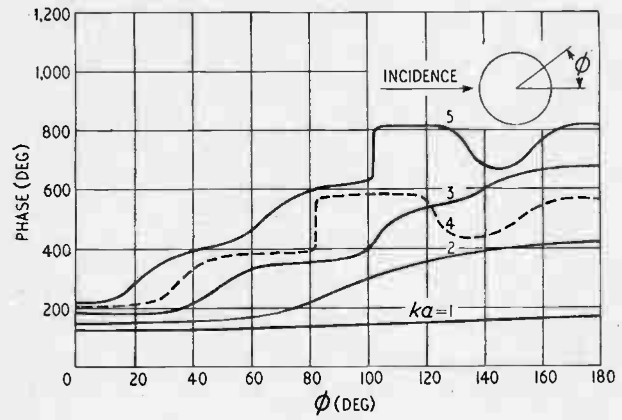


Fig. 13. *Phase of the far-scattered field of cylinders with* $k_r = 1 \cdot 60$

(b) The direction of convergence to the plane-wave conditions, in the case of the amplitude, is not always monotonic.

(c) Rapid rates of change in amplitude are accompanied by corresponding rapid phase changes, as in the plane-wave case.

(d) The forward-scattered field amplitude increases with the curvature, while the opposite holds for the back scattering.

(e) There is some shift in the positions of the side lobes, but it is not always in the same direction.

In the case of the near fields the curvature effect is most marked for back scattering.

From an analysis of the curvature factor it would appear, from the cases considered, that a b/a ratio of the order of 25 would be required to reduce the curvature effect to a negligible value.

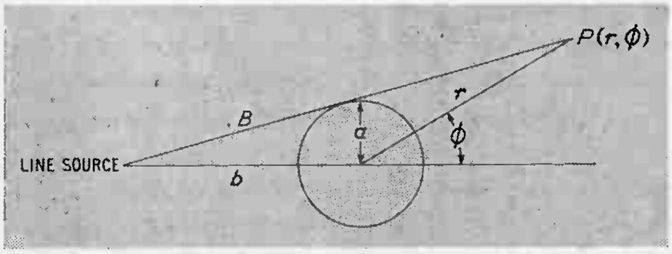
Two observations on the R-factor can be made on the basis of the reciprocity theorem—

(1) If one has calculated the field at the point (b, π) for plane-wave incidence, in the direction $\phi = 0$, then there has also been calculated the far field in the direction $\phi = \pi$ for excitation from a line source at (b, π).

Since the former shows fluctuations about the unobstructed value as b is changed–the amplitude of the fluctuations decreasing with increasing b–then the far field of a line source in front of the cylinder will show the same characteristics.

(2) If one has calculated the field at the point $(b, 0)$ for a plane-wave incident in the direction $\phi = 0$, then there has also been calculated the far field in the direction $\phi = \pi$ for a line source at $(b, 0)$.

Fig. 14. *Configuration for cylindrical-wave scattering*



The former has been shown to reveal a zero field for certain combinations of b and ka and the far field at $\phi = \pi$ should thus be zero for that cylinder and the line source at that value of b. This condition does not seem to occur for metal cylinders (except for the obvious case $b = a$)*.

Internal Field of a Dielectric Cylinder

The internal field of a scattering dielectric cylinder is usually not discussed in papers on the subject, probably because it cannot be measured. If the distribution of this field were known, an expression for the external field could be written, even though it might not be possible to evaluate it.

Rayleigh's method of treating scattering by large cylinders whose electric and magnetic properties differ but slightly from those of the external medium—by considering that the internal field is that existing there with the cylinder absent—would not be valid for the cases being considered here and one is led to inquire as to the complexity, in amplitude and phase, of the internal field in some typical cases.

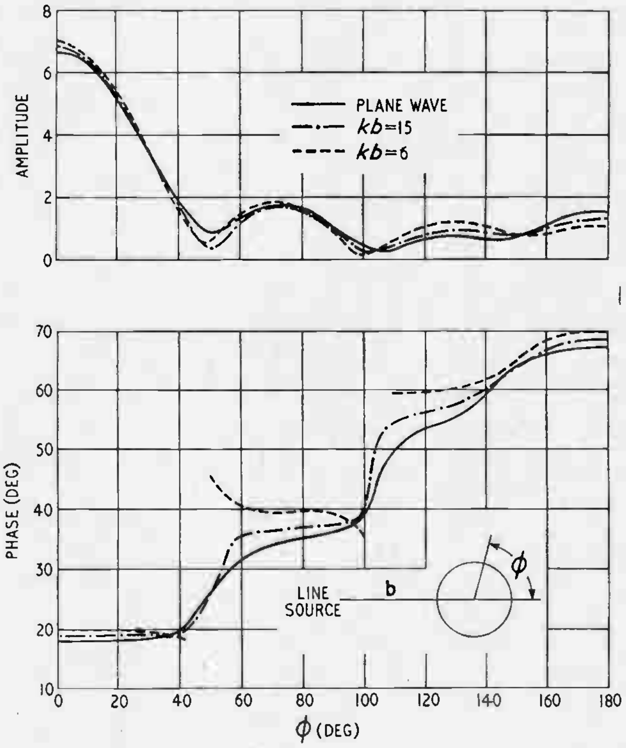


Fig. 15. *Far-field scattering of a cylindrical wave by a cylinder with* $k_r = 1 \cdot 60$, $ka = 3$

The internal field is given by (2), (5) and (6). If only the field on the axis of the cylinder is being considered, only the first term of the series is required.

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* After this work had been completed, a letter by Faran[18], who treated the far field of the acoustic case with boundary conditions corresponding to a metal cylinder in electromagnetics, indicated that the normalizing procedure given in this section had been discussed by Morse *et al*[19]. Faran's numerical results indicated the same trends as found here.

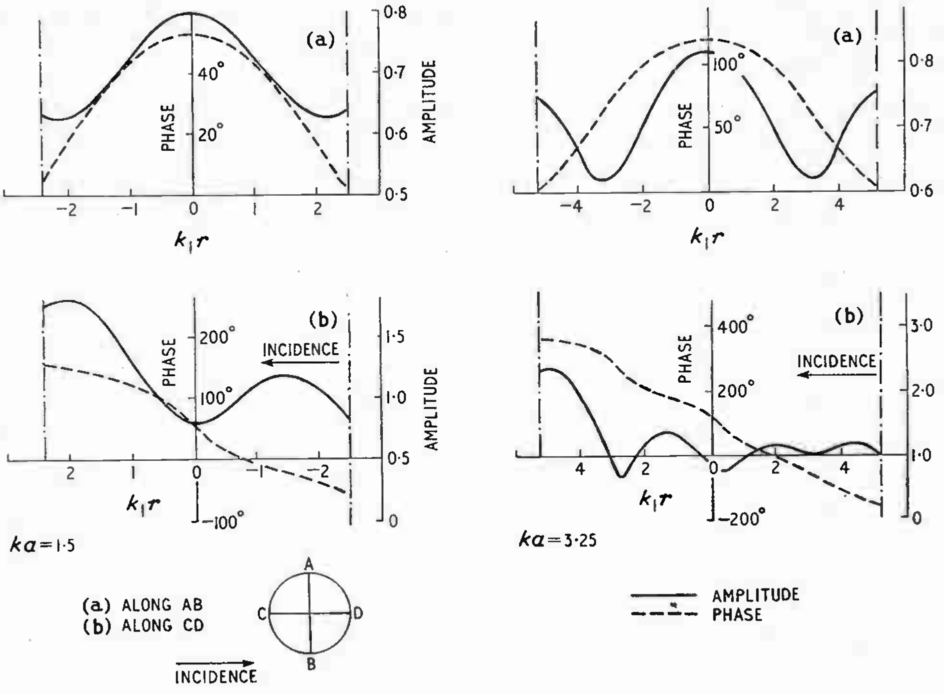


Fig. 16. *Internal field along a diameter of a cylinder of $k_r = 1 \cdot 60$; ———— amplitude; – – – – – phase*

Then

$$E_{INT}{}^{(0)} = A_0 = \left(\frac{2}{i\,\pi\,ka}\right) \Big/ \left[\mathrm{J}_0\,(k_1a)\;\mathrm{H}_1{}^{(1)}\,(ka) - \frac{k_r}{\mu_r}\mathrm{J}_1\,(k_1a)\;\mathrm{H}_0{}^{(1)}\,(ka)\right] \qquad (31)$$

where $\mu_r = \mu_1/\mu$, $\epsilon_r = \epsilon_1/\epsilon$, $k_r = \sqrt{\epsilon_r\,\mu_r}$.

For small cylinders this reduces to—

$$E_{INT}{}^{0}/k_1a \text{ Small} = \frac{4}{i}\left[\frac{1}{\pi\,\epsilon_r\,\mu_r\,(ka)^2\,(1-\epsilon_r) - 2i\left\{2 - \epsilon_r\,(ka)^2 \log_e \dfrac{2}{\gamma\,ka}\right\}}\right] \qquad (32)$$

where γ is Euler's constant.

For large cylinders it becomes—

$$E_{INT}{}^{(0)}/k_1a \text{ large} = \sqrt{k_r}\,e^{i\,\tau\pi/4}\,E_s \quad \ldots \quad \ldots \qquad (33)$$

where $\tau = (k_r - 1)/k_r$. E_s is the field at the centre of a parallel-sided slab of the same material as that of the cylinder, centred on the origin, of half-thickness d given by—

$$k_1d = k_1a - \pi/4$$

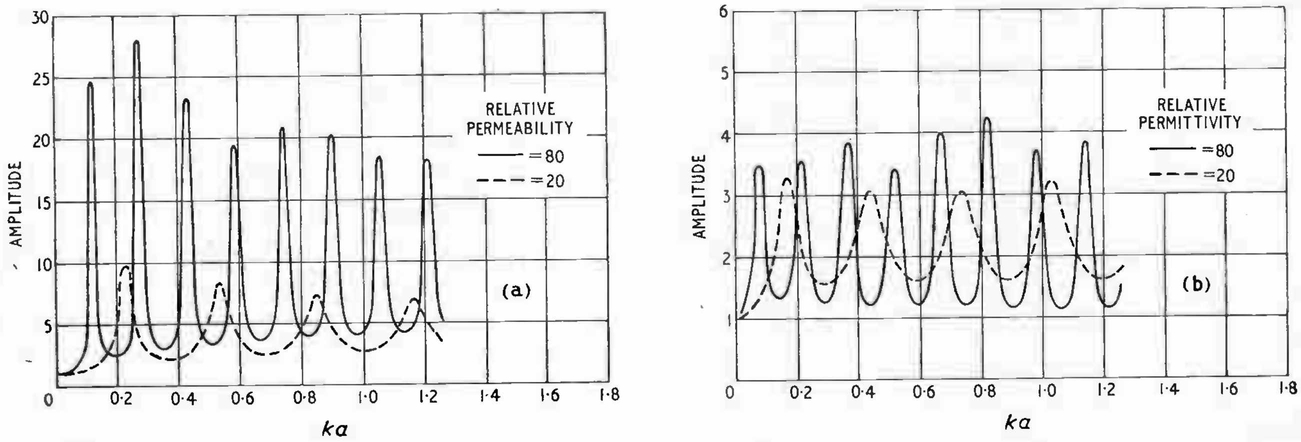
and on which a plane wave of unit amplitude is incident normally from medium k. Calculations for a slab and a cylinder of $\mu_r = 1$ and $k_r = 1 \cdot 60$ showed these slab-cylinder relations to hold, even for cylinders less than one external wavelength in diameter. Also, whereas the slab-field amplitude, plotted as a function of ka, never exceeds the free-space value, the cylinder field oscillates about it between the values k_r and $1/k_r$ (for the values of μ_r used in these calculations).

Additional calculations were made for a cylinder of this material. The fields are taken along two mutually perpendicular diameters, along and normal to the direction of incidence. They were made to trace the variation in the internal field pattern through the equivalent of a period in the thickness of the parallel-sided slab. In addition there was the possibility of observing some aspect of correlation between the variation of the form of the internal field and that found externally, particularly for the radii giving the deep shadow effect.

Some of the results are given in Fig. 16. No correlation was observed, to the extent that no marked change in the field pattern appeared along the two diameters considered. But the correlation must obviously exist, when the complete cylinder cross-section is considered.

The curves show also (a) the standing wave of

Fig. 17. *Variation with radius of the amplitude of the field at the axis of a cylinder with (a) $\epsilon_r = 5$, (b) $\mu_r = 5$*



amplitude along the incidence diameter, the repetitions increasing in number with the radius and (b) the tendency for the field to be concentrated on the shadow side of the cross-section, in a form of focusing effect, with a large peak just inside the shadow surface.

Resonances of the Internal Field

It has been seen that the scattering from a dielectric cylinder is governed by the resonance equation—

$$\frac{J'_n\,(k_1a)}{J_n\,(k_1a)} = \sqrt{\frac{\mu_r}{\epsilon_r}}\,\frac{Y'_n\,(ka)}{Y_n\,(ka)} \qquad \ldots \quad \ldots \quad \ldots \quad (34)$$

which reduces, for small cylinders, to—

$$\frac{J_1\,(k_1a)}{J_0\,(k_1a)} = \sqrt{\frac{\mu_r}{\epsilon_r}}\,\frac{Y_1\,(ka)}{Y_0\,(ka)} \qquad \ldots \quad \ldots \quad \ldots \quad (35)$$

If $\mu_r \gg \epsilon_r$, this latter condition corresponds to the zeros of $J_0(k_1a)$. If $\mu_r \ll \epsilon_r$, the zeros of $J_1(k_1a)$ control. In this case, however, there is the further possibility of resonance corresponding to the tendency of μ_r/ϵ_r and $Y_1\,(ka)/Y_0\,(ka)$ to balance each other at a finite value.

The two cases are illustrated by the curves of Fig. 17 for the axial field. As for the slab case, the spacing between the peaks tends toward π/k_r. The curves oscillate about the value $\sqrt{\mu_r}$. All the peaks, except for the first one for each curve of the second set, correspond to the zeros of the Bessel functions, as discussed above. This peak corresponds to the balancing tendency mentioned.

The results of Fig. 17 can be compared with those of Page[20] for the other polarization of the magnetic field parallel to the cylinder axis. There it was found that a plot of the total flux through a cross-section showed a peak for one value of the variable ka (except for the case $\mu_r = 100$, the highest value considered, when some secondary ripples appeared). Page's interest was in the efficiency of the rods as cores for loop receiving aerials. The analogous application here would be that of a very-thin-wire receiving aerial, coated with a layer of high-μ material, of such a thickness as to produce a resonance on the axis ; i.e., at the wire.

Acknowledgements

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The experimental results corresponding to the calculations of Fig. 12 were obtained by Dr. P. Sollom, whose assistance is acknowledged.

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New Books

Transistor Circuit Engineering

Edited by Richard F. Shea. Pp. 468 + xx. John Wiley & Sons Inc., available in U.K. from Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 96s.

In their first book "Principles of Transistor Circuits", Shea and his co-authors went as far as they were able at that time to show how the transistor would be used for circuit design with a background knowledge of the physics of the device and a mental picture of its operation. In this book, they have departed radically from their previous approach in that practically no mention is made of how the transistor works and very little information is given about the physics of the transistor. Their approach may be called the 'black box' approach in that the transistor is regarded as a box having certain published characteristics which enable the engineer to design circuits without knowing what is going on inside the box. The design of linear circuits is based on the *h*-parameters and the use of matrix methods of analysis is advocated. The design of non-linear switching circuits is based on the negative-resistance characteristic produced either by internal feedback in the device or external feedback in the circuit.

This black-box approach is advantageous to the engineer experienced in valve circuit design, as he does not need to spend time learning about the transistor, but it has a considerable disadvantage as it is not possible to express many of the dynamic characteristics of these new devices very accurately by published curves and limits. This means that the engineer using the black-box approach will meet many anomalies in his designs which may only be conveniently explained by a basic knowledge of the operation of the transistor and an elementary knowledge of the physics of the device.

Engineers who wish to use the *h*-parameter approach to the design of transistor circuits will find this a useful book as it gives a good basic introduction to a wide variety of applications. Most of the book is confined to the use of the junction transistor, and the obsolete point-contact transistor is almost absent. Some mention is made of the tetrode and the double-base diode.

Chapter one deals with characteristics and characteristic curves and may be summed up by quoting from the introduction—"The utilization of semiconductor devices is most efficiently realized by the combination of the general technical knowledge of the engineer and the specific engineering data supplied by the manufacturer." Some of the characteristic curves are then briefly described.

A short description of the *h*-parameters and the principle transistor equivalent circuits takes the reader to chapter three where bias and its stabilization is mathematically analysed using current and voltage stability factors. Chapter four gives a comprehensive treatment of audio amplifiers.

D.C. amplifiers are interposed between audio amplifiers and tuned amplifiers, and are designed by the use of compensation or chopping techniques. Chapter six, on tuned amplifiers, analyses coupling networks first and then goes on to the transistor itself and neutralization, the major part being devoted to the networks.

Low-frequency compensation and high-frequency peaking are used as the basis for chapter seven on video amplifiers. Chapters eight and nine briefly describe a number of harmonic oscillators, relaxation oscillators, the known methods of modulation applied to transistors, mixers and various transistor detectors including regenerative detectors.

Transient and pulse characteristics were omitted in chapters one and two and are described in the first half of chapter ten. The second half describes transistor gates and pulse circuits based on the Eccles-Jordan circuit and the transformer-coupled pulse-regenerative amplifier or blocking oscillator.

In chapter eleven, entitled "Systems," the section on a.m. receivers is useful, but the other sections on carrier-current systems, f.m. receivers and television receivers are not very helpful, as they are rather ahead of their time in that no details are given of the use of the new v.h.f. drift transistor.

The final chapter on special circuits describes low-level limiters, transistor active filters, d.c. converters and regulated power supplies.

Shea's first book came as a very useful introduction to the subject and a revision of that book would have been very helpful. In compiling this new book, too much has been lost from the old to make it a useful introduction for beginners and not enough new material has been added to make it a standard reference for experienced engineers.

E.W.

Low-Cost High-Quality Amplifier. With Alternative Pre-amplifiers.

By P. J. Baxandall, B.Sc.(Eng.). Pp. 29. Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 3s. 6d. (by post 4s.).

A collection of articles reprinted from *Wireless World.*

A statistical analysis by the designer of the main amplifier described in this booklet has shown that an undistorted output of more than 5 watts is rarely called for, even by sound-reproduction enthusiasts when given a free hand to adjust the volume level to their own liking.

Careful choice of operating conditions for the output valves, and the application of adequate feedback has enabled an output transformer of comparatively simple design to be employed with complete stability.

Frequency-Modulated Radio (2nd Edition)

By K. R. Sturley, Ph.D., M.I.E.E., Sen.M.I.R.E. Pp. 120. George Newnes Ltd., Tower House, Southampton Street, London, W.C.2. Price 15s.

Written for 'the practical man, whose job is to make an f.m. receiver work', this book nevertheless contains one chapter on general principles which discusses f.m. waves by means of vectors and explains the advantage of f.m. systems over a.m. systems. Another chapter deals with methods of producing frequency modulation. The remaining three chapters are devoted to reception and testing receivers, the last being a description of a typical combined f.m./a.m. receiver.

Transistor Electronics

By David De Witt and Arthur L. Rossoff. Pp. 381. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 60s.

The earlier chapters deal with semiconductor physics. The authors go on to discuss transistor characteristics and equivalent circuits. The major part of the book is devoted to circuitry (10 chapters) and there is one chapter on special types of transistor and one on noise. Almost every application of transistors is included, though the treatment is necessarily rather brief at times.

Synthesis of Passive Networks

By Ernst A. Guillemin. Pp. 741. John Wiley & Sons Inc., available in the U.K. from Chapman & Hall, 37 Essex Street, London, W.C.2. Price £6.

Chapters on Properties of Driving-Point and Transfer Impedances; Driving-Point and Transfer Functions of Two-Element-Kind Networks; Synthesis of *LC* Driving-Point Impedances; Synthesis of *RC* and *RL* Driving-Point Impedances; More about Equivalent and Reciprocal Networks; Properties of Two Terminal-Pair Networks; Real-Part Sufficiency and Related Topics; Synthesis of *RLC* Driving-Point Impedances; Transformerless Driving-Point Impedance Synthesis; Conventional Methods of Transfer Function Synthesis; Other Methods of Realizing Transfer Functions; *RC* Transfer Function Synthesis; The Approximation Problem and Time-Domain Synthesis.

Portable Transistor Receiver

By S. W. Amos, B.Sc.(Hons.), A.M.I.E.E. Pp. 15. Published for *Wireless World* by Iliffe & Sons Ltd. Price 2s. 6d. (By post 2s. 10d.)

This reprint from *Wireless World* describes the theoretical and practical design of a battery-operated, self-contained, portable superheterodyne receiver using a total of seven junction transistors, four of which are of the a.f. type and three of the latest r.f. type. A point-contact diode is also included.

The Ultra-High Frequency Performance of Receiving Valves

By W. E. Benham, B.Sc., F.Inst.P., and I. A. Harris, A.M.I.E.E., A.M.Brit.I.R.E. Pp. 173. Macdonald & Co. Ltd., 16 Maddox Street, London, W.1. Price 28s.

Deals theoretically with conventional space-charge control valves for use up to about 3000 Mc/s.

Telecommunications

By W. Fraser, B.Sc.(Eng.), A.M.I.E.E. Pp. 772. Macdonald & Company (Publishers) Ltd., 16 Maddox Street, London, W.1. Price 65s.

A textbook for students who intend to sit B.Sc. or equivalent examinations. The 22 chapters include one on telephone systems and one on telegraph systems. Radio coverage includes waveguides, u.h.f. valves, propagation and aerials, and 'radio transmission systems'. Questions from examination papers are given at the end of each chapter, and there are some worked examples in the mathematical chapters.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)
Deviations from nominal frequency for February* 1958

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

* *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. N.T. = No Transmission.*

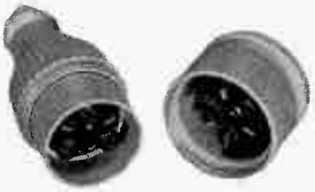
New Products

Plugs and Sockets

A new range of plugs and sockets, which meets stringent electrical and mechanical requirements, is to be manufactured in the U.K. under licence from Tuchel Kontact, Germany.

The patented contact principle employed in this new range of connectors is said to differ radically from more traditional systems in that the construction is based on the theory that surfaces kept free from oxide film and dirt are more important factors than contact pressure.

The plugs and sockets available range from small ¾-in. diameter units with six poles or less, each carrying 5 A at 125 V, to larger 5-in. diameter units with 36 poles, each carrying 60 A at 500 V. The connector illustrated is rated at 20 A, 1,000 V per pole.



Modern Acoustics Ltd.,
Manor Way, Boreham Wood, Herts.

Disturbance Recorder

The instrument is described as an oscillographic chart recorder with memory and other special features which should make it suitable for applications in industrial-control schemes apart from its normal use of fault recording on electricity transmission systems.

Seven oscillographic traces and twelve d.c. marker traces are made continuously on a drum but only transferred to paper on receipt of a signal.

The memory feature causes the chart to record the conditions starting 0·5 second before the signal. The chart speed is approximately 6 inches per second.

The chart can be handled immediately and the markings are permanent. The instrument operates for over a year without re-inking. Oscillographic elements are suitable for voltage or current measurements on 50-cycle supplies. A full range of accessories is available.

Ferranti Ltd.,
Hollinwood, Lancs.

New Klystrons

The K 345 is described as a newly-developed medium-power reflex klystron oscillator giving an output of 1·2 W, and is an equivalent of the American type VA 220. The K 345 has been designed particularly for microwave links. Good linearity of the frequency versus reflector voltage characteristic ensures low f.m. distortion. Frequency stability is improved by the addition of an external tuning cavity.

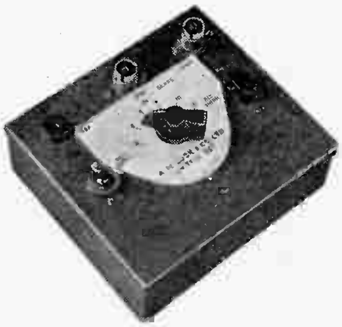
There are seven frequency variants which together cover the band from 5925 to 8025 Mc/s with one small gap.

Type No.	*Mechanical Tuning Ranges*
K 345Z	7725 — 8025 Mc/s
K 345A	7425 — 7725 ,,
K 345B	7125 — 7425 ,,
K 345C	6875 — 7125 ,,
K 345D	6575 — 6875 ,,
K 345E	6225 — 6525 ,,
K 345F	5925 — 6225 ,,

English Electric Valve Co. Ltd.,
Chelmsford, Essex.

pH Meter Test Set

An inexpensive instrument which enables pH meters to be checked quickly by unskilled personnel has recently been introduced. It consists of a dry cell feeding a potential divider made up of high-stability resistors. A switch selects various outputs, which correspond to steps of about 2 pH. The output may be taken at low or high (250 MΩ) impedance.

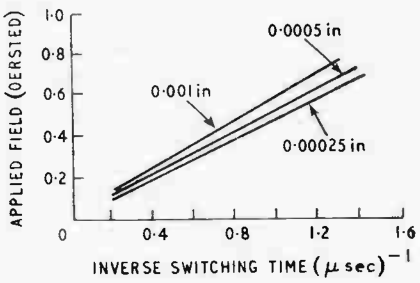


A. M. Lock & Co. Ltd.,
Prudential Buildings, 79 Union Street,
Oldham, Lancs.

Microstrip Magnetic Cores

Spirally-wound cores of very thin strips of magnetic material are now available from T.M.C. The materials used are mumetal and a new alloy—H.S. alloy. Strip thicknesses are 0·001 in., 0·0005 in., and 0·00025 in. and strip widths ⅛ in. and ¼ in.

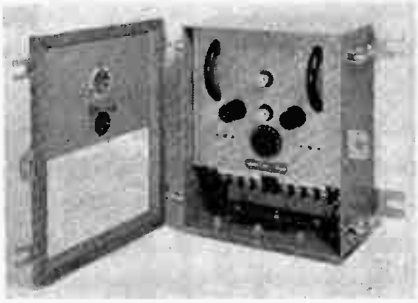
The cores are intended mainly for pulse applications, but can also be used for inductors up to at least 10 Mc/s. H.S. alloy has a fast switching time; i.e., the tim taken to drive a core from one state o remanence to the opposite state. The switching time varies with applied field and the strip thickness. The accompanying graph shows how the reciprocal of the switching time varies with the applied field, and it will be seen that a typical switching time is 1 μsec.



The Telegraph Construction & Maintenance Co. Ltd.,
Telcon Works, Manor Royal, Crawley, Sussex.

Level-Controller

The Fielden 'Tektor Major' is a new capacity-operated level-controller which is claimed to be easy to install, reliable and sensitive. The capacitor formed by the probe electrodes is one arm of a capacitance bridge which is located in the positive-feedback loop of an oscillator. A change in the probe capacitance therefore alters the amplitude of oscillation.



In this instrument, the oscillator is amplitude-modulated and the change in output of the detected modulation-frequency is made to operate a trigger circuit. The sensitivity is better than 1 pF and is said to be virtually unaffected by a ± 10% mains-voltage variation, an ambient temperature change of −20°C to +50°C, or lapse of time. Zero drift is small.

The detector operates a contactor-type relay with a switching capacity of 440 V, 5 A a.c., and this can be used for control purposes. The capacitance probe may be located up to 35 feet from the instrument proper. Operation from mains supplies of 110–400 V, 40–60 c/s is catered for. The instrument is housed in a waterproof metal case.

Fielden Electronics Ltd.,
Wythenshawe, Manchester.

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 : 541.135 **991**

U-Effect, II, an Electrokinetic Phenomenon.—W. W. Fain, S. L. Brown & A. E. Lockenvitz. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 902–908.) "When a glass capillary tube filled with a column of alternate layers of mercury and an electrolytic solution is forced to vibrate mechanically, an a.c. voltage is generated across the ends of the column. Certain aspects of this electrokinetic phenomenon have been investigated both experimentally and theoretically."

534.2-14 **992**

Comparison of Experimental Underwater Acoustic Intensities of Frequency 14·5 kc/s with Values Computed for Selected Thermal Conditions in the Sea.—F. H. Sagar. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 948–965.) See 1277 of 1956.

534.2-8 **993**

Rendering Standing Ultrasonic Fields Visible and Acoustic-Optical Image Conversion.—G. Keck. (*Acustica*, 1956, Vol. 6, No. 6, pp. 543–548. In German.) A survey of existing methods of exploring ultrasonic fields with further details of new photographic methods [see e.g. 963 of 1956 (Hauer & Keck)].

534.23 + 621.396.677 **994**

Some Aspects of the Design of Strip Arrays.—D. G. Tucker. (*Acustica*, 1956, Vol. 6, No. 5, pp. 403–411.) The design and performance of linear arrays are examined from the point of view of far-field directional patterns. Patterns are synthesized by the linear superposition of curves of the $\sin x/x$ type; this method clarifies the relation between directional pattern and the distribution of excitation (or sensitivity, in the case of reception) over the length of the array.

534.232 : 537.228.1 **995**

Piezoelectric Transducers.—A. C. Dobelli. (*Acustica*, 1956, Vol. 6, No. 4, pp. 346–356.) Review of the properties and parameters of the most widely used piezoelectric materials. Typical applications are mentioned and the choice of materials and the appropriate dimensions for the element are discussed. Reference is made to recent developments of piezoelectric polycrystalline ceramics.

534.232 : 621.317.35 **996**

Measurements on Electromechanical Transducers.—Diestel. (See 1214.)

534.232 : 621.372.5 **997**

The Ideal Forms of Electroacoustic Transducers and the Characteristics of Coupled Systems Formed by Them.—F. A. Fischer. (*Acustica*, 1956, Vol. 6, No. 5, pp. 421–424. In German.) The ideal form of the four classes of electroacoustic transducer (piezoelectric, dielectric, electrodynamic and electromagnetic) is discussed. The characteristics of the electric quadripoles resulting from the mechanical coupling of two ideal transducers of identical or different classes are investigated.

534.52 **998**

Scattering of Sound by Sound.—P. J. Westervelt. (*J. acoust Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 934–935.) Earlier analysis of the mutual nonlinear interaction of two plane sound waves (2657 of 1957) is extended to include arbitrary directions of travel of one wave with respect to the other. An exact solution for the first-order scattering process is obtained.

534.612 **999**

Acoustical Radiation Pressure due to Incident Plane Progressive Waves on Spherical Objects.—G. Maidanik & P. J. Westervelt. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 936–940.)

534.615 **1000**

Automatic Recorder of Lines of Equal Phase Change.—V. Gavreau & A. Calaora. (*Acustica*, 1956, Vol. 6, No. 6, pp. 539–542. In French.) Apparatus for tracing curves of equal phase change in a plane section of a three-dimensional sound field is described. Its application to the sound field of a loudspeaker and the effects of interposed obstacles are illustrated.

534.641 : 621.395.92 **1001**

The Acoustical Impedance Presented by some Human Ears to Hearing-Aid Earphones of the Insert Type.—J. Y. Morton & R. A. Jones. (*Acustica*, 1956, Vol. 6, No. 4, pp. 339–345.) Report of measurements made over the frequency range 220 c/s–4 kc/s on persons of normal hearing. The impedance of the ear at the ear drum is calculated from that measured 1·6 cm from the ear drum with a complete seal between the ear mould and the ear canal. A continuously variable acoustic impedance developed for the investigation is described.

534.75 **1002**

Recorded Group Audiometer Test Comparisons at the 1956 Southern California Exposition.—J. C. Webster & P. O. Thompson. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 895–899.)

534.75 : 621.391 **1003**

The Basic Elements for Determining Information Capacity in Hearing.—E. Zwicker. (*Acustica*, 1956, Vol. 6, No. 4, pp. 365–381. In German.) Audibility changes are considered as a function of just perceptible levels of a.m. and f.m. for pure tones or white noise. By means of a model based on the perception of a change in sound intensity of 1 dB within a group of frequencies with a mean time constant of 20 ms, the effects of phase differences as well as masking are explained. The maximum information perceptible is estimated and its application to the intelligibility of syllables is illustrated.

534.846 **1004**

Orthophonic Surfaces in Auditorium Design.—C. Codegone. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 885–888.) " A study is made, in a particular case, of the shape to be given to the ceiling of a large auditorium such that the sum of the direct intensity and the intensity reflected by the ceiling is constant."

621.395.61 : 534.844.1 **1005**

Investigation of the Directional Characteristics of the Pick-Up Microphone in Pulse Measurements of Room Acoustics.—H. Niese. (*Hochfrequenztech. u. Elektroakust.*, May 1957, Vol. 65, No. 6, pp. 192–200.) A stereophonic microphone simulating the directional characteristics of the human head is described, which is used in tests to assess subjectively the effect of the angle of incidence of echoes, and of amplitude differences at the ears. See also 2658 of 1957.

621.395.623.5 **1006**

Wedge-Shaped Acoustic Horns for Underwater Sound Applications.—C. M. McKinney & W. R. Owens. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 940–947.) Report of measurements on horns constructed of rubber-covered Al plates [1241 of 1955 (McKinney & Anderson)] modified to operate at 10 kc/s.

621.395.623.64 + 534.833 **1007**

Attenuation of Ear Protectors by Loudness Balance and Threshold Methods.—J. Hershkowitz & L. M. Levine. (*J. acoust. Soc. Amer.*, Aug. 1957, Vol. 29, No. 8, pp. 889–894.) Results of measurements on earmuffs and earphone sockets using three different methods are compared.

621.395.623.7 **1008**

Design for a Folded Corner Horn.—H. J. F. Crabbe. (*Wireless World*, Feb. 1958, Vol. 64, No. 2, pp. 57–62.) Details are given of the construction of low- and middle-frequency horns in a compact unit to match an 8-in. speaker, for domestic use.

621.395.625.3 **1009**

The Infrared Transparency of Magnetic Tracks.—G. Lewin. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1957, Vol. 66, No. 9, pp. 517–522. Discussion, p. 522.) Magnetic tracks are substantially transparent to infrared light. The infrared sensitivity of the PbS photoconductive cell enables excellent reproduction to be obtained from an optical track completely covered by a magnetic stripe, eliminating the need for half-width stripes and giving superior magnetic recording quality.

621.395.625.3 : 389.6 **1010**

The International Standardization of Magnetic-Tape Recording.—P. H. Werner. (*Tech. Mitt. PTT*, 1st July 1957, Vol. 35, No. 7, pp. 266–273. In French & German.) Summary of the principal international standards for professional and amateur recordings, including C.C.I.R. recommendations.

621.395.625.3 : 681.85 **1011**

Magnetic Recording Tape—Manufacture and Properties.—H. G. M. Spratt. (*Brit. Commun. Electronics*, July 1957, Vol. 4, No. 7, pp. 418–421.) Tables of representative British recording tapes and magnetic-tape base materials are given.

AERIALS AND TRANSMISSION LINES

621.315.212 **1012**

Coaxial Transmission Lines.—S. Mahapatra. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 63–67.) Approximate calculations are made for the distributed constants R, L, G and C of a coaxial transmission line with the inner conductor of elliptical cross-section. In conditions of restricted space higher Q and lower attenuation may be possible than with a conventional coaxial line.

621.315.212.029.63 : 621.316.543 **1013**

A Contactless Wide-Band Switch for 20-cm Wavelengths.—L. Mollwo. (*Hochfrequenztech. u. Elektroakust.*, May 1957, Vol. 65, No. 6, pp. 181–188.) The mechanical construction and electrical characteristics of a rotary-type coaxial-line switch with matching stubs are detailed. The pass band is 80 Mc/s at 19·5 cm λ with more than 20 dB attenuation of the blocked circuit.

621.315.212.029.63 : 621.372.512]
+621.372.832.43 **1014**

Criteria for the Design of Loop-Type Directional Couplers for the L Band.—P. P. Lombardini, R. F. Schwartz & P. J. Kelly. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 234–239. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, pp. 113–114.)

621.372.029.64 : 537.226 **1015**

The Use of Dielectric Materials to Enhance the Reflectivity of a Surface at Microwave Frequencies.—G. B. Walker & J. T. Hyman. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 73–76.) The condition that the reflection at the dielectric shall be better than at the (metal) surface alone is derived. As a test, a disk of TiO_2 of thickness λ/4 was built into a metal cavity, and was estimated to produce a reflection about twice as great as that from the associated metal surface alone.

621.372.2 **1016**

Open Wire Lines.—G. Goubau. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 197–200. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.372.2+621.372.8] : 537.226 **1017**

Propagation of Microwaves along a Single Conductor Embedded in Three Coaxial Dielectrics:—Part 2.—S. K. Chatterjee & R. Chatterjee. (*J. Indian Inst. Sci.*, Section B, April 1957, Vol. 39, No. 2, pp. 71–82.) "The characteristic equation for the EH wave has been derived. It is shown, as a special case, that the asymmetric wave EH_1 cannot be propagated along a solid conductor embedded in free space due to very high attenuation. Field components in terms of the axial power flow have been derived." Part 1: 16 of 1957.

621.372.413+621.372.8] : 518.5 **1018**

Special Slide Rule for Calculating the Internal Wavelength of E.M. Waves in Waveguides and Cavity Resonators.—W. Otto. (*NachrTech.*, July 1957, Vol. 7, No. 7, pp. 294–296.)

621.372.8 **1019**

Contribution to the General Theory of Irregular Waveguides.—B. Z. Katsenelenbaum. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Sept. 1957, Vol. 116, No. 2, pp. 203–206.) Waveguides of nonuniform circular cross-section are examined and their main characteristics obtained by an approximation method. See also 676 of 1957.

621.372.8 **1020**

Miniaturization of Microwave Assemblies.—L. Lewin. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 261–262. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.)

621.372.821 : 621.372.86 **1021**

The Excitation of Surface Waveguides and Radiating Slots by Strip-Circuit Transmission Lines.—A. D. Frost, C. R. McGeoch & C. R. Mingins. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 218–222. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.372.825 **1022**

A Note on the Fourier Series Representation of the Dispersion Curves for Circular Iris-Loaded Waveguides.—P. N. Robson. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 69–72.) A rapid way of determining the dispersion curve to the accuracy necessary when designing slow-wave structures for electron accelerators. See also 19 of 1956 (Grosjean).

621.372.83 : 621.372.029.6 **1023**

Recent Advances in Finline Circuits.—S. D. Robertson. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 263–267. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.) See also 2555 of 1955.

621.372.832.6 **1024**

Recent Advances in Waveguide Hybrid Junctions.—P. A. Loth. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 268–271. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.)

621.372.832.8 **1025**

A Broad-Band Microwave Circulator.—E. A. Ohm. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 210–217. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.372.832.8 **1026**

The Turnstile Circulator.—P. J. Allen. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 223-227. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 113.)

621.372.85 : 621.318.134 **1027**

Propagation in Ferrite-Filled Transversely Magnetized Waveguide.—P. H. Vartanian & E. T. Jaynes. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp. 140–143. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1898.)

621.372.852.363 : 621.318.134 **1028**

Improved Rectangular-Waveguide Resonance Isolators.—M. T. Weiss. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 240–243. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.)

621.372.853 : 537.562 **1029**

Properties of Ion-Filled Waveguides.—L. D. Smullin & P. Chorney. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 360–361.) A mathematical analysis suggests that two pass bands exist at frequencies below the usual pass band for the empty waveguide.

621.396.67 : 537.226 **1030**

Some Investigations on Dielectric Aerials.—R. Chatterjee & S. K. Chatterjee. (*J. Instn Telecommun. Engrs, India*, Sept. 1957, Vol. 3, No. 4, pp. 280–284.) A comparative study of the expressions obtained for the radiation pattern of a circular dielectric rod aerial, excited in the HE_{11} mode (see 3637 of 1956 and *J. Indian Inst. Sci.*, Section B, July 1957, Vol. 39, No. 3, pp. 134–140) using (*a*) Schelkunoff's equivalence principle, (*b*) the application of Huyghen's principle over the whole rod, and (*c*) the theory of Halliday and Kiely (3350 of 1948).

621.396.674.3 : 621.396.11 **1031**

Radiation resulting from an Impulsive Current in a Vertical Antenna Placed on a Dielectric Ground.—C. L. Pekeris & Z. Alterman. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1317–1323.) The electromagnetic field produced by passing a current with the form of a delta function $\delta(t)$ through a vertical dipole on a dielectric ground is calculated with the aid of an electronic computer, and the solutions discussed. This method of treatment has certain advantages over the assumption of a periodic current and can also be generalized to deal with an arbitrary current variation.

621.396.676.012.12 **1032**

Currents Excited on a Conducting Surface of Large Radius of Curvature.—J. R. Wait. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp, 143–145. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1898.) See also 2296 of 1956.

621.396.677 + 534.23 **1033**

Some Aspects of the Design of Strip Arrays.—Tucker. (See 994.)

621.396.677 : 621.372.826 **1034**

An Investigation of Periodic Rod Structures for Yagi Aerials.—J. O. Spector. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 38–44.) Experiments suggest that the radiation mechanism can be explained in terms of a surface wave propagating along the aerial together with a radiating aperture at the end. The side-lobe structure is explained by interference with direct radiation from the driven element caused by inefficient launching of the surface wave.

621.396.677.029.62 : 523.164 **1035**

A Radio Telescope.—J. Firor. (*QST*, Sept. 1957, Vol. 41, No. 9, pp. 32–36.) A description of an interferometer system including constructional details of an aerial system suitable for tracking an earth satellite on a frequency of 108 Mc/s.

621.396.677.029.62 : 523.164 : 629.19 **1036**

Mark II Minitrack Base-Line Components.—R. L. Easton. (*QST*, Sept. 1957, Vol. 41, No. 9, pp. 37–41.) Full constructional details of an aerial system suitable for earth satellite tracking on 108 Mc/s. See also *ibid.*, July 1956, Vol. 40, No. 7, pp. 38–41, 134.

621.396.677.31 : [523.164 + 523.5 **1037**

An Antenna Array for Studies in Meteor and Radio Astronomy at 13 Metres.—P. B. Gallagher. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 89–92.) Short description of a broadside array at Stanford, California, designed for radar studies of very small meteors. Preliminary results indicate that the system should be sensitive to meteors down to the 15th visual magnitude. The theoretical half-power azimuthal beam width is 1·2°. Other possible applications of the system are noted.

621.396.677.83 : 523.164 **1038**

Radio-Telescope Antennas of Large Aperture.—J. D. Kraus. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 92–97.) A discussion of the detection and resolving powers of radio telescopes illustrates the need for larger apertures, and a design of a low-cost aerial system which provides a large effective aperture at 300 Mc/s is described. It consists of a combination of long parabolic and flat sheet reflectors built on the ground.

621.396.677.833.2.095 **1039**

The Radiation Diagram of the Paraboloid under Various Conditions of Illumination.—G. Barzilai, C. Montebello & F. Serracchioli. (*Note Recensioni Notiz.*, July/Aug. 1957, Vol. 6, No. 4, pp. 525–538.) Radiation diagrams obtained at 9 375 Mc/s with different types of radiator are reproduced and discussed.

621.396.677.833.2.095 **1040**

Microwave Antenna Characteristics in the Presence of an Intervening Ridge.—R. Vikramsingh, M. N. Rao & S. Uda. (*J. Instn Telecommun. Engrs, India*, Sept. 1957, Vol. 3, No. 4, pp. 274–279.) A description of parabolic-aerial characteristics at 1 860 Mc/s and 1 940 Mc/s as measured over a 14-km path. The polar diagrams, and the effects of polarization, defocusing and varying angle of elevation are recorded.

AUTOMATIC COMPUTERS

681.142 **1041**

Logical Design of a Computer for Business Use.—R. J. Froggat. (*J. Brit. Instn Radio Engrs*, Dec. 1957, Vol. 17, No. 12, pp. 681–696.) Paper presented at the Convention on Electronics in Automation, Cambridge, England, June 1957, describing an E.M.I. serial machine operating at a clock rate of 115 kc/s with magnetic-drum storage.

681.142 **1042**

Torsional Waves in Wires: Disk-Loading increases Computer Memory.—(*Engineering, Lond.*, 21st June 1957, Vol. 183, No. 4763, p. 787.) Note on the characteristics of a disk-loaded delay line machined from a solid brass rod.

681.142 **1043**

Digital/Analogue Converter Provides Storage.—H. N. Putschi, J. A. Raper & J. J. Suran. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 148–151.) A transistorized converter changes eight binary bits, received in parallel from a shift register, to 128 steps in amplitude of a 400-c/s sine wave. One binary bit is used to obtain phase information. Operation is performed within a 4-ms sampling period occurring at an average rate of 20 c/s.

681.142 : 621.039 **1044**

Analogue and Digital Computers in Nuclear Engineering.—(*Nucleonics*, May 1957, Vol. 15, No. 5, pp. 53–88.) A group of articles outlining the principles of operation and giving a survey of applications in the U.S.A.

681.142 : 621.314.7 **1045**

A.D.A.—a Transistor Decimal Digital Differential Analyser.—N. W. Allen. (*J. Instn Engrs Aust.*, Oct./Nov. 1957, Vol. 29, Nos. 10/11, pp. 255–262.) Circuit techniques and assembly and construction methods are discussed. The specification of the machine is summarized and some typical problems are considered.

681.142 : 621.318.57 : 537.227 **1046**

Signals from Switched Ferroelectric Memory Capacitors.—Pulvari & McDuffie. (See 1053.)

681.142 : 621.372.4/.5 **1047**

Electric Nonlinear Computation Performed by Linear Elements.—Abou-Hussein. (See 1061.)

681.142 : 621.375.4 **1048**

Application of Junction Transistors to Carrier-Frequency Computing Amplifiers.—W. A. Curtin. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 746–752.) A discussion of the properties required by a transistorized summing amplifier in a carrier-frequency analogue computer. Details are given of a practical 400-c/s amplifier using standard commercial transistors.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 **1049**

Moulding moves in on Printed Circuits.—N. L. Greenman. (*Product Engng*, 30th Sept. 1957, Vol. 28, No. 12, pp. 90–93.) The moulding process described permits the use of heavy-gauge conductor material in circuits of high current-carrying capacity.

621.3.078 : 621.372.852.3 **1050**

An Automatic Gain Control System for Microwaves.—J. P. Vinding. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 244–245. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.)

621.314.2 : 621.372.51 **1051**

Wide-Band Balun Transformer.—A. I. Talkin & J. V. Cuneo. (*Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 808–815.) Description of a balun transformer developed for use with a c.r.o. deflection amplifier. Bandwidth extends from below 50 kc/s to above 500 Mc/s. Input impedance data and pulse response characteristics are shown. One balun coil can be used alone as a passive wide-band pulse inverter.

621.314.2.029.55/.62 **1052**

Design of Wide-Band R.F. Transformers utilizing a Synthesized Equivalent Network.—H. H. Kajihara. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 802–805.)

621.318.57 : 537.227 : 681.142 **1053**

Signals from Switched Ferroelectric Memory Capacitors.—C. F. Pulvari & G. E. McDuffie, Jr. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 681–685.) The equivalent circuit for a bistable storage capacitor is a current generator with a resistance in parallel. The properties of the output signals from switched and unswitched stores are discussed using this model.

621.318.57 : 621.314.63 **1054**

High-Speed Microwave Switching of Semiconductors.—R. V. Garver, E. G. Spencer & R. C. LeCraw. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1336–1338.) A microwave switch using an *n*-type Ge diode has been shown to give pulse rise and decay times as low as 3 mμs.

621.318.57 : 621.314.7 **1055**

Transistor NOR Circuit Design.—W. D. Rowe & G. H. Royer. (*Commun. & Electronics*, July 1957, No. 31, pp. 263–267.) A NOR circuit gives a signal output only when no input signal is present. Examination of transistor characteristics gives a stable design for an arbitrary number of inputs and outputs on a single Type-2N109 transistor. Crosstalk factor and speed of operation are considered.

621.318.57 : 621.314.7 **1056**

A New Family of Transistor Switching Circuits.—M. Rubinoff. (*Commun. & Electronics*, July 1957, No. 31, pp. 286–289.) Two pairs of output levels are obtained from the 'dual range' circuits described which incorporate both *n-p-n* and *p-n-p* transistors. The possibility of lower stand-by power dissipation and greater switching speed than the DCTL circuits (see 2855 of 1955) is offered.

621.318.57 : 621.314.7 **1057**

Transistor 2-Terminal Switches.—A. Har'el. (*Commun. & Electronics*, July 1957, No. 31, pp. 328–338.) Discussion showing how the blocking-voltage concept may be applied to obtain negative-resistance characteristics using a junction transistor, a point-contact transistor, a modified double-base diode, or a modified tetrode. A few applications mainly in telephone systems are indicated.

621.372 : 621.396.822 **1058**

A Note on Noise Temperature.—P. D. Strum. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp. 145–151. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1898.) Expressions are given for the effective noise temperature of a lossy passive network with internal noise sources. Their application in different receiving systems is discussed.

621.372.029.6 : 621.372.83 **1059**

Recent Advances in Finline Circuits.—Robertson. (See 1023.)

621.372.2 : 512.831 **1060**

Equivalent Admittance, Impedance and Scattering Matrices.—G. C. Corazza & F. Serracchioli. (*Note Recensioni Notiz.*, July/Aug. 1957, Vol. 6, No. 4, pp. 502–510.) See also 694 of March.

621.372.4/.5 : 681.142 **1061**

Electric Nonlinear Computation Performed by Linear Elements.—M. S. M. Abou-Hussein. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 378–380.) Simple linear circuits are shown for performing multiplication, division and inversion, and forming powers, positive and negative roots, logarithms, antilogarithms and hyperbolic functions.

621.372.5 : 621.3.015.3 **1062**

Interpretation of Network Theorems in Terms of Laplace Transforms.—V. M. Narbutt. (*J. Instn Telecommun. Engrs, India*, Sept. 1957, Vol. 3, No. 4, pp. 304–309.)

621.372.54 : 621.396.62 **1063**

Greater Selectivity with the C.W. Clipper Filter.—L. I. Albert. (*QST*, Sept. 1957. Vol. 41, No. 9, pp. 24–26.) A two-stage amplifier with variable bandwidth.

621.372.54(083.57) **1064**

Charts Simplify Passive *LC* Filter Design.—D. R. J. White. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 160–163.) "Universal design data permit design of Butterworth and Tchebycheff filters of prescribed steady-state insertion-loss characteristics. Design of band-pass prototypes of lumped-element configuration is given."

621.372.54.029.6 **1065**

A Travelling-Wave Directional Filter.—F. S. Coale. (*Trans. Inst. Radio Engrs*, Oct. 1956, Vol. MTT-4, No. 4, pp. 256–260. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1957, Vol. 45, No. 1, p. 114.)

621.372.54.049 **1066**

Design and Manufacture of Practical Filter Circuits.—S. Boyle. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 154–157.) Discusses the effects of distributed capacitance, component proximity, encapsulating material and impedance mismatch.

621.372.56 : 621.385.029.6 **1067**

Travelling - Wave - Tube Limiters.—Fank & Wade. (See 1288.)

621,372.6 : 621.3.016.35 **1068**

A Phasor Method of Nonlinear Network Analysis.—J. P. Neal. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 630–636. Discussion, p. 636.) The network is represented by the sum of a finite number of simple sinusoidal functions with time, chosen from a knowledge of the network, experiment and preliminary analysis. An appendix illustrates stability tests and an application of the method to a typical problem.

621.373 : 517.942.932 **1069**

Fine Structure of Response Curves of Frequency-Entrained Oscillations.—C. A. Ludeke & J. D. Blades. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1326–1328.) Jump and hysteresis phenomena suggested by Cartwright (see 2740 of 1948) together with a dependence of entrainment range on forcing amplitude have been observed in an electromechanical analogue of a forced self-excited system.

621.373.029.4 : 621.396.822 **1070**

A Low - Frequency Random - Signal Generator.—J. C. West & G. T. Roberts. (*J. sci. Instrum.*, Nov. 1957, Vol. 34, No. 11, pp. 447–450.) The 5-kc/s and 8-kc/s noise components, each in an overall bandwidth of 250 c/s, are selected from the noise output of an argon-filled thyratron operating in a magnetic field. After peak rectification the resulting noise signals are added so that their d.c. components cancel. The amplitude distribution of the resultant output then approximates very closely to a Gaussian distribution. The output noise spectrum is flat from zero frequency up to 100 c/s. A full circuit diagram is given.

621.373.029.45 : 621.372.5 **1071**

Measurements of the Quality of Very-Low-Frequency *RC* Oscillators.—A. Zanini. (*Note Recensioni Notiz.*, July/Aug. 1957, Vol. 6, No. 4, pp. 459–477.) A 'figure of merit' is calculated for a number of different *RC* networks, and the experimental determination of this coefficient is described.

621.373.421.14.029.63 **1072**

Double Excitation in Decimetre-Wave Oscillators with Coaxial Lines.—W. Rohde. (*NachrTech.*, July 1957, Vol. 7, No. 7, p. 311.) The simultaneous oscillation at a frequency corresponding to a setting of about $\lambda/4$ in a disk-seal triode oscillator tuned to $3\lambda/4$ is investigated and the calculations are verified by experiment.

621.373.431.1 : 621.314.7 : 621.397.621 **1073**

Monovibrator has Fast Recovery Time.—A. I. Aronson & C. F. Chong. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 158–159.) "Use of complementary transistors decreases recovery time of monostable multivibrator. This prevents erratic operation when circuit is used in a television sync generator. Since both transistors are off during timing cycle, circuit is relatively insensitive to transistor variations and operates reliably from −50 to +70°C for input frequencies from 250 c/s to 1 Mc/s."

621.373.44 **1074**

A Simple Rectangular-Pulse Generator.—H. L. Armstrong. (*Canad. J. Phys.*, Oct. 1957, Vol. 35, No. 10, pp. 1250–1252.) The cathode-coupled multivibrator circuit discussed produces pulses of up to 100-V amplitude with rise times from 0·1 to 1 μs.

621.373.44 **1075**

Method of Obtaining Integral Mark/Space Ratios of High Accuracy with a Continuously Variable Control Frequency.—R. Rickert. (*NachrTech.*, July 1957, Vol. 7, No. 7, pp. 303–304.) A pulse generator circuit is described in which the mark/space ratio can be adjusted to within 0·1% over the range 1 : 1 to 1 : 9 for control frequencies from 10 c/s to 1 kc/s.

621.373.5 : 621.3.018.41(083.74) **1076**

Increasing Quartz Oscillator Stability.—(*Engineering, Lond.*, 18th Oct. 1957, Vol. 184, No. 4780, p. 496.) Brief description of a 5-Mc/s quartz servo oscillator forming part of frequency-standard equipment Type RD101. A frequency error of less than 1 in 10^{10} occurs for a 5% change in supply voltage or 5°C change in temperature.

621.374.32 **1077**

Fast Gray Wedge Analyser for High Input Rates.—J. T. Flynn & F. A. Johnson. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 867–874.) A full description is given of the circuits and operation of a pulse height analyser designed for positive input pulses at rates up to 10^7/s with less than 10% distortion of the spectrum. The resolution is approximately 10 mμs.

621.374.5 : 621.396.962.3 **1078**

Pulse Compression: Part 2.—R. Krönert. (*NachrTech.*, July 1957, Vol. 7, No. 7, pp. 305–308.) Description of an experimental realization of the method discussed in 72 of January.

621.375.1.011.6 **1079**

The Time Behaviour of Logarithmic Amplifier Input Circuits.—T. P. Flanagan. (*J. sci. Instrum.*, Nov. 1957, Vol. 34, No. 11, pp. 450–452.) "The non-linear nature of the logarithmic amplifier input circuit causes the time constant to vary with current, and three definitions of effective time constant are examined, based on an analysis of the time behaviour of the input circuit to a step change of input current."

621.375.122 **1080**

New Types of D.C. Amplifier: Part 1—The Cascade-Balance System.—D. J. R. Martin. (*Electronic Radio Engr*, Jan. 1958, Vol. 35, No. 1, pp. 2–7.) The Owen-Prinz method of zero-drift correction [see e.g. 1602 of 1948 (Prinz)] is applied to the first of two identical amplifier stages; the residual drift is then balanced against that of the second stage. The overall drift is reduced 100 times as compared with the parallel balance system. A practical design is fully discussed.

621.375.122 **1081**

New Types of D.C. Amplifier: Part 2—The Reflex-Monitor System.—D. J. R. Martin. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 56–62.) A correcting amplifier or monitor, identical to the input stage of the main amplifier, corrects alternately its own drift and that of the main amplifier. A differential input is used which rejects in-phase drift components, and residual drifts are balanced between the two amplifiers. Full details are given of a practical circuit which has certain advantages over the simpler cascade-balance system described in Part 1 (1080 above).

621.375.127 **1082**

Push-Pull Amplifier Design.—R. G. Christian. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 72–73.) "A method of designing class-AB push-pull amplifiers which eliminates the need to plot composite characteristics is discussed, and some examples are given which compare favourably with practical results."

621.375.132 **1083**

Total Differential Feedback.—J. C. H. Davis. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 40–44.) A device for squaring the effective possible feedback in a feedback amplifier.

621.375.3 **1084**

The Simple Reactor Circuit; its Operation and Mode Transition.—J. F. Ringelman & A. L. Fenaroli. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 660–668. Discussion, pp. 668–669.)

621.375.3 **1085**

The Series Magnetic Amplifier.—R. C. Barker. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 819–830. Discussion, pp. 830–831.) A theoretical analysis is given of the steady-state operation of the amplifier for saturation states of the cores. The control-circuit-impedance and control-voltage relations determining the mode of operation are established, and the gain characteristic and time constant are considered.

621.375.3 **1086**

Analysis of the Full-Wave Magnetic-Amplifier Circuits considering the Change of the Width of the Dynamic Hysteresis Loop.—T. Kikuchi. (*Commun. & Electronics*, July 1957, No. 31, pp. 241–249.)

621.375.3 **1087**

The Operation of the Current-Type Self-Balancing Magnetic Amplifier.—A. D. Krall. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 380–384.) A theoretical analysis giving the voltage gain and time constant in terms of the circuit parameters. Within the range of ideal components the input impedance is zero.

621.375.4 **1088**

High-Frequency Amplification using Junction Transistors.—L. E. Jansson. (*Mullard tech. Commun.*, Oct. 1957, Vol. 3, No. 26, pp. 174–187.) The maximum possible stage gain of a unilateralized grounded-emitter narrow-band amplifier stage is calculated and the design of inter-stage coupling transformers is described. A typical 470-kc/s i.f. amplifier stage is designed and its stability factor calculated.

621.375.4 **1089**

An Analysis of Transient Response of Junction-Transistor Amplifiers.—J. C. Bhattacharyya. (*J. Instn Telecommun. Engrs, India*, Sept. 1957, Vol. 3, No. 4, pp. 297–303.) An exact solution of the one-dimensional diffusion equation is obtained by the method of Laplace's transform. The short-circuited output collector current is calculated for a step-input forcing function. Experimental results agree with the theory, and for ordinary conditions the physical process underlying transistor action must be diffusion of minority carriers across the base region.

621.375.4+621.376.233] : 539.169 **1090**

How Transistors Operate under Atomic Radiation.—R. L. Riddle. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 125–127.) Results of tests on an amplifier/detector system show that degrading effects of irradiation can be controlled to some degree by negative feedback. Measurements were also made on a single separate transistor and a coaxial cable.

621.375.4 : 621.396.822 **1091**

Internal Noise of Transistor Amplifiers.—J. J. Brophy & A. R. Reinberg. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 965–966.) The internal noise levels of two commercial transistor amplifiers were measured for various input impedances and the results are shown graphically.

621.375.4.029.3 **1092**

Transistors in Speech Equipment.—H. J. Albrecht. (*QST*, Sept. 1957, Vol. 41, No. 9, pp. 19–22.) The use of transistors in a.f. circuits is described and constructional details are given of a speech amplifier.

621.375.9 : 538.221 **1093**
: 538.569.4.029.63/.64

Theory of the Ferromagnetic Microwave Amplifier.—H. Suhl. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1225–1236.) "All three possible types of operation using respectively, two electromagnetic cavity modes, two sample modes, and one sample and one cavity mode, are discussed. One especially simple case, that of a sphere in the first type of operation, is treated separately. Thereafter all three cases are discussed in terms of scalar and vector potentials. An appendix deals with the gain-bandwidth problem and gives an expression for the equivalent 'negative Q' of the sample."

621.375.9 : 538.569.4.029.6 **1094**

Twin Cavity for NH_3 Masers.—J. Bonanomi, J. Herrmann, J. De Prins & P. Kartaschoff. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 879–881.) "A system of two coupled cavities is described replacing the single cavity of an NH_3 maser. Using this system the curve of the oscillation frequency against cavity temperature presents a plateau, thus reducing considerably the 'pulling' effects of the cavity."

621.375.9 : 538.569.4.029.6 **1095**
: 621.396.822

Maser Noise Considerations.—J. Weber. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 537–541.) A discussion of the effect of the saturation r.f. field on the noise figure of a three-level maser. For the conditions considered, the effect is small. The spontaneous-emission equivalent temperature for a free-electron vacuum-tube amplifier is one-half that of a maser. In principle, more general quantum-mechanical amplifiers can be constructed which will not have spontaneous-emission noise.

621.376.32 : 621.318.134 **1096**

Compact Microwave Single-Sideband Modulator using Ferrites.—J. C. Cacheris & H. A. Dropkin. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp. 152–155. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, pp. 1898–1899.) See also 3183 of 1954 (Cacheris).

621.376.332 **1097**

Foster-Seeley Discriminator.—C. G. Mayo & J. W. Head. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 44–51.) A detailed analysis is given of the circuit and the use of the design formulae is illustrated. With the circuit arrangement described a sensitivity of 1·4 V (a.f. signal) per mA (i.f.) can be obtained with less than 0·1% second-harmonic distortion at 100% modulation.

GENERAL PHYSICS

531.19-2 **1098**

Statistical Properties of an Isotropic Random Surface.—M. S. Longuet-Higgins. (*Phil. Trans. A*, 17th Oct. 1957, Vol. 250, No. 975, pp. 157–174.)

535.37 **1099**

Possibility of Luminescent Quantum Yields Greater than Unity.—D. L. Dexter. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 630–633.) "It is shown that an excited sensitizer can transfer its energy simultaneously to two activators, under suitable conditions, leading to two emitted photons per incident higher-energy photon. The probability of this transfer process is computed, and the process is shown to be experimentally feasible."

537 + 538] : 621.38 **1100**

Physics and the New Electronics.—R. A. Smith. (*J. sci. Instrum.*, Oct. 1957, Vol. 34, No. 10, pp. 377–382.) Vacuum electronics and the new crystalline-solid electronics are compared. A brief account of recent work on semiconductors and maser amplifiers is given and the effect of these new developments on the training of electronic engineers is discussed.

537.122 **1101**

Some Remarks about Electron Correlation.—O. Krisement. (*Phil. Mag.*, Feb. 1957, Vol. 2, No. 14, pp. 245–258.) A comparison is made between theories representing different approaches to the inclusion of correlation in the wave-mechanics treatment of a free electron gas.

537.122 **1102**

Correlation Energy of an Electron Gas at High Density: Plasma Oscillations.—K. Sawada, K. A. Brueckner, N. Fukuda & R. Brout. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 507–514.) An analysis of the contribution made by zero-point oscillations to the correlation energy of an electron gas at high density. This contribution is expressed in terms of that from the scattering states by using the analytic properties of the scattering amplitudes. The results are compared with those of Bohm & Pines (1375 of 1954).

537.122 **1103**

Correlation Energy of a High-Density Gas: Plasma Coordinates.—R. Brout. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 515–517.) "The model Hamiltonian of Sawada [*ibid.*, 15th April 1957, Vol. 106, No. 2, pp. 372–383] which describes electron correlation at high density is examined. It is shown that the set of scattering modes for momentum transfers below a certain $q_{\max}$ is not complete. It is completed by the plasma mode. $(q_{\max})^{-1}$ is the natural Debye length of the theory."

537.122 **1104**

Characteristic Energy Loss of Electrons passing through Metal Foils: Momentum-Exciton Model of Plasma Oscillations.—R. A. Ferrell & J. J. Quinn. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 570–575.) Previous work is extended to obtain a unified theory of collective and individual electron effects in the excitation of a degenerate electron gas by fast incident electrons. The theory is equivalent to the screening out by the conduction electrons of the external field set up by the incident electrons.

537.311.1 **1105**

Quantum Theory of Electrical Transport Phenomena.—W. Kohn & J. M. Luttinger. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 590–611.) Using a simple model a technique is developed which gives the entire density matrix of the system of charge carriers in the steady state. The model consists of noninteracting free (or Bloch) electrons scattered by 'random' rigid impurity centres. The density matrix is developed in ascending powers of the strength of the scattering potential. The Boltzmann transport equation represents an approximation valid in the limiting cases of very weak or very dilute scatterers. Higher-order corrections are given.

537.311.31 : 536.63 **1106**

The Specific Heat of the Electrons of Metals.—F. Kaschluhn. (*Ann. Phys., Lpz.*, 20th Dec. 1956, Vol. 19, Nos. 3–5, pp. 94–101.) A new approximation method is used to calculate the specific heat for low temperatures taking account of electron interaction. Results are compared with those obtained by other methods.

537.525 **1107**

Exact Nonlinear Plasma Oscillations.—I. B. Bernstein, J. M. Greene & M. D. Kruskal. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 546–550.) A solution is given for the problem of a one-dimensional stationary nonlinear electrostatic wave in a plasma free from inter-particle collisions. In the limiting case of small-amplitude waves, the linearized theory can still be applied by using singular first-order distribution functions.

537.525 **1108**

Contribution to the Diffusion Oscillations in Gas-Discharge Plasma.—O. V. Prudkovskaya. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Dec. 1957, Vol. 117, No. 4, pp. 601–604.)

537.525.8 : 538.569 **1109**

Quenching of the Negative Glow by Microwaves in Cold-Cathode Gaseous Discharges.—J. M. Anderson. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 898–899.) A discussion of results obtained in He at a pressure of 10·2 mm Hg using a 40-μs pulse at 9 375 Mc/s. Enhanced emission was observed in Ne, A, and Xe discharges at 1–10 mm Hg.

537.533.8 **1110**

The Theory of the Surface Effect of Secondary Emission.—W. Brauer & W. Klose. (*Ann. Phys., Lpz.*, 20th Dec. 1956, Vol. 19, Nos. 3–5, pp. 116–132.) For primary energies of about 100 eV about 30% of the secondary-electron yield in light metals can be ascribed to surface effects. For higher energies and heavier metals surface effects can be neglected.

537.56 **1111**

Effective Electron-Electron and Electron-Ion Cross-Sections in Plasmas.—M. Bayet. (*J. Phys. Radium*, June 1957, Vol. 18, No. 6, pp. 380–386.) Cross-sections for elastic scattering are infinite for strictly Coulomb interaction, but finite values are found by using Debye's theory of electrolytes.

538 : 061.3 **1112**

Report on the 4th Course of the International School of Physics of the Italian Physical Society, Varenna, 15th July–4th August 1956.—(*Nuovo Cim.*, 1957, Vol. 6, Supplement No. 3, pp. 805–1237.) Report of the proceedings of the course on magnetic properties held at the Villa Monastero, Varenna. The text is given of lectures and communications, including the following:

(*a*) Paramagnetism in Crystals.—M. H. L. Pryce (pp. 817–856, in English).

(*b*) Magnetic Properties of Metals.—J. H. Van Vleck (pp. 857–886, in English).

(*c*) Paramagnetic Relaxation.—C. J. Gorter (pp. 887–894, in English).

(*d*) Ferromagnatism.—C. Kittel (pp. 895–922, in English).

(*e*) Antiferromagnetism.—C. J. Gorter (pp. 923–941, in English).

(*f*) Metamagnetics or Critical-Field Antiferromagnetic Materials.—L. Néel (pp. 942–960, in French).

(*g*) Nuclear Magnetism and Nuclear Relaxation.—E. M. Purcell (pp. 961–992, in English).

(*h*) Line Breadths and the Theory of Magnetism.—J. H. Van Vleck (pp. 993–1014, in English).

(*i*) The Influence of Electrons on Nuclear Spin Resonance in Diamagnetic Materials: 'Chemical' Displacement and Indirect Interaction.—A. Abragam (pp. 1015–1062, in French).

(*j*) Stochastic Theory of Magnetic Resonance.—R. Kubo (pp. 1063–1080, in English).

(*k*) The Concept of Temperature in Magnetism.—J. H. Van Vleck (pp. 1081–1100, in English).

(*l*) Magnetism at Very Low Temperatures and Nuclear Orientation.—N. Kurti (pp. 1101–1139, in English).

(*m*) Cyclotron Resonance in Crystals.—C. Kittel (pp. 1140–1147, in English).

(*n*) Optical Methods for Radio-Frequency Resonance.—A. Kastler (pp. 1148–1167, in French).

(*o*) Magnetic Properties of Superconductors.—C. J. Gorter (pp. 1168–1176, in English).

(*p*) Hall Effect in Ferromagnetics.—J. Smit (pp. 1177–1182, in English).

(*q*) Critical Scattering of Neutrons from Ferromagnets.—W. Marshall (pp. 1183–1184, in English. Discussion, pp. 1184–1185).

(*r*) The Van Vleck Model of Ferromagnetism.—W. Marshall (pp. 1186–1187, in English).

(*s*) Influence of the Apparatus on Nuclear Magnetic Resonance.—H. Pfeifer (pp. 1188–1189, in English).

(*t*) Paramagnetic Resonance of Impurities in a Semiconductor.—A. Abragam & J. Combrisson (pp. 1197–1211, in French. Discussion, p. 1212). See also 165 of 1957.

(*u*) The Electric Analogue to Antiferromagnetism: Antiferroelectricity.—H. Gränicher (pp. 1220–1222, in English. Discussion, p. 1223).

(*v*) Magnetic Susceptibility of Electrons in Periodic Fields.—C. P. Enz (pp. 1224–1229, in English).

538.3 **1113**

Equivalence Theorems for Electromagnetic Fields.—G. C. Corazza. (*Ricerca sci.*, July 1957, Vol. 27, No. 7, pp. 2109–2119.) Theorems are considered with reference to electric and magnetic surface currents.

538.3 **1114**

Forces and Stresses in an Electromagnetic Field.—T. H. Lee. (*Commun. & Electronics*, July 1957, No. 31, pp. 267–271. Discussion, pp. 271–274.) The theories of Faraday, Larmor and Livens, and the 'energy method' are compared and applied to some illustrative examples. The three theories may lead to different results for nonrigid magnetic materials, and the Larmor and Livens theory may not be applicable to magnetostriction. A discussion by R. M. Lichtenstein giving a further 'thermodynamic' method is included.

538.311 **1115**

Coils for the Production of High-Intensity Pulsed Magnetic Fields.—S. Foner & H. H. Kolm. (*Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 799–807.) The design, performance and applications of a pulsed-field system for 750 000 G are described. It consists of a 2 000-μF capacitor bank charged to 3 kV which is discharged through a Be-Cu helix of $\frac{3}{16}$ in. internal diameter and $\frac{1}{2}$ in. long. Detailed design data are also given for a range of coils giving increased volume and field uniformity at lower field intensities and include a coil providing transverse access to the field.

538.311 **1116**

Production and Use of High Transient Magnetic Fields: Part 2.—H. P. Furth, M. A. Levine & R. W. Waniek. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 949–958.) The mechanical and thermal limitations of solid metal coils are discussed and illustrated by experiment. These limitations can be removed by force-free coil designs. Part 1: 3030 of 1956 (Furth & Waniek).

538.311 **1117**

Variable and Reversible Magnetic Fields Obtained from a Permanent Magnet.—J. A. Dalman & L. S. Goodman. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 961–962.) A permanent magnet forms the core of a cylinder whose surface is made of two armco iron shoes and two brass shoes. Rotation of the cylinder directs the flux either to a shunt or across the gap of the magnet, for deflection of an atomic beam.

538.566 : 537.122 **1118**

Oscillations of Electron Cloud in External Fields.—M. Sumi. (*J. phys. Soc. Japan*, Oct. 1957, Vol. 12, No. 10, pp. 1110–1117.) The density fluctuations are analysed in a pure electron gas not neutralized by positive ions and under the influence of an external magnetic field.

538.566 : 621.396.677.85 **1119**

A Semi-infinite Array of Parallel Metallic Plates of Finite Thickness for Microwave Systems.—R. I. Primich. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp. 156–166. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1899.)

538.569.4 : 538.221 **1120**

On the Line Width in Ferromagnetic Resonance.—S. Takeno. (*Progr. theor. Phys.*, Oct. 1957, Vol. 18, No. 4, pp. 448–449.) It is suggested that the line width is modified by a nonuniformity of the demagnetization field throughout the specimen. It is shown that the broadening is greater for a disk magnetized normal to its plane than for one magnetized in its plane, and also that the effect would be greater for ferrites than metals; both results are in agreement with experiment.

538.569.4.029.63/.64 : 538.221 **1121**
: 621.375.9

Theory of the Ferromagnetic Microwave Amplifier.—Suhl. (See 1093.)

538.569.4.029.65 : 535.34 **1122**

High-Temperature Molecular-Beam Microwave Spectrometer.—A. K. Garrison & W. Gordy. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 899–900.) A description of measurements at 3 mm λ on the $J = 12 \rightarrow 13$ rotational transition of KCl.

538.652 **1123**

Effective Magnetic Anisotropy and Magnetostriction of Monocrystals.—P. K. Baltzer. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 580–587.) A nine-constant expression is derived which describes empirically the spontaneous magnetostriction to the sixth order in the direction cosines of the magnetization. The relation between the effective anisotropy in domains and domain walls is also studied.

538.691 **1124**

Diffusion of Charged Particles in a Homogeneous Electromagnetic Field.—Kh. Ya. Khristov. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Sept. 1957, Vol. 116, No. 2, pp. 213–216.) An approximate solution is considered.

539.11 : 548.4 **1125**

The Elastic Impurity-Centre Model.—J. Teltow. (*Ann. Phys., Lpz.*, 20th Dec. 1956, Vol. 19, Nos. 3–5, pp. 169–174.) An approximation is derived for the purely elastic interaction between two extraneous atoms in a crystal lattice.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164 **1126**

Radio Astronomy.—(*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1.) The main part of this issue is devoted to a group of 50 papers dealing with practical and theoretical aspects of radio astronomy. Abstracts of some of the papers are given individually; titles of others are as follows:

(*a*) Introduction to Radio Astronomy.—F. T. Haddock (pp. 3–12).

(*b*) The Discovery and Identification by Karl Guthe Jansky of Electromagnetic

Radiation of Extraterrestrial Origin in the Radio Spectrum.—C. M. Jansky, Jr, (pp. 13–15).

(c) Early Radio Astronomy at Wheaton, Illinois.—G. Reber (pp. 15–23).

(d) The Telescope Program for the National Radio Astronomy Observatory at Green Bank, West Virginia.—R. M. Emberson & N. L. Ashton (pp. 23–35).

(e) Noise Levels at the National Radio Astronomy Observatory.—J. W. Findlay (pp. 35–38).

(f) Radio Astronomy at the Meudon Observatory.—E. J. Blum, J. F. Denisse & J. L. Steinberg (pp. 39–43).

(g) A High-Resolution Radio Telescope for Use at 3·5 m.—B. Y. Mills, A. G. Little, K. V. Sheridan & O. B. Slee (pp. 67–84).

(h) The Sydney 19·7-Mc/s Radio Telescope.—C. A. Shain (pp. 85–88).

(i) Measurements of Solar Radiation and Atmospheric Attenuation at 4·3 Millimetres Wavelength.—R. J. Coates (pp. 122–126).

(j) Scanning the Sun with a Highly Directional Array.—W. N. Christiansen & D. S. Mathewson (pp. 127–131).

(k) A Wide-Band Antenna System for Solar Noise Studies.—H. Jasik (pp. 135–142).

(l) The Radio Spectrum of Solar Activity.—A. Maxwell, G. Swarup & A. R. Thompson (pp. 142–148).

(m) A Swept-Frequency Interferometer for the Study of High-Intensity Solar Radiation at Metre Wavelengths.—J. P. Wild & K. V. Sheridan (pp. 160–171).

(n) The Cornell Radio Polarimeter.—M. H. Cohen (pp. 183–190).

(o) A Time-Sharing Polarimeter at 200 Mc/s.—S. Suzuki & A. Tsuchiya (pp. 190–194).

(p) A Polarimeter in the Microwave Region.—K. Akabane (pp. 194–197).

(q) Radio Sources and the Milky Way at 440 Mc/s.—N. G. Roman & B. S. Yaplee (pp. 199–204).

(r) Flux Measurements of Cassiopeia A and Cygnus A between 18·5 Mc/s and 107 Mc/s.—H. W. Wells (pp. 205–208).

(s) The Distribution of Cosmic Radiation Background Radiation.—H. C. Ko (pp. 208–215).

(t) A Galactic Model for Production of Cosmic Rays and Radio Noise.—L. Marshall (pp. 215–220).

(u) Hydrogen-Line Study of Stellar Associations and Clusters.—T. K. Menon (pp. 230–234).

(v) Extragalactic 21-cm-Line Studies.—D. S. Heeschen & N. H. Dieter (pp. 234–239).

(w) Measurements of Planetary Radiation at Centimetre Wavelengths.—C. H. Mayer, T. P. McCullough & R. M. Sloanaker (pp. 260–266).

(x) Planetary and Solar Radio Emission at 11 Metres Wavelength.—J. D. Kraus (pp. 266–274).

(y) Radio Emission from Comet 1956 h on 600 Mc/s.—R. Coutrez, J. Hunaerts & A. Koeckelenbergh (pp. 274–279).

(z) Lunar Thermal Radiation at 35 kMc/s.—J. E. Gibson (pp. 280–286).

(aa) Lunar Radio Echoes.—J. H. Trexler (pp. 286–292).

(bb) Radar Echoes from the Moon at a Wavelength of 10 cm.—B. S. Yaplee, R. H. Bruton, K. J. Craig & N. G. Roman (pp. 293–297).

(cc) A Phase Tracking Interferometer.—H. Penfield (pp. 321–325).

(dd) Radio Astronomy Measurements at V.H.F. and Microwaves.—J. Aarons, W. R. Barron & J. P. Castelli (pp. 325–333).

(ee) Cosmical Electrodynamics.—J. H. Piddington (pp. 349–355).

523.164 **1127**

Spectral Lines in Radio Astronomy.—A. H. Barrett. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 250–259.) A review of atomic and molecular resonance lines occurring in the radio spectrum. It is shown how radiation from the galaxy is modified by interstellar gas.

523.164 **1128**

Radio Astronomy Polarization Measurements.—M. H. Cohen. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 172–183.) A survey of possible techniques for measuring polarization, and for studying the Faraday effect, shows which components should be measured. Methods for deducing the electron density along the path are discussed and applied to data on radiation from the sun and the Crab Nebula.

523.164 **1129**

Absorption Techniques as a Tool for 21-cm Research.—A. E. Lilley & E. F. McClain. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 221–229.) Absorption lines in the spectra of radio stars are used to study the distribution of interstellar gas, minimum distances to radio stars and problems related to cosmology.

523.164 **1130**

Excitation of the Hydrogen 21-cm Line.—G. B. Field. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 240–250.) The influences of certain mechanisms on the spin temperature such as electron collision and interaction with light. are considered, and their importance to radio observations is discussed. The deuterium line at λ = 91·6 cm is also considered.

523.164 : 53.088 **1131**

Restoration in the Presence of Errors.—R. N. Bracewell. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 106–111.) A discussion of the relationship between observed aerial temperatures and the true distribution of a celestial source. Techniques for allowing for certain errors are described.

523.164 : 621.317.794 : 621.396.822 **1132**

A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy.—Drake & Ewen. (See 1229.)

523.164 : 621.317.794.089 : 621.396.822 **1133**

A Method of Calibrating Centimetric Radiometers using a Standard Noise Source.—Hey & Hughes. (See 1230.)

523.164 : 621.396.677.029.62 **1134**

A Radio Telescope.—Firor. (See 1035.)

523.164+523.5] : 621.396.677.31 **1135**

An Antenna Array for Studies in Meteor and Radio Astronomy at 13 Metres.—Gallagher. (See 1037.)

523.164 : 621.396.677.83 **1136**

Radio-Telescope Antennas of Large Aperture.—Kraus. (See 1038.)

523.164 : 629.19 : 621.396.677.029.62 **1137**

Mark II Minitrack Base-Line Components.—Easton. (See 1036.)

523.164.3 **1138**

The Large-Scale Structure of the Galaxy.—F. J. Kerr, J. V. Hindman & M. S. Carpenter. (*Nature, Lond.*, 5th Oct. 1957, Vol. 180, No. 4588, pp. 677–679.) Observations at 21 cm λ are being made at Sydney, Australia, using an aerial of 36 ft diameter and beam width 1·5°, and a receiver with bandwidth 40 kc/s. A composite diagram of the galaxy is shown based on these observations and similar observations at Leyden, Netherlands.

523.164.3 : 523.6 **1139**

Radio Observations of the Comet Arend-Roland.—G. R. Whitfield & J. Högbom. (*Nature, Lond.*, 21st Sept. 1957, Vol. 180, No. 4586, p. 602.) Observations made at Cambridge between 12th March and 13th May 1957 at 38 and 81·5 Mc/s using 5 different instruments gave negative results. Emission at 600 and 27·6 Mc/s has previously been reported elsewhere.

523.164.32 **1140**

Duration of Transients in Solar Radio Noise.—Ö. Elgaröy. (*Nature, Lond.*, 19th Oct. 1957, Vol. 180, No. 4590, pp. 808–809.) Histograms of observations made at 200 Mc/s at Oslo are given. Reber's theory (see 1637 of 1955) is confirmed for this frequency. See also 439 of February (de Groot). (Please note change of U.D.C. number.)

523.164.32 **1141**

Frequency Drift of Short-Time Transients in Solar Radio Noise.—Ö. Elgaröy. (*Nature, Lond.*, 26th Oct. 1957, Vol. 180, No. 4591, p. 862.) An analysis of records made at Oslo on a two-channel receiver is being made to give a detailed study of solar-noise storm bursts. Channels are centred on 199 and 200·5 Mc/s, bandwidth is 0·3 Mc/s, time constant 0·01 s and noise factor about 4. It is suggested that the transients in most cases have a frequency drift in either direction ranging from about 10 Mc/s^2 to higher values.

523.164.32 : 53.008 **1142**

Discussion of 10·7-cm Solar Radio Flux Measurements and an Estimation of the Accuracy of Observations.—W. J. Medd & A. E. Covington. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 112–118.) Revision of previously published data is recommended.

523.164.32 : 538.566.2(083.57) **1143**

Critical Frequency, Refractive Index, and Cone of Escape in the Solar Corona.—R. N. Bracewell & C. V. Stableford. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 198–199.) "Nomograms give refractive index governing radio propagation in the solar corona and the semivertical angle of the cone of escape."

523.164.32 : 523.75 **1144**

Studies at the McMath-Hulbert Observatory of Radio-Frequency Radiation at the Time of Solar Flares.—H. W. Dodson. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 149–159.) Records of radiation at 80, 200 and 2 800 Mc/s during several hundred flares are compared with associated data obtained from spectroheliograms.

523.164.32 : 621.317.75 **1145**

A Dynamic Spectrum Analyzer for Solar Studies.—J. Goodman & M. Lebenbaum. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 132–135.) The range 90–580 Mc/s is swept three times a second in three overlapping bands. Results appear as a frequency-intensity display on film.

523.164.42 : 535.417 **1146**

Radio Interferometry of Discrete Sources.—R. N. Bracewell. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 97–105.) "Salient features of the theory and practice of radio interferometry are presented with special attention to assumptions and to the specifically two-dimensional aspects of the subject. A theorem is proved according to which only certain discrete stations on a rectangular lattice need be occupied for full determination of a discrete source distribution. Procedures in interferometry are discussed in the light of this result and an optimum procedure is deduced. Current practice is considered over-conservative, e.g. independent data in the case of the sun are obtainable only at station spacings of about 100 wavelengths on the ground, a fact which has not hitherto been taken into account."

550.38 **1147**

Preliminary Report of Geomagnetic Observations at Prince Harald Coast, Antarctica.—T. Nagata, T. Oguti & K. Momose. (*Rep. Ionosphere Res. Japan*, June 1957, Vol. 11, No. 2, pp. 41–49.) The results of geomagnetic observations during the first Japanese Antarctic Research Expedition (1956/1957) are briefly summarized. The geomagnetic total intensity was found to be 10% less than the value in Vestine's world map. Correlations between geomagnetic and ionospheric phenomena are discussed.

550.389.2 : 629.19 **1148**

Apsidal Motion of an I.G.Y. Satellite Orbit.—L. Blitzer. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, p. 1362.) An error in previous work [756 of 1957 (Blitzer et al.) and *ibid.*, Feb. 1957, Vol. 28, No. 2, p. 279 (Blitzer & Wheelon)] is corrected, concerning the movement of the line of apsides relative to a node. (Please note change of U.D.C. number.)

551.510.53 : [551.557 + 551.524.7 **1149**

Winds and Temperatures between 20 km and 100 km—a Review.—R. J. Murgatroyd. (*Quart. J. R. met. Soc.*, Oct. 1957, Vol. 83, No. 358, pp. 417–458.) The results of measurements, by various methods, of temperature and wind velocity in the northern hemisphere are briefly outlined. The results are used to obtain consistent systems of zonal wind and temperature for the winter and summer seasons. A standard atmosphere is suggested for northern latitudes to a height of 60 km, extended to 100 km for latitudes 30° N to 60° N, for both seasons.

551.510.535 : 523.164 **1150**

The Use of Radio Stars to Study Irregular Refraction of Radio Waves in the Ionosphere.—H. G. Booker. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 298–314.) Information from various sources on the amplitude and phase scintillation of radio-star radiation is reviewed, and the experimental data compared with theory. There is general agreement on most, but not all, of the characteristics of the scintillation and its variation with time and zenith angle. The rate of scintillation increases under magnetically disturbed conditions. There is good correlation with spread-F ionospheric reflections and some with sporadic-E reflections. The scale of the irregularities causing the scintillation is of the order of a kilometre. The most puzzling problem is the cause of the night-time maximum in scintillation. 41 references.

551.510.535 : 523.164 **1151**

An Investigation of the Perturbations Imposed upon Radio Waves Penetrating the Ionosphere.—R. S. Lawrence. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 315–320.) A new method is described for measuring continuously the phase deviations introduced by the ionosphere into the radiation from discrete sources. Records at 53, 108 and 470 Mc/s are shown. In future work vertical incidence measurements will be made of the ionosphere at the point where the extraterrestrial waves penetrate.

551.510.535 : 523.164 **1152**

Some Measurements of High-Latitude Ionospheric Absorption using Extraterrestrial Radio Waves.—C. G. Little & H. Leinbach. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 334–348.) A theory of absorption and a technique of measurement are described. Observations at 30 Mc/s have shown that regions of anomalous absorption have dimensions in excess of 100 km and that marked differences can occur, in disturbed periods, between stations 800 km apart. Most of the absorption occurs below the E region and it correlates well with the geomagnetic *K* index.

551.510.535 "1957" : 621.396.11 **1153**

Ionosphere Review 1957.—T. W. Bennington. (*Wireless World*, Feb. 1958, Vol. 64, No. 2, pp. 77–78.) Sunspot activity has reached unprecedented high values; average F_2-layer critical frequencies have increased by a factor of just over two since 1954; monthly mean values for noon and midnight were considerably higher that at last sunspot maximum. M.u.f.'s as high as 50 Mc/s have been reported, and examples of long-distance reception at 41·5 and 45 Mc/s are noted. Though sunspot maximum may have occurred before the end of 1957 little change in propagation conditions is expected over various paths in 1958 as compared with those during similar months of 1957. 'Quasi-maximum' conditions may well prevail throughout 1959.

LOCATION AND AIDS TO NAVIGATION

621.396.932.2 **1154**

'S.A.R.A.H.': a U.H.F. (243-Mc/s) Pulse Coded Air/Sea Rescue System.—D. Kerr. (*J. Brit. Instn Radio Engrs*, Dec. 1957, Vol. 17, No. 12, pp. 669–680.) The portable beacon emits 160–240 pairs of 5–10-μs pulses per sec and the pulse spacing can be varied from 100–300 μs. The double pulse permits discrimination between several beacons, and facilitates synchronization at the receiving station. Speech may be transmitted by pulse frequency modulation. The search equipment can detect pulses at a range of 145 km at a height of 3 000 m, or 9 km at sea level.

621.396.933 **1155**

Some Radio Aids for High-Speed Aircraft.—J. S. McPetrie. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 11–13.) Discusses some of the newer navigational aids which might be expected to come into civil and military use during the next few years.

621.396.962.3 : 621.374.5 **1156**

Pulse Compression: Part 2.—Krönert. (See 1078.)

621.396.963.3 : 621.396.822 **1157**

Detection of Pulse Signals in Noise: Trace-to-Trace Correlation in Visual Displays.—M. I. Skolnik: D. G. Tucker. (*J. Brit. Instn Radio Engrs*, Dec. 1957, Vol. 17, No. 12, pp. 705–706.) Comment on 3149 of 1957 and author's reply.

621.396.969.3 **1158**

'Angels' on Centimetric Radars caused by Birds.—W. G. Harper. (*Nature, Lond.*, 26th Oct. 1957, Vol. 180, No. 4591, pp. 847–849.) A study of unexplained echoes recorded at East Hill 30 miles NW of London from 1952 onwards suggests that these are due to birds only and not to insects or meteorological effects. 10-cm-λ p.p.i. equipment was used with peak power 500 kW, pulse length 2 μs, and half-power beam width 1¼°.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.289 **1159**

Copper-Doped Germanium as a Model for High-Resistivity Photoconductors.—P. J. van Heerden. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 230–238.) In the electrical behaviour of high-resistivity copper-doped Ge the nature of the electrodes plays an essential role. The primary and secondary photocurrents can be sharply distinguished experimentally and form the basic concepts for an explanation of the photocurrent. The secondary photocurrent can be quantitatively expressed in terms of the primary photocurrent and the variation of the dark current with a voltage step-function. Observations are given on the space-charge-limited current and some light is thrown on the mechanism of electrical breakdown.

535.215 : 546.682.86 **1160**

New Infrared Detectors using Indium Antimonide.—D. G. Avery. (*J. sci. Instrum.*, Oct. 1957, Vol. 34, No. 10, pp. 394–395.) Description of the properties of photoconductive InSb detectors at room temperature, and of experimental work on *p-n* junctions cooled to 90°K used as photovoltaic cells.

535.215 : 546.817.231 **1161**

The Reaction of Oxygen with Lead Selenide.—R. H. Jones. (*Proc. phys. Soc.*, 1st Nov. 1957, Vol. 70, No. 455B, pp. 1025–1032.) Oxidation processes were investigated at various temperatures from ambient to 280°C, to gain fundamental information about the photoconductive sensitization of evaporated layers of lead selenide by oxygen. Results indicate that two processes occur during oxidation and this postulation is supported by earlier electrical measurements on similar layers. Theories of photoconductivity in such layers are discussed in relation to these experiments. See also 3887 of 1957.

535.37 : 53.082.5 **1162**

The Measurement of the Decay of Phosphors by means of a New Type of Phosphoroscope.—D. Hahn & H. J. Kösel. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 137–140.) Decay curves for periods ranging from 10^{-1} to 10^{-4} s can be obtained.

535.37 : 546.472.21 **1163**

Influence of the Intensity and Duration of Excitation on the Photoelectric Properties of Copper-Activated Zinc Sulphide.—B. Hagène & J. J. Le Fèvre. (*J. Phys. Radium*, June 1957, Vol. 18, No. 6, pp. 412–413.)

535.372 **1164**

The Effectiveness of Some Fluorescence Quenchers in Relation to the Position of the Fluorescence Spectrum.—V. V. Zelinskiĭ, V. P. Kolobkov & N. I. Kondaraki. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Nov. 1957, Vol. 117, No. 3, pp. 391–394.) The effect of aniline and tri-ethylamine iodide on several types of amino-N-methylphthalimide are examined and results tabulated.

535.376 : 546.472.21 **1165**

The Electroluminescence of ZnS Phosphors as an Equilibrium Process.—W. Lehmann. (*Optik, Stuttgart*, July/Aug. 1957, Vol. 14, Nos. 7/8, pp. 319–327.) The time average of electroluminescence intensity is characterized by dynamic equilibrium between monomolecular excitation and a recombination mechanism. The dependence of electroluminescence on field strength, temperature, activator and electron-trap concentrations is considered theoretically and a comparison is made with experimental results.

537.226/.227 : 546.431.824-31 **1166**
: 621.317.335.3.029.64

The Dielectric Constant of Barium Titanate at 10 Gc/s.—H. J. Schmitt. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 107–111.) The results of measurements at 9·4 kMc/s are discussed. Values of dielectric constant obtained range from 173 to 560 according to the grade of material. Comparison with 2-kc/s values indicates a relaxation frequency of about 10^{10} c/s which appears to be increased when a strong steady field is applied.

537.227 : 546.431.824-31 **1167**

Domain Conversion of Multidomain Barium-Titanate Single Crystals.—P. H. Fang, S. Marzullo & W. S. Brower. (*Phys. Rev.*, 15h Oct. 1957, Vol. 108, No. 2, pp. 242–243.) "It is found that complete domain conversion of single crystal $BaTiO_3$ can be achieved by passing the crystal through the orthorhombic-tetragonal transition under an applied d.c. field. By using this process either complete *a* or *c* domain crystals can be prepared."

537.311.33 **1168**

On the Electron-Lattice Interaction in Nonpolar Semiconductors.—S. Koshino. (*Progr. theor. Phys.*, July 1957, Vol. 18, No. 1, pp. 23–32.) The acoustical-mode scattering mobility is derived for the two-phonon process which is shown to be predominant at higher temperatures. The transition from the one-phonon to the two-phonon region can account for the dependence of mobility on temperature for both electrons and holes in Ge and Si.

537.311.33 **1169**

Studies on Group III-V Intermetallic Compounds.—C. Kolm, S. A. Kulin & B. L. Averbach. (*Phys. Rev.*, 15th Nov. 1957, Vol. 108, No. 4, pp. 965–971.) The effects of additions of Si, Ge, Sn and Pb to GaAs and InSb are described. The forbidden energy gap (determined by infrared absorption measurement), varies inversely with the lattice spacing for the GaAs alloys, except for the GaAs-Ge alloy where the reverse effect occurs. From infrared transmission and electrical resistivity data it is concluded that the Group IV elements substitute for nearest-neighbour pairs in GaAs, but with Si and Ge in InSb it appears that the substitution of only one atom in the unit polyhedron occurs. Similar measurements on InSb-GaSb alloys are also described.

537.311.33 **1170**

Calculation of the Electrical Conductivity of a Semiconductor with High Impurity Concentration.—R. Ziegenlaub. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Nov. 1957, Vol. 117, No. 3, pp. 395–398.) A brief mathematical analysis based on a new method for the evaluation of conductivity in semiconductors without the use of kinetic equations. The scattering of current carriers by ionized impurity centres is investigated.

537.311.33 **1171**

Electron and Hole Mobility in a Liquid Semiconductor.—I. Z. Fisher. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Nov. 1957, Vol. 117, No. 3, pp. 399–402.) A mathematical analysis of the diffusion of electrons and holes in a liquid is presented and an expression for electron or hole mobility is derived. The diffusion coefficient is estimated to be approximately 1 cm²/s.

537.311.33 : 535.214 **1172**

Light-Induced Plasticity in Semiconductors.—G. C. Kuczynski & R. F. Hochman. (*Phys. Rev.*, 15th Nov. 1957, Vol. 108, No. 4, pp. 946–948.) "It has been found that light of wavelength between 2·0 and 4·0 μ, decreases the hardness of the surface layer of *n*-type germanium by 10 to 60%. The softened layer extends to a depth of one to two microns. A similar but less intense effect was observed in *p*-type germanium and *n*-type InSb and InAs. On the other hand, *p*-type silicon seems to soften to a greater extent than germanium. The effect is proportional to light intensity and is affected by surface preparation."

537.311.33 : 535.215 **1173**

Measurement of Minority-Carrier Lifetimes with the Surface Photovoltage.—E. O. Johnson. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1349–1353.) The method is based on the junction-like properties of a semiconductor surface. With a capacitive contact the photovoltage (in the millivolt range) may be detected, and is a linear function of excess-carrier density. This gives the same carrier decay constant as the photoconductivity method when the lowest diffusion mode prevails.

537.311.33 : 535.37 **1174**

Anomalous Variation of Band Gap with Composition in Zinc Sulpho- and Seleno-tellurides.—S. Larach, R. E. Shrader & C. F. Stocker. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 587–589.) "A monotonic variation of band gap with composition occurs for many binary solid solutions. Of some Group II-Group VI systems, ZnS-ZnSe shows this type of variation of band gap with composition, whereas ZnSe-ZnTe, ZnS-ZnTe show an anomalous minimum in a plot of band gap *versus* composition of the solid solution."

537.311.33 : 539.16 **1175**

Radiation-Induced Expansion of Semiconductors.—D. Kleitman & H. J. Yearian. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, p. 901.) Small areas (3 mm × 3 mm) of polished GaSb, InSb and Ge were irradiated with 9-MeV deuterons at temperatures below −130°C. Surface contours were then determined by interferometers.

537.311.33 : 546.28 **1176**

Effect of Oxygen on the Carrier Lifetime in Silicon.—D. H. Roberts, P. H. Stevens & P. H. Hunt. (*Nature, Lond.*, 28th Sept. 1957, Vol. 180, No. 4587, pp. 665–666.) Results of measurements made using an infrared absorption technique to determine oxygen content show that the presence of oxygen increases the carrier lifetime. This is explained by the hypothesis of recombination centres which can be nullified by chemical reaction with the oxygen or which compete with it for a place in the lattice.

537.311.33 : 546.28 1177

Behaviour of Oxygen in Plastically Deformed Silicon.—S. Lederhandler & J. R. Patel. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 239–242.) For a deformed sample with a dislocation density of 10^7/cm^2, the amplitude of the 9-μ infrared absorption band is reduced appreciably in 15 min at 1 000°C, while the undeformed sample with a dislocation density of 10^4/cm^2 requires about 12 h. The effect of deformation on light scattering is described.

537.311.33 : 546.28 1178

Energy Band Structure in Silicon Crystals by the Orthogonalized Plane-Wave Method.—F. Bassani. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 263–264.) Energies at the point $\mathbf{K} = 2\pi a^{-1}$ (1, 0, 0) are calculated, and energy curves are drawn as a function of **K** in the [100] direction.

537.311.33 : 546.28 1179

Infrared Absorption in *n*-Type Silicon.—W. Spitzer & H. Y. Fan. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 268–271.) Experiments using samples of various impurity in the spectral region 1–45 μ, show an absorption band at 1·5–5 μ as well as a smooth rise with wavelength. The absorption is proportional to carrier concentration.

537.311.33 : 546.28 1180

Annealing of Electron Bombardment Damage in Silicon Crystals.—G. Bemski & W. M. Augustyniak. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 645–648.) The annealing was studied between 200°C and 400°C by observing the recovery of the minority-carrier lifetime. The annealing was found to proceed with an activation energy of 1·3 eV. The significance of the results is discussed briefly.

537.311.33 : 546.28 1181

Thermal Breakdown in Silicon *p-n* Junctions.—J. Tauc & A. Abrahám. (*Phys. Rev.*, 15th Nov. 1957, Vol. 108, No. 4, pp. 936–937.) "It is shown that the normal avalanche breakdown of well-etched silicon *p-n* junctions can pass at higher temperatures into a kind of thermal breakdown at which most of the current flows through a small hot hole in the potential barrier."

537.311.33 : 546.28 1182

Surface Protection and Selective Masking during Diffusion in Silicon.—C. J. Frosch & L. Derick. (*J. electrochem. Soc.*, Sept. 1957, Vol. 104, No. 9, pp. 547–552.) In the apparatus described a carrier gas containing impurity elements passes over heated Si. Carriers such as H_2 and He produce serious pitting of the Si surface; oxidizing carriers, however, cause a protective SiO_2 layer to form. This layer also provides a selective mask against diffusion of certain donors and acceptors into Si.

537.311.33 : 546.28 1183

Preparation of Large-Area *p-n* Junctions in Silicon by Surface Melting.—E. Billig & D. B. Gasson. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1242–1245.) The surface melting was produced either by using direct r.f. power coupling or by using radiation heating. The junctions, which were formed by overdoping of the molten material, sustained a peak inverse voltage of approximately 40 ρ_n and 10 ρ_p for *n*-type and *p*-type base material of resistivity ρ_n and ρ_p respectively.

537.311.33 : 546.28 : 621.314.63 1184

Semiconductor Properties of Recrystallized Silicon in Aluminium Alloy Junction Diodes.—R. A. Gudmundsen & J. Maserjian, Jr. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1308–1316.) The Al concentration was found to be largest in the region first recrystallized decreasing in a linear manner to the surface immediately beneath the Al-Si eutectic; the average concentration was about 7×10^{18} cm^{-3}. The average Hall hole mobility was about 55 cm^2/V.sec, while the conductivity varied from 72 to 35 mho. cm^{-1}.

537.311.33 : 546.289 1185

Dislocation Arrays in Germanium.—W. W. Tyler & W. C. Dash. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1221–1224.) "Dislocation arrays have been observed in deformed germanium using a technique based on the selective etching of lithium precipitates which prefer to nucleate on or near dislocations. Evidence is presented which indicates that both internal Frank-Read sources and surface dislocation sources are of importance in the deformation of germanium."

537.311.33 : 546.289 1186

Hillocks, Pits, and Etch Rate in Germanium Crystals.—B. W. Batterman. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1236–1241.) Symmetric, faceted etch hillocks have been produced on certain Ge surface orientations by an H_2O_2-HF etchant. General characteristics of hillock formation are explained in terms of the variation of etch rate with surface orientation.

537.311.33 : 546.289 1187

Surface Studies on Single-Crystal Germanium.—S. G. Ellis. (*J. appl. Phys.*, Nov. 1957, Vol. 28, No. 11, pp. 1262–1269.) The condition of the surface has been studied by chemical, electron-microscope, electron-diffraction, and other techniques after several of the standard etching procedures. It was found that the surface was often partially covered by particles believed to be GeO. Details of an etch which gives a controlled thickness of GeO and of an etch which minimizes the oxide formation are given.

537.311.33 : 546.289 : 538.24 1188

Magnetic Susceptibility of Germanium.—R. Bowers. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 683-689.) A study of the contribution of extrinsic charge carriers to the magnetic susceptibility of Ge as deduced from measurements of susceptibility between 1·3° K and 300° K. The results are compared with theoretical estimates based on the effective-mass values given by cyclotron-resonance experiments. The susceptibility of high-purity Ge is found to be independent of temperature below 60° K.

537.311.33 : 546.289 : 538.63 1189

Effect of Impurity Scattering on the Magnetoresistance of *n*-Type Germanium.—M. Glicksman. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 264–267.) A decrease in anisotropy of the magnetoresistance with increased ionized-impurity content and decreased temperature is shown. Deviations from symmetry conditions are observed for electron concentrations $>4 \times 10^{15}$ cm^{-3} at 77° K.

537.311.33 : 546.47-31 1190

Hall-Effect Studies of Doped Zinc Oxide Single Crystals.—A. R. Hutson. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 222–230.) Doping with H or interstitial Zn or Li permitted a single-donor-level analysis. The conductivity and Hall coefficient were measured between 55°K and 300°K. An experimental curve shows Hall mobility between 80°K and 600°K. At 300°K it is 180 cm^2/V. sec, and increases with decreasing temperature (T). The probable active scattering modes are discussed. Optical-mode scattering appears important above 200°K. Curves of electron concentration and carrier concentration as functions of $1/T$ are given. The low-frequency dielectric constant of ZnO was found to agree with Kamiyoshi's value of 8·5.

537.311.33 : 546.561-31 : 539.23 1191

Crystal Structure and Magnetic Susceptibility of Rectifying Cuprous Oxides.—K. R. Dixit & V. V. Agashe. (*Indian J. Phys.*, Sept. 1957, Vol. 31, No. 9, pp. 466–482.) The properties of films formed at different temperatures have been studied. Above 800°C rectification is appreciable and is accompanied by changes in crystal structure and susceptibility. A mechanism of film formation giving crystallites of Cu_2O with an excess of oxygen is suggested. See also 1140 of 1957 (Dixit & Agashe).

537.311.33 : 546.682.86 : 537.312.9 1192

Piezoresistance of Indium Antimonide.—R. F. Potter. (*Phys. Rev.*, 1st Nov. 1957, Vol. 108, No. 3, pp. 652–658.) The piezoresistivity coefficients were measured for both *n*- and *p*-type InSb over the range 77°K–300°K. The results for extrinsic *p*-type material are similar to those for *p*-type Si and Ge. The results for extrinsic *n*-type material confirm the picture of a conduction band having its minimum at the centre of the Brillouin zone.

537.311.33 : 546.682.86 : 538.63 1193

Galvanomagnetic Effects in *n*-type Indium Antimonide at Very Low Temperatures.—Y. Kanai & W. Sasaki. (*J. Phys. Soc. Japan*, Oct. 1957, Vol. 12, No. 10, p. 1169.) An anomalous Hall effect was investigated.

537.311.33 : 546.817.241 : 538.63 1194

The Galvanomagnetic Effects in Single Crystal of PbTe.—K. Shogenji & S. Uchiyama. (*J. phys. Soc. Japan*, Oct. 1957, Vol. 12, No. 10, p. 1164.) Results of measurements of magnetoresistance and planar Hall effect in *p*-type material are given.

537.311.33 : 546.824-31 1195

Electrical Resistivity of some Titanium Dioxide Semiconductors.—V. Andresciani, L. Nicolini & D. Sette. (*Note Recensioni Notiz.*, July/Aug. 1957, Vol. 6, No. 4, pp. 511–524.) Report of experimental investigations on three specimens, one being pure rutile, and the others containing small additions of Ba and Ca, respectively. See also 757 of 1954 (Breckenridge & Hosler).

537.312.62 : 538.569.4 **1196**

Microwave Measurements of the Energy Gap in Superconducting Aluminium.—M. A. Biondi, M. P. Garfunkel & A. O. McCoubrey. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 495–497.) Experimental evidence is presented for a temperature-dependent energy gap, extrapolating to a value between 3·0 and 3·5 kT_c at absolute zero.

537.312.62 : 538.569.4 **1197**

Millimetre-Wave Studies of Superconducting Tin.—M. A. Biondi, A. T. Forrester & M. P. Garfunkel. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 497–498.) A description of measurements of the temperature variation of the power transmitted by a superconducting tin waveguide, and brief comparison with theory.

537.312.62 : 539.23 **1198**

Conductivity of Superconducting Films for Photon Energies between 0·3 and $40kT_c$.—R. E. Clover, III & M. Tinkham. (*Phys. Rev.*, 15th Oct. 1957, Vol. 108, No. 2, pp. 243–256.) An investigation of far-infrared and mm-wave transmission through thin superconducting lead and tin films are reported, including measurements in the unexplored frequency region in which superconduction changes to normal. At $T=0$, σ_1 is very small for photon energies below $3\,kT_c$; above $3\,kT_c$, σ_1 rises rapidly to a limit σ_n at $20\ kT_c$. A gap in the excitation spectrum, of width $3\,kT_c$, is suggested.

538.22 **1199**

Magnetic Properties of the $FeTiO_3$-Fe_2O_3 Solid Solution Series.—Y. Ishikawa & S. Akimoto. (*J. phys. Soc. Japan*, Oct. 1957, Vol. 12, No. 10, pp. 1083–1098.) Experiments on $x\mathrm{FeTiO_3}.(1-x)\,\mathrm{Fe_2O_3}$, showed that its properties could be classified into three groups according to x. Simple molecular field theory does not explain all magnetic properties although a preferential distribution of Ti ions could account for some of them.

538.22 **1200**

Magnetic Properties of $NiTiO_3$-Fe_2O_3 Solid Solution Series.—Y. Ishikawa. (*J. phys. Soc. Japan*, Oct. 1957, Vol. 12, No. 10, p. 1165.)

538.221 **1201**

Effect of Isothermal Magnetic Annealing on the Magnetic Properties at Room Temperature in Ni-Co Alloys.—H. Masumoto, H. Saitô & M. Takahashi. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, Oct. 1957, Vol. 9, No. 5, pp. 374–394.) The magnetostriction and the magnetization at room temperature in a wide range of Ni-Co alloys were measured after annealing at high temperatures under external magnetic fields of various strengths. The results of the measurements are presented in 36 families of graphs.

538.221 : [621.318.124+621.318.134 **1202**

Convention on Ferrites.—(*Proc. Instn elect. Engrs*, Part B, 1957, Vol. 104, Supplement No. 5, pp. 127–266.) The following papers were included among those read at the I.E.E. Convention held in London 29th October–2nd November 1956.

Chemical and Physical Properties and Preparation:

(*a*) The Chemistry and Crystal Structure of Ferrites and Other Magnetic Oxides.—A. J. E. Welch (pp. 138–140).

(*b*) Structural and Magnetic Properties of Solid Solutions of Lithium Ferrite with Cadmium Ferrite and with Lithium Aluminate.—A. E. Carter, P. A. Miles & A. J. E. Welch (pp. 141–144).

(*c*) The Physical and Chemical Properties of some Nickel-Zinc Ferrite Compounds.—N. C. Tombs & J. Watkins (pp. 145–151).

(*d*) A Note on Crystal Formation in the Metallic Oxides.—L. R. Maxwell & S. J. Pickart (pp. 152–153).

(*e*) Ferrite Materials for Faraday Rotation at Wavelengths of 3, 6 and 10 cm.—P. E. Ljung (pp. 154–158).

(*f*) The Preparation of Magnesium-Manganese Ferrite for Microwave Applications.—A. E. Robinson (pp. 159–164.)

(*g*) The Properties of Manganese-Zinc Ferrites and the Physical Processes Governing them.—C. Guillaud (pp. 165–173).

Discussion (pp. 174–178).

Magnetic Spectra:

(*h*) Magnetic Spectra.—J. B. Birks (pp. 179–188).

(*i*) Losses in Ferrites: Single-Crystal Studies.—J. K. Galt (pp. 189–197).

(*j*) Effect of Magnetocrystalline Anisotropy on the Magnetic Spectra of Mg-Fe Ferrites.—G. T. Rado, V. J. Folen & W. H. Emerson (pp. 198–205).

(*k*) Ferrimagnetic Resonance in a Magnesium-Manganese Ferrite.—K. J. Standley & J. Peters (pp. 206–208). See also 1153 of 1957 (Standley & Stevens).

(*l*) Complex Susceptibility of High-Resistivity Ferrites.—P. M. Prache & B. Chiron (pp. 209–212).

Discussion (pp. 213–216).

Molecular Interaction:

(*m*) Neutron-Diffraction Studies of the Manganese-Magnesium Ferrite System.—R. Nathans, S. J. Pickart, S. E. Harrison & C. J. Kriessman (pp. 217–220). See also 510 of 1957 (Callen et al.).

(*n*) Magnetic Exchange Mechanisms in Magnesium-Manganese Ferrites—W. P. Osmond (pp. 221–227).

Discussion (pp. 228–230).

Direct-Current and Low-Frequency Properties:

(*o*) Heat Changes Accompanying Magnetization Processes in Ferrites.—L. F. Bates & D. A. Christoffel (pp. 231–237). See also 3569 of 1957 (Christoffel).

(*p*) Magnetocrystalline Anistropy in Cobalt-Substituted Magnetite Single Crystals.—L. R. Bickford, Jr, J. M. Brownlow & R. F. Penoyer (pp. 238–244).

Discussion (pp. 245–248).

New Materials:

(*q*) A New Class of Oxidic Ferromagnetic Materials with Hexagonal Crystal Structures.—G. H. Jonker, H. P. J. Wijn & P. B. Braun (pp. 249–254).

(*r*) Saturation Magnetization of Some Ferrimagnetic Oxides with Hexagonal Crystal Structures.—E. W. Gorter (pp. 255–260).

(*s*) Crystalline Structure and Magnetic Properties of Ferrites having the General Formula $5Fe_2O_3.3M_2O_3$.—F. Bertaut & R. Pauthenet (pp. 261–264).

Discussion (pp. 265–266).

538.221 : [621.318.124+621.318.134 **1203**

Synthesis of Some Ferrites.—H. Kedesdy & A. Tauber. (*J. Metals, N.Y.*, Sept. 1957, Vol. 9, No. 9, pp. 1140–1148.) Results are discussed of systematic investigations of the formation of a basic ferrite, such as Ni ferrite, a mixed ferrite, such as NiZn ferrite, and a ferrite involving a complex formation process, such as Mn ferrite. Trends in magnetic characteristics are related to changes in firing cycle and furnace atmosphere.

538.24 : 621.318.1 **1204**

Investigations of the Internal Demagnetization Factor.—G. Vogler. (*Ann. Phys., Lpz.*, 20th Dec. 1956, Vol. 19, Nos. 3–5, pp. 229–232.) Report on magnetic measurements on substances consisting of spherical ferromagnetic particles embedded in nonmagnetic material. The demagnetization factor is defined and its characteristics are discussed.

621-762.8 : 621.318.134 **1205**

Hermetic Seal for Ferrites.—A. H. Iversen. (*Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 797–799.) The electrical characteristics of ferrites are adversely affected by moisture. A hermetic seal is obtained by coating the ferrite surface with a thin layer of glass after sintering a thin layer of finely ground ferrite surface to reduce porosity. The glass seal has good mechanical and thermal shock properties and negligible effect on electrical properties.

537.311.33 **1206**

Rectifying Semiconductor Contacts. [Book Review]—H. K. Henisch. Publishers: Clarendon Press, Oxford, 1957, 372 pp., 70s. (*Nature, Lond.*, 21st Sept. 1957, Vol. 180, No. 4586, pp. 566–567.) ". . . a basic fund of knowledge upon which any investigator of surface and contact phenomena may usefully draw."

MEASUREMENTS AND TEST GEAR

53.082.5 : 519.24 : 621.397.3 **1207**

Optical Autocorrelation Measurement of Two-Dimensional Random Patterns.—L. S. G. Kovásznay & A. Arman. (*Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 793–797.) A function of two independent variables can be regarded as an image. Using a novel optical method the autocorrelation coefficient of such a function is formed as another image. The basic optical equipment is improved by adopting electronic scanning techniques.

621.3.018.41(083.74) **1208**
: 621.396.11.029.45

Comparison of Caesium Resonators by Transatlantic Radio Transmission.—L. Essen, J. V. L. Parry & J. A. Pierce. (*Nature, Lond.*, 14th Sept. 1957, Vol. 180, No. 4585, pp. 526–528.) A resonator at Teddington was indirectly compared with one of a different type at Camden, New

York, by means of the GBR 16-kc/s transmission from Rugby at about 0300 G.M.T. A table of results from 16th November 1956 to 1st June 1957 is given. Simultaneous measurements of suitable duration using day-time transmission should give results accurate to at least one part in 10^{10}.

621.317.2.013.782 **1209**

A Test Chamber Magnetically Screened by Magnetized Dynamo Laminations.—W. Albach & G. A. Voss. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 111–115.) Three concentric cylinders of lamination material of 120, 190 and 300 cm diameter form a screened enclosure. A 25-fold increase of the screening effect is achieved by magnetizing the toroidally wound cylinders with 50-c/s a.c.

621.317.335 : 518.4 **1210**

Rapid Measurement of Dielectric Constant and Loss Tangent.—D. M. Bowie & K. S. Kelleher. (*Trans. Inst. Radio Engrs*, July 1956, Vol. MTT-4, No. 3, pp. 137–140. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1898.)

621.317.335.2 : 621.319.4 **1211**

Precision Measurements on an Imperfectly Screened Three-Plate Capacitor.—W. Wiessner. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 120–125.) Sources of errors and their elimination are discussed.

621.317.335.3.029.64 : 537.226.8 **1212**

A Three-Centimetre Microwave Bench for Studying the Pressure and Temperature Effects on Dielectric Behaviour of Gases.—Krishnaji & G. P. Srivastava. (*J. sci. industr. Res.*, July 1957, Vol. 16A, No. 7, pp. 289–294.)

621.317.34.029.63/.64 : 621.372.5 **1213**

Three-Point Method of Measuring U.H.F. Quadripoles.—L. Mollwo. (*Hochfrequenztech. u. Elektroakust.*, May 1957, Vol. 65, No. 6, pp. 188–192.) A simplified method of determining network characteristics is described. It is applicable to simple types of quadripole and is based on locating three node positions.

621.317.35 : 534.232 **1214**

Measurements on Electromechanical Transducers.—H. G. Diestel. (*Acustica*, 1956, Vol. 6, No. 4, pp. 357–360. In German.) Apparatus is described for the continuous recording of frequency response curves of electromechanical transducers. Feedback into the driving system is compensated by means of a control circuit.

621.317.351.018.756 **1215**

A Simple Pulse Spectrograph.—N. R. Hansen. (*J. sci. Instrum.*, Oct. 1957, Vol. 34, No. 10, pp. 402–404.) Pulses from a radiation detector are displayed as bright dots along a nonlinear vertical timebase on a c.r. tube screen. The timebase is deflected towards the right by an amount proportional to the pulse amplitude. A photograph of the screen with time exposure thus shows a variable density pattern. Lines of constant density represent pulse frequency as a function of pulse amplitude.

621.317.361 **1216**

Pulse-Type Frequency Measurement.—J. C. Muller. (*Wireless World*, Feb. 1958, Vol. 64, No. 2, pp. 83–85.) Discusses the design of a beat-counting frequency comparator for the range 60–600 kc/s in which an unknown frequency is compared with a harmonic of a 1-kc/s standard. The harmonics are derived using a thyratron/delay-line circuit to generate a rectangular 1-μs pulse.

621.317.382.029.64 : 538.632 **1217**

The Hall Effect and its Application to Power Measurement at 10 Gc/s.—H. E. M. Barlow & S. Kataoka. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 53–60.) The instrument described is a transmission wattmeter based on the Hall effect produced in single crystals of *n*- and *p*-type Ge erected on the axis of a hollow metal rectangular waveguide carrying the power. The preparation of suitable crystals is described, and the reduction of errors due to the finite admittance of the crystal, thermal e.m.f.s and residual rectifier output is discussed. The Hall coefficient at nearly 10 kMc/s appeared to be of the same order of magnitude as the theoretical d.c. value. See also 1488 of 1956 (Barlow & Stephenson).

621.317.6.029.45 **1218**

Measurement of Voltage Ratio at Audio Frequencies.—W. C. Sze. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 444–449.) Detailed description of bridge circuits for precise voltage ratio and phase angle measurements. An accuracy within 0·005% in ratio and 0·2′ in phase angle is achieved at frequencies from 100 c/s to 10 kc/s.

621.317.7 : 621.376.2 : 621.396.62 **1219**

Modulation Check Meter for the Receiver.—M. W. Kirby. (*Short Wave Mag.*, Sept. 1957, Vol. 15, No. 7, pp. 350–351.) A circuit for direct modulation depth measurement.

621.317.7 : 621.383.27 : 621.396.822 **1220**

Photomultiplier Tubes as Standard Noise Sources.—Chenette, Shimeda & van der Ziel. (See 1286.)

621.317.714.023.43 **1221**

Clamp-on Microammeter Measures A.C. Current.—G. F. Montgomery & C. Stansbury. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 152–153.) "Small toroidal current transformer clamped around unknown current measures 0 to 200-microampere range over frequency band of 50 c/s to 100 kc/s with negligible reaction upon measured circuits. Feedback to tertiary winding supplies frequency correction to transistor amplifier."

621.317.725.082.72 **1222**

A Compact Rotating-Electrode Voltmeter for Wide Voltage Ranges.—W. Knauer. (*Z. angew. Phys.*, March 1957, Vol. 9, No. 3, pp. 115–119.) In the portable e.s. voltmeter described the rotating multisegment electrode is motor-driven and contained in a pressure chamber together with the h.v. electrode. Linear indication is provided in 6 ranges, 0–3 V up to 0–300 kV.

621.317.733.029.5 **1223**

A 'Subtractor' for Radio-Frequency Bridges.—E. A. Faulkner. (*J. sci. Instrum.*, Nov. 1957, Vol. 34, No. 11, p. 461.) A two-valve device with high-impedance input terminals. The output is proportional to the vector difference of the input potentials, with respect to earth, and can be fed into a normal receiver. Balancing is simple at frequencies up to 5 Mc/s. A full circuit diagram is given.

621.317.755 **1224**

Sampling Oscilloscope for Statistically Varying Pulses.—R. Sugarman. (*Rev. sci. Instrum.*, Nov. 1957, Vol. 28, No. 11, pp. 933–938.) Pulses of constant or random repetition rate may be displayed on the oscilloscope which has a rise time of $5{\cdot}6 \times 10^{-10}$s, maximum sweep speed of 10^{-10} s/in. and a maximum sensitivity of 15 mV/in.

621.317.755.087 : 621.3.015.3 **1225**

A High-Sensibility Cathode-Ray Tube for Millimicrosecond Transients.—K. J. Germeshausen, S. Goldberg & D. F. McDonald. (*Trans. Inst. Radio Engrs*, April 1957, Vol. ED-4, No. 2, pp. 152–158. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1957, Vol. 45, No. 8, p. 1164.)

621.317.784 **1226**

A Precision Thermoelectric Wattmeter for Power and Audio Frequencies.—J. J. Hill. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 61–68.) Compensated thermojunctions are used with negative-feedback amplifiers to achieve accuracies within about 0·1% over the range 200 c/s–10 kc/s and 0·3% for 50 c/s–30 kc/s. A method of extending the frequency range to 100 kc/s is described.

621.317.794 : 621.372.822 **1227**

Transverse-Film Bolometers for the Measurement of Power in Rectangular Waveguides.—J. A. Lane. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 77–80.) Describes a plunger-backed bolometer consisting of a relatively narrow absorbing strip (platinum sputtered on mica) placed symmetrically in the region of the transverse plane where the electric and magnetic fields are substantially uniform. Substitution of d.c. for microwave power is therefore used, the power range being 1–100 mW with an error of 3% or less.

621.317.794 : 621.396.822 **1228**

Considerations in High-Sensitivity Microwave Radiometry.—P. D. Strum. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 43–60.) A discussion of the factors which limit sensitivity of receiving systems. Apart from fluctuations in the receiver and its aerial, background radiation from space, atmospheric oxygen, water vapour and earth-bound radiators must be considered. A square-wave switched type of receiver is most likely to yield satisfactory sensitivity and the optimum performance is discussed.

621.317.794 : 621.396.822 : 523.164 **1229**

A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy.—F. D.

Drake & H. I. Ewen. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 53–60.) A travelling-wave-valve radiometer operating at 8 kMc/s with bandwidth 1 kMc/s and sensitivity of the order of 0·01°K is described. Effects of gain fluctuations are eliminated by introducing compensating noise. Results obtained using a 28-ft. reflector are discussed.

621.317.794.089 : 523.164 **1230**
A Method of Calibrating Centimetric Radiometers using a Standard Noise Source.—J. S. Hey & V. A. Hughes. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 119–121.) The method is suitable for aerial temperatures in the range of 0–1 000°K with an accuracy within 1°K. The source is an argon discharge tube.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.384.61 **1231**
Circular Electron Accelerators.—M. Seidl. (*Slab. Obz., Praha*, Dec. 1956, Vol. 17, No. 12, pp. 698–702.) A description is given of particle accelerators made in Czechoslovakia including a 15 Mev betatron made in 1955, earlier smaller betatrons and a 3-MeV synchrotron.

621.384.61 **1232**
C.E.R.N. 600-MeV Synchrocyclotron.—(*Engineer, Lond.*, 11th Oct. 1957, Vol. 204, No. 5307, pp. 538–540.) Constructional details of the equipment.

621.384.613 **1233**
Engineering Problems in the Development of Glass Betatron Toroids.—S. J. Morrison. (*J. Soc. Glass Tech.*, Dec. 1956, Vol. 40, No. 197, pp. 520T–541T.) Three different types of toroidal glass vacuum chambers which have been developed for use in betatrons operating at 16 and 20 MeV are reviewed.

629.113 : 621.317.39 **1234**
Electronics in Automobile Research.—C. H. G. Mills & J. F. Winterbottom. (*Brit. Commun. Electronics*, July 1957, Vol. 4, No. 7, pp. 410–415.) Illustrated description of equipment used for electronic recording of stress measurements, speed measurement by means of lamp/photocell units, and vehicle noise measurement using ⅓-octave band-pass filters.

681.61 : 621.318.57 **1235**
***L'Electrostyl* to Revolutionize Stenography?**—A. V. J. Martin. (*Radio-Electronics*, March 1957, Vol. 28, No. 3, pp. 38–39.) Brief description of an electronically controlled typewriter, a French invention in which the conventional keyboard is replaced by a flat panel with contact studs representing characters. A single conducting probe is drawn across the panel completing a switching circuit whenever a contact is touched; this actuates the keys of a typewriter.

PROPAGATION OF WAVES

621.396.11 : 523.5 **1236**
Meteor Signal Rates Observed in Forward Scatter.—E. L. Vogan & L. L. Campbell. (*Canad. J. Phys.*, Oct. 1957, Vol. 35, No. 10, pp. 1176–1189.) Measurements at 49·98 Mc/s over a path 860-km long in eastern Canada are described. Variations in meteor signal rate on an hourly, daily and monthly time scale show a large spread, the maximum rate occurring in the summer months. An explanation of certain features in the diurnal variation is based on possible attenuation in the E region.

621.396.11 : 551.510.535 **1237**
New Type of Scattering Echo Observed by the Ship-Borne Ionospheric Sounder over the Sea.—H. Okamoto, M. Ose & K. Aida. (*Rep. Ionosphere Res. Japan*, June 1957, Vol. 11, No. 2, pp. 50–54.) A note on new short-range echoes having variation with time of day, frequency, and type of terrain; records are shown. The scattered echoes appear to be related to tropospheric conditions.

621.396.11 : 551.510.535 **1238**
The Attenuation of Radio Waves Reflected from the E Region of the Ionosphere.—R. W. Meadows. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 22–26.) The absorption measured at 5·1 Mc/s over a 740-km path between Slough and Inverness is compared with that measured simultaneously at Slough at vertical incidence on 2–5 Mc/s. It is suggested tentatively that the absorption measurements which have been made for many years on 2 Mc/s at noon at Slough should prove a useful guide to the absorption likely to be obtained over other oblique E-region paths during any part of the sunspot cycle, and they have been formulated accordingly.

621.396.11 : 551.510.535 : 550.38 **1239**
The Effect of the Earth's Magnetic Field on Absorption from a Single-Hop Ionospheric Path.—R. W. Meadows & A. J. G. Moorat. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 33–37.) Magneto-ionic calculations show that deviative absorption is not necessarily negligible at vertical incidence for waves reflected from the E region at frequencies considerably below the penetration value. Deviative absorption on paths sufficiently oblique is, however, negligible. The effect of the earth's magnetic field on Martyn's absorption theorem is shown to be similar to the effect of losses due to partial reflections. The absorption to be expected on a radio path might best be calculated by applying the conventional non-deviative formula to measurements made at oblique rather than at vertical incidence.

621.396.11 : 551.510.535 **1240**
: 621.396.812.3
The Effect of Fading on the Accuracy of Measurement of Ionospheric Absorption.—R. W. Meadows & A. J. G. Moorat. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 27–32.) An analysis is made of measurements of the amplitude of first-order reflections from the E region. The accuracy of measurement of the smoothed value of field strength is defined as the range of values having a 99% chance of containing the correct value. On this basis, the accuracy of noon absorption, as calculated from a single day's observations at one frequency, has been estimated to be within about +4 and −12 dB at vertical incidence and +4 and −11 dB for a particular oblique-incidence case.

621.396.11.029.55 **1241**
Broadcasting Plan: 26 Mc/s and 'Phase: June 70'.—R. Gea Sacasa. (*Rev. Telecommunicación, Madrid*, March 1957, Vol. 11, No. 47, pp. 43–49.) Critical comparison of some m.u.f. charts issued for various epochs by the International Frequency Registration Board with predictions based on the Gea method (see e.g. 558 of 1957).

621.396.62 : 621.317.794 **1242**
Present and Future Capabilities of Microwave Crystal Receivers.—C. T. McCoy. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 61–66.) The effective temperatures of present-day receivers using crystal mixers are about 600°K at frequencies up to 10 kMc/s but increase rapidly at higher frequencies. Future developments in crystals and the introduction of cooling mechanisms may reduce effective temperatures to 150°K at all frequencies up to 100 kMc/s.

621.396.62 : 621.375.2 **1243**
High-Selectivity I.F./A.F. Amplifier Unit.—A. C. Edwards. (*Short Wave Mag.*, Sept. 1957, Vol. 15, No. 7, pp. 344–349.) Design details for amateur construction of a circuit incorporating many of the latest reception techniques.

RECEPTION

621.396.62 : 621.317.7 : 621.376.2 **1244**
Modulation Check Meter for the Receiver.—Kirby. (See 1219.)

621.396.621 : 621.314.7 **1245**
Audio Stages for All-Transistor Portables.—O. J. Edwards & L. H. Light. (*Mullard tech. Commun.*, Oct. 1957, Vol. 3, No. 26, pp. 188–194.) A driver stage is transformer coupled to a class B single-ended output stage. Two versions are described having outputs of 200 and 100 mW and battery drains of 20 and 10 mA respectively.

621.396.621 : 621.396.215 **1246**
New Frequency-Shift Telegraphy System.—(*Wireless World*, Feb. 1958, Vol. 64, No. 2, pp. 93–94.) Brief account of a system of frequency-diversity reception described in 2278 of 1957 (Allnatt et al.).

621.396.621.53 : 621.396.662.1 **1247**
A New Method of Analysing Tracking in Superheterodyne Receivers.—J. Holownia. (*Nachr Tech.*, July 1957, Vol. 7,

No. 7, pp. 289–293.) The method described is independent of the i.f., the tuning range and tuning system of the receiver.

621.396.665 : 621.375.4 **1248**

Transistor Amplified A.G.C. — W. Woods-Hill. (*Wireless World*, Feb. 1958, Vol. 64, No. 2, pp. 94–95.) Description of a circuit suitable for miniature receivers, in which very large variations of signal strength are encountered. A single junction transistor combines the function of second detector, a.f. amplifier and a.g.c. amplifier.

621.396.822 : 621.317.794 **1249**

Considerations in High-Sensitivity Microwave Radiometry.—Strum. (See 1228.)

621.396.828 **1250**

The Problem of Eliminating Radio Interference Originating from High-Frequency Heating Installations.—E. Rosemann. (*NachrTech.*, July 1957, Vol. 7, No. 7, pp. 297–302.) The design of screens and suppressors is discussed and the difficulties of measuring the interference field strength are outlined.

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 **1251**

Distortion in F.M. Systems.—A. Ditl. (*Slab. Obz., Praha*, Nov. 1956, Vol. 17, No. 11, pp. 609–617.) See also 3291 of 1957.

621.396.3 : 621.396.41 **1252**

A Predicted Wave-Signalling Phase-Shift Telegraph System.—E. T. Heald & R. G. Clabaugh. (*Commun. & Electronics*, July 1957, No. 31, pp. 316–319. Discussion, p. 319.) Highly stable mechanical resonator filters in the detection equipment are turned on and off in step with the arrival and conclusion of each signal pulse; the detector is thus subjected to noise only when a signal is present. A typical 40-channel system ('kineplex') is described, and a superiority in signal/noise ratio of 8 dB over a wide-band frequency-shift keyed system is claimed.

621.396.3 : 621.396.43 : 523.5 **1253**

The Janet Communication System.—G. W. L. Davis. (*Brit. Commun. Electronics*, July 1957, Vol. 4, No. 7, pp. 402–407.) Description of the method of operation and a note on equipment used. See 3533 of 1956 (Cocking).

621.396.41 **1254**

The Canadian Transcontinental Microwave System.—S. Bonneville. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 473–477.) The system under construction will cover 4 800 miles with about 137 repeater points. It operates in the 4-kMc/s band, with six radio channels in each direction. Each 2-way channel can handle two television programmes or 600 telephone circuits.

621.396.41 : 621.314.7 **1255**

A Transistorized Time-Division Multiplex Telegraph Set.—F. D. Biggam. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 780–785.) A description of a fully transistorized 4-channel equipment which saves 80% in weight and volume and 95% in power consumption compared with the equivalent valve equipment Type AN/FGC-5.

621.396.41 : 621.376 **1256**

The Third Method of S.S.B.—H. F. Wright. (*QST*, Sept. 1957, Vol. 41, No. 9, pp. 11–15.) A description of the method previously given by Weaver (*Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, pp. 1703–1705) which produces the s.s.b. signal directly on the required radio frequency without the use of wide-band a.f. phase-shift networks.

621.396.44 : 621.395.97 : 621.396.82 **1257**

High-Frequency Broadcasting over Lines and R.F. Interference.—J. Meyer de Stadelhofen. (*Tech. Mitt. PTT*, 1st July 1957, Vol. 35, No. 7, pp. 257–265. In French.) A description of methods used for measuring and eliminating interference from various sources in program transmission over telephone lines, with particular reference to the Swiss system.

621.396.65 **1258**

Light-Route Microwave Systems in Canada.—A. J. Dinnin. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 488–492.) Description of systems and equipment used for auxiliary links in the Trans-Canada microwave system (1254 above). Frequencies in the 900-Mc/s 2-, 4- and 6-kMc/s bands are used, and diversity reception is found necessary, particularly for telephone transmission and at the higher frequencies.

621.396.933 : 621.396.41 **1259**

Dual-Purpose Circuitry cuts Transceiver Size.—P. G. Wulfsberg & C. H. Kirkpatrick. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 134–138.) Reflex circuit techniques are applied in the design of a 1 750-channel transceiver using only 35 crystals and 28 valves. The unit is intended for fixed- or mobile-station use in a ground-to-air communications system in the frequency range 225–400 Mc/s. Transmitter power is 15 W and receiver sensitivity better than 5 μV for a 10-dB signal-plus-noise/noise ratio. The operation of a 'squelch' circuit is controlled by this ratio.

SUBSIDIARY APPARATUS

621-519 : 621.318.57 : 621.314.7 **1260**

Transistor Relays Have Low Idling Current.—D. W. R. McKinley. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, p. 147.) Two transistor circuits for remote control are shown, which operate a 2- or 3-W electromechanical relay from the output signal of a receiver. This may be a carrier modulated at 1 kc/s or a pulsed microwave signal. Using Si transistors the idling current is a few microamps.

621.311.6 : 621.314.7 : 621.316.72 **1261**

Magnetic-Rehulation Transistor Power Supply.—L. F. Lyons. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 643–645.) The method uses saturable reactors in the input and load circuit of the supply. For closer regulation a silicon-diode reference voltage may be used.

621.311.62 **1262**

Compact Supplies Have Wide-Range Regulation.—W. F. Schreiber. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 168–169.) "Use of solid-state rectifiers and high-current regulator tubes eliminates power transformer and reduces size and cost of power supplies. Units have regulation better than 1%, ripple of 0·01% and stacking factor of about 150 mA capacity per inch of rack space."

621.311.62 : 621.314.2 **1263**

A Method of Designing Small Power Transformers for Communication Equipment.—H. S. Sear. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 758–761.)

621.311.62 : 621.372.54 **1264**

The Effect of Capacitance on Power-Supply Filter Bounce.—D. T. Geiser. (*QST*, Sept. 1957, Vol. 41, No. 9, pp. 27–30.) A discussion on the design of power-supply filters.

621.311.62 : 621.375.4 **1265**

Regulated Supply for Transistor Amplifiers.—H. R. Lowry. (*Radio TV News*, Oct. 1957, Vol. 58, No. 4, pp. 55, 193–194.)

621.314.63 **1266**

Semiconductor Rectifiers.—D. Ashby. (*Elect. Rev., Lond.*, 4th Oct. 1957, Vol. 161, No. 14, pp. 587–592.) Summary of characteristics and review of industrial applications of Si, Ge, Se and copper-oxide rectifiers.

621.314.63 **1267**

A New Voltage-Divider Circuit for Semiconductor Rectifiers.—I. K. Dortort. (*Commun. & Electronics*, July 1957, No. 31, pp. 356–358. Discussion, p. 358.) Auxiliary transformers divide the inverse voltage across high-power rectifiers.

621.314.63 : 546.28 **1268**

The Fused Silicon Rectifier.—H. W. Henkels. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 733–746.) A comprehensive review dealing with (*a*) rectification theory (*b*) basic fabrication techniques and the properties of Si, the counter-electrode alloy solder, the base solder and the base electrode, (*c*) encapsulation and heat dissipation, (*d*) theoretical and experimental properties of Si rectifiers.

621.314.63 : [546.289+546.28 **1269**

Rating and Application of Germanium and Silicon Rectifiers.—F. W. Gutzwiller. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 753–757.) A discussion of the factors limiting the life and rating of semiconductor rectifiers. Practical design criteria are considered including fusing, surge voltage protection and parallel and series operation.

621.314.63-713 : 546.289 **1270**

Direct - Water - Cooled Germanium Power Rectifier.—R. E. Wahl. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 832–841.) A description of the development, performance and operation of various types of water-cooled unit.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 (71) **1271**

The Development of Television in Canada.—J. E. Hayes. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 482–484.) General description of the growth of the television service, which now covers 80% of the population. Discussion of improvements in production facilities and system design is included.

621.397.5 : 535.623 **1272**

Colour-Television-System Performance Requirements.—R. Kennedy. (*Commun. & Electronics*, Jan. 1957, No. 28, pp. 653–659.) A discussion of the basic requirements of phase, amplitude and transient response which a transmission system must meet to handle a standard colour-television signal. The effect of departures from the required standards are illustrated and test procedures are described.

621.397.5 : 621.391 **1273**

Some Relations between Television Picture Redundancy and Bandwidth Requirements.—K. H. Powers & H. Staras. (*Commun. & Electronics*, Sept. 1957, No. 32, pp. 492–496. Discussion, p. 496.) Analysis of the statistical redundancy, i.e. correlations existing between the elements of a given picture or between corresponding elements of successive frames, in terms of information theory. It is concluded that the best way of achieving bandwidth reduction is by cleverly degrading the information in a picture in such a way that the observer would not notice it appreciably.

621.397.611.2 : 681.42.089 **1274**

New Equipment and Methods for the Evaluation of the Performance of Lenses for Television.—W. N. Sproson. (*B.B.C. Engng Div. Monographs*, Dec. 1957, No. 15, pp. 5–16.)

621.397.62 **1275**

An Original Television Receiver for 4 Standards and 12 Channels.—K. Lamus. (*Télévision*, Feb. 1957, No. 71, pp. 56–60.) Circuit details are given of a commercial-type receiver with switching facilities for adapting the set to the 625-line Belgian or European and the 819-line Belgian or French standards.

621.397.62 **1276**

Black Level—the Lost Ingredient in Television Picture Fidelity.—R. G. Neuhauser. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1957, Vol. 66, No. 10, pp. 597–601. Correction, *ibid.*, Dec. 1957, Vol. 66, No. 12, p. 775.) The necessity for correct control of black level (d.c. restoration) in television pictures of good fidelity is discussed. Advances in equipment design to achieve this control are described and a waveform standard is suggested for uniformity in black-level reproduction.

621.397.62 **1277**

Television Line Structure Suppression.—F. T. Thompson. (*J. Soc. Mot. Pict. Telev. Engrs*, Oct. 1957, Vol. 66, No. 10, pp. 602–606. Discussion, p. 606.) Methods of eliminating line structure are described. Experimental results obtained in comparing pictures with and without line structure suppression are given together with photographic comparisons.

621.397.62.001.4 : 535.623 **1278**

Flat-Field Generator speeds Colour TV Testing.—R. W. Cook. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 139–141.) Description of a generator providing video signals of any hue or saturation as well as a luminance signal variable from black to white.

TRANSMISSION

621.396.61 : 621-519 **1279**

A New Transmitter for Long Distance Communications.—V. Nemecek. (*Slab. Obz., Praha*, Jan. 1957, Vol. 18, No. 1, pp. 15–18.) The transmitter, using d.s.b. or s.s.b., operates in the 3–27·5-Mc/s band at powers of 6–60 kW. Remote tuning to any of eight carrier frequencies is possible.

621.396.61 : 621.376.22 **1280**

Transmitter Circuits for Suppressed-Carrier A.M.—J. P. Costas & R. W. French. (*Electronics*, 1st Dec. 1957, Vol. 30, No. 12, pp. 128–131.) Description and circuit details of a suppressed-carrier d.s.b. system applied to two communication transmitters. They operate with peak sideband power 100 W and 1 kW, respectively, over a frequency range 2–32 Mc/s, using class-C push-push power amplifiers with push-pull screen-grid modulation.

621.396.61 : 621.376.3 **1281**

New Trends in the Development of Large Transmitters.—J. Vackár. (*Slab. Obz., Praha*, Jan. 1957, Vol. 18, No. 1, pp. 2–9.) A survey of trends in broadcasting, communication, television and v.h.f. transmitters is given with special reference to modulation methods.

VALVES AND THERMIONICS

621.314.63 **1282**

On the Switching Transient in the Forward Conduction of Semiconductor Diodes.—H. L. Armstrong. (*Trans. Inst. Radio Engrs*, April 1957, Vol. ED-4, No. 2, pp. 111–113. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1957, Vol. 45, No. 8, p. 1163.) See also 2624 of 1957.

621.314.63 : 537.311.33 **1283**

On the Forward Characteristic of Semiconductor Diodes.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 361.) One expression, frequently quoted, for the current/voltage characteristic is said to be restricted to small voltages. Alternative formulae are given which fit experimental curves more closely.

621.314.7 **1284**

Transistor Cut-Off Frequency.—W. L. Stephenson. (*Electronic Radio Engr*, Feb. 1958, Vol. 35, No. 2, pp. 69–71.) The variation of cut-off frequency with collector voltage is calculated from known physical relations and provides a method for the measurement of the Early feedback factor. (See 874 of 1953.)

621.314.7 : 546.289 **1285**

Some Measurements on Commercial Transistors and their Relation to Theory.—F. J. Hyde. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 45–52.) The effective lifetimes of minority carriers in the bases of 4 types of Ge *p-n-p* alloy transistor and on one Ge surface-barrier transistor were measured. One - dimensional small - signal internal - transitor equations could be adapted to explain current-gain and cut-off-frequency parameters, provided that the bulk lifetime and diffusion constant of minority carriers in the base were replaced by effective values.

621.383.27 : 621.396.822 : 621.317.7 **1286**

Photomultiplier Tubes as Standard Noise Sources.—E. R. Chenette, K. Shimada & A. van der Ziel. (*Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 835–836.) A brief article indicating that photomultipliers may be used to produce very large 'equivalent saturated - diode currents' with negligible flicker effect down to 1 c/s.

621.385.002.2 **1287**

Background, Foreground and Horizon—the Radio Valve Industry in Prospect and Retrospect.—T. E. Goldup. (*Proc. Instn elect. Engrs*, Part B, Jan. 1958, Vol. 105, No. 19, pp. 1–10.)

621.385.029.6 : 621.372.56 **1288**

Travelling-Wave-Tube Limiters.—F. B. Fank & G. Wade. (*Trans. Inst. Radio Engrs*, April 1957, Vol. ED-4, No. 2, pp. 148–152. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1957, Vol. 45, No. 8, p. 1164.)

MISCELLANEOUS

621.37/.39 : 061.3 **1289**

Radio Scientists Attend U.R.S.I. Conference.—(*Tech. News Bull. nat. Bur. Stand.*, Dec. 1957, Vol. 41, No. 12, pp. 187–189.) Brief summary of the proceedings of the 12th general assembly held in Boulder, Colorado, from 22nd August to 5th September 1957.

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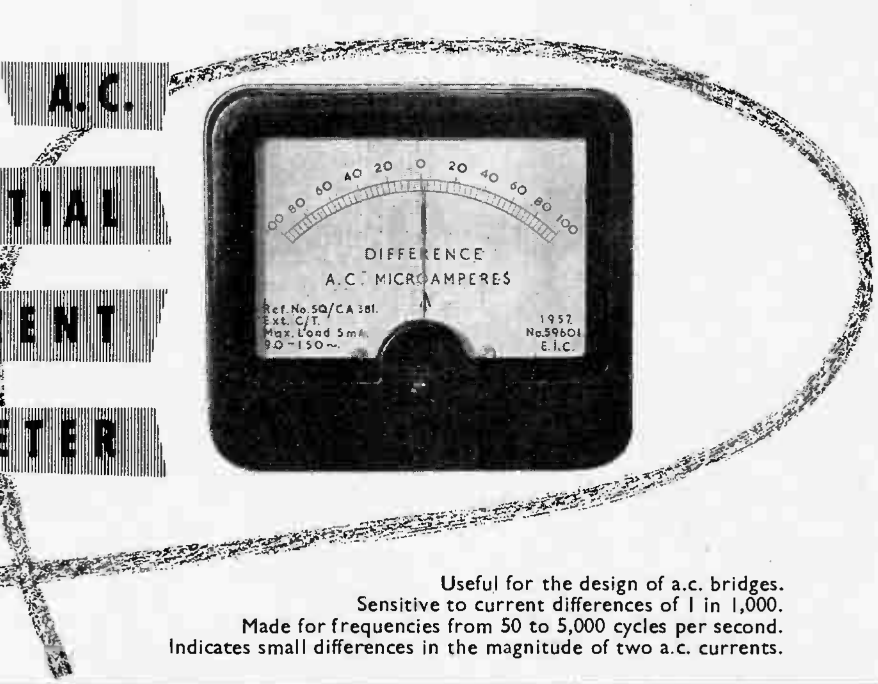

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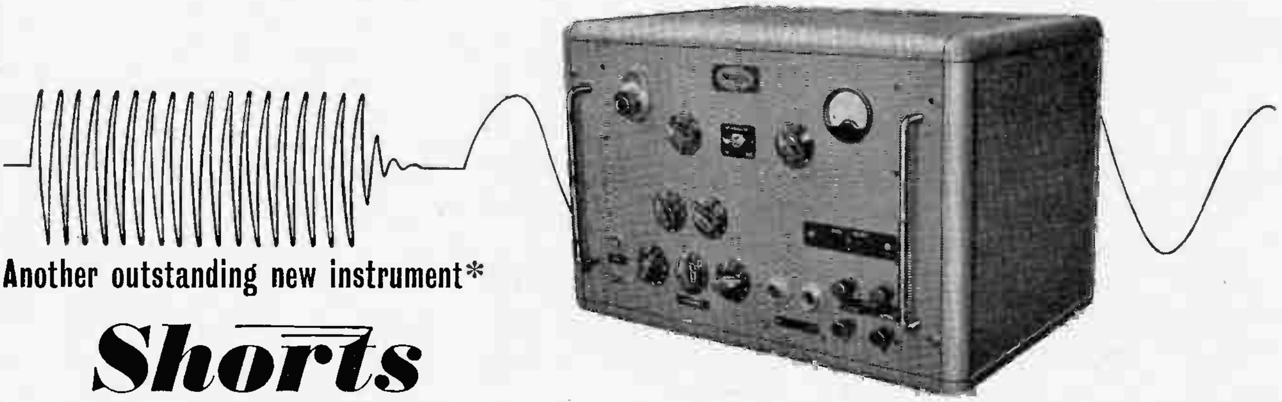

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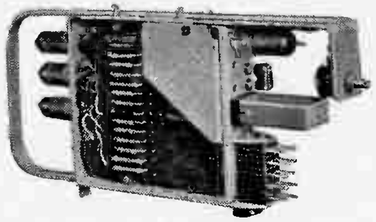

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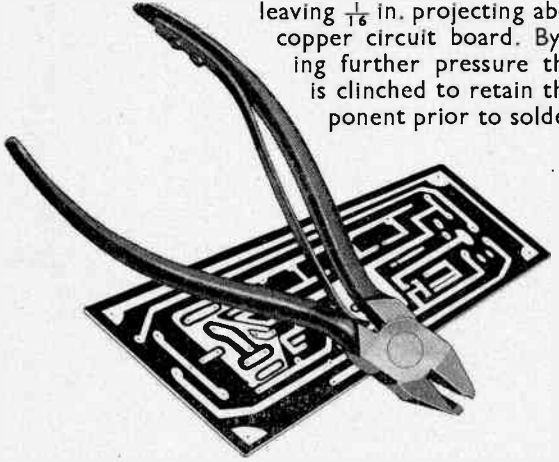

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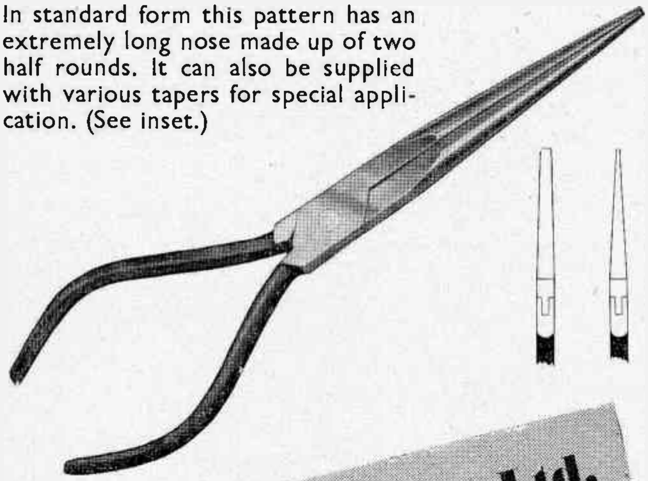

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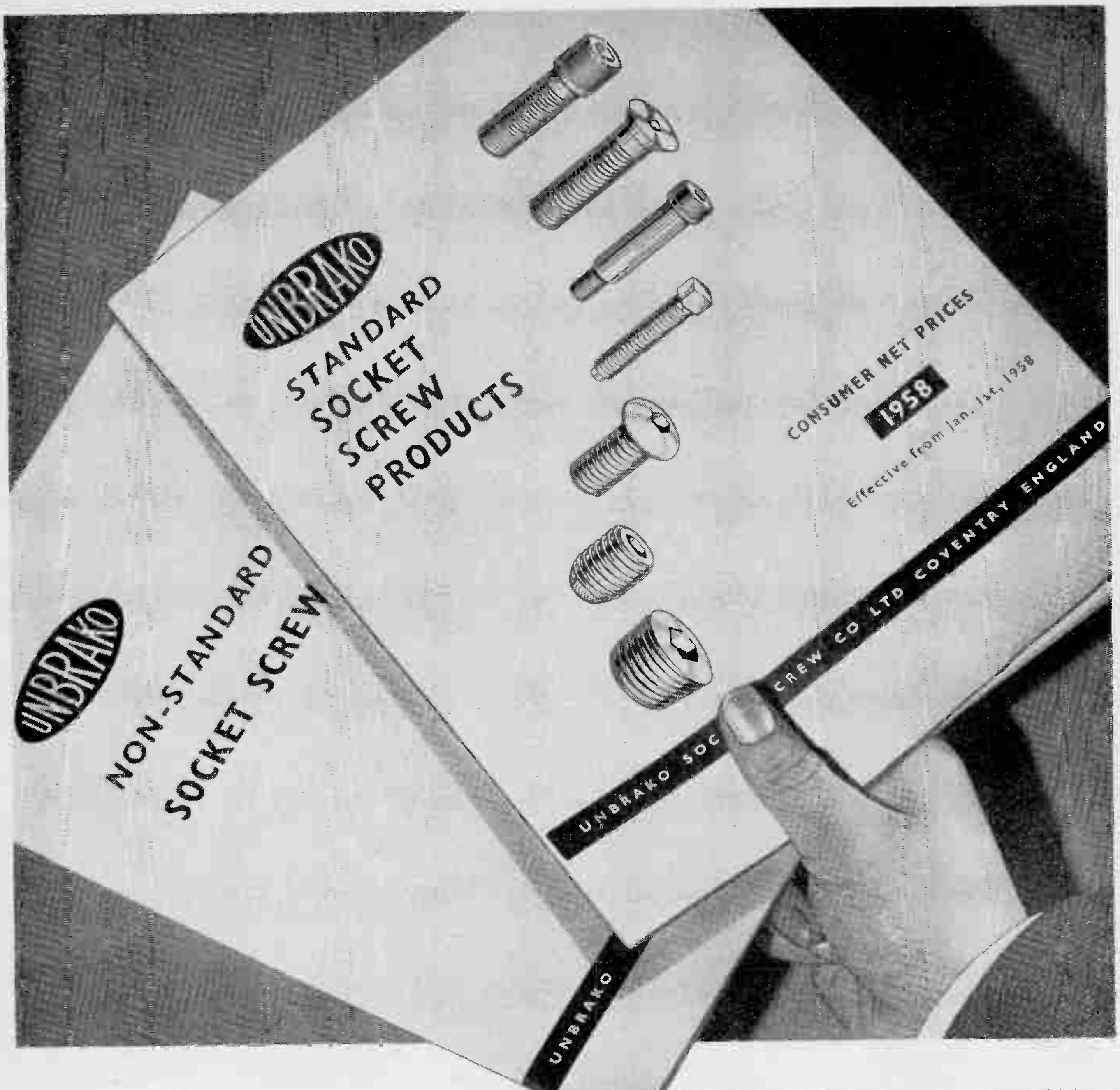

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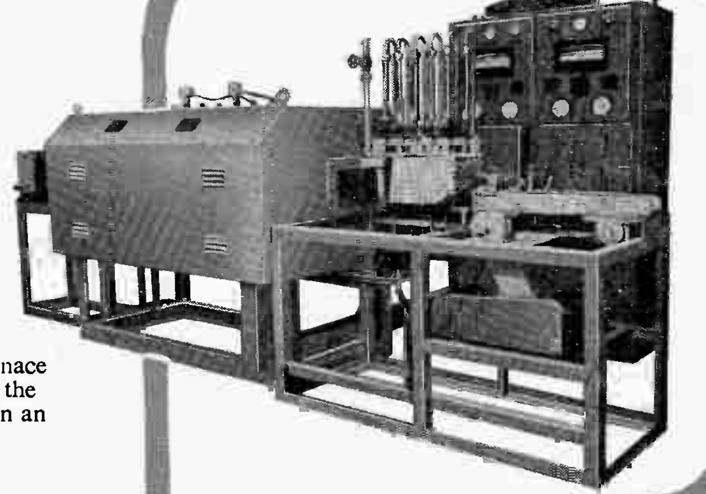

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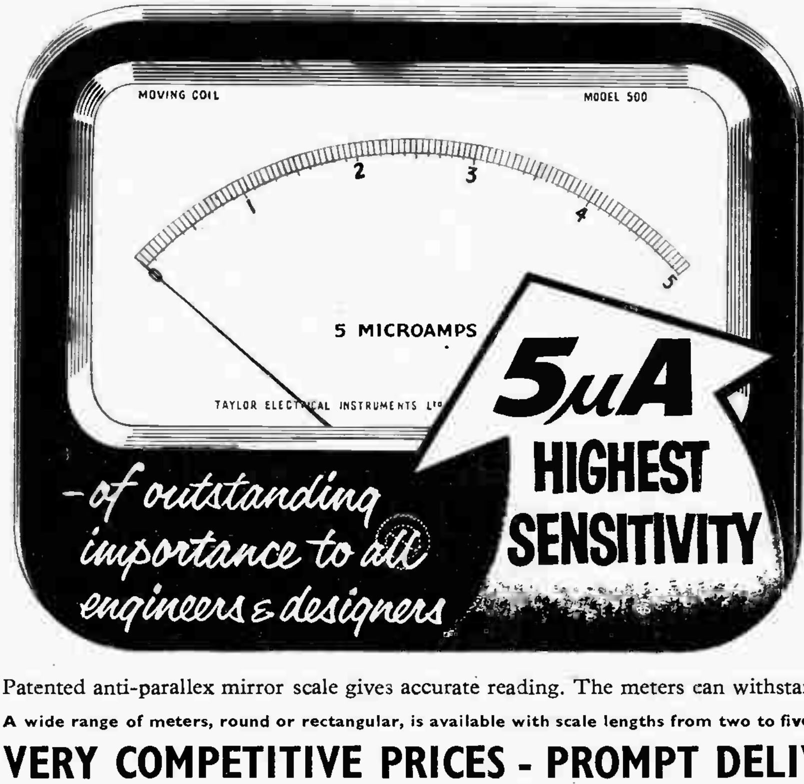

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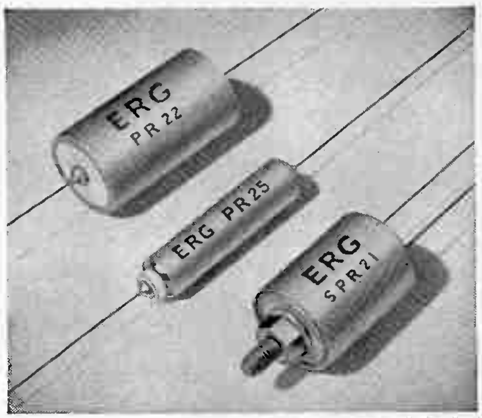

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PR 22A	SPR 22A	1½″	¾″	1·0	·65	·32	0·1	750K
PR 23	SPR 23	1¾″	1″	1·5	1·0	·5	0·1	2 Meg
PR 24		1⅞″	½″	1·0	·65	·32	0·1	350K
PR 25		1⅜″	⅜″	0·5	·32	·16	0·1	100K
PR 26	SPR 26	2½″	1″	2·0	1·32	·65	0·1	2·5 Meg

Test	Duration	Spec.	Average % Change
Full DC Load	1,000 hours		·013
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Climatic Cycles ...	6 Cycles	RCS II Climatic	·01
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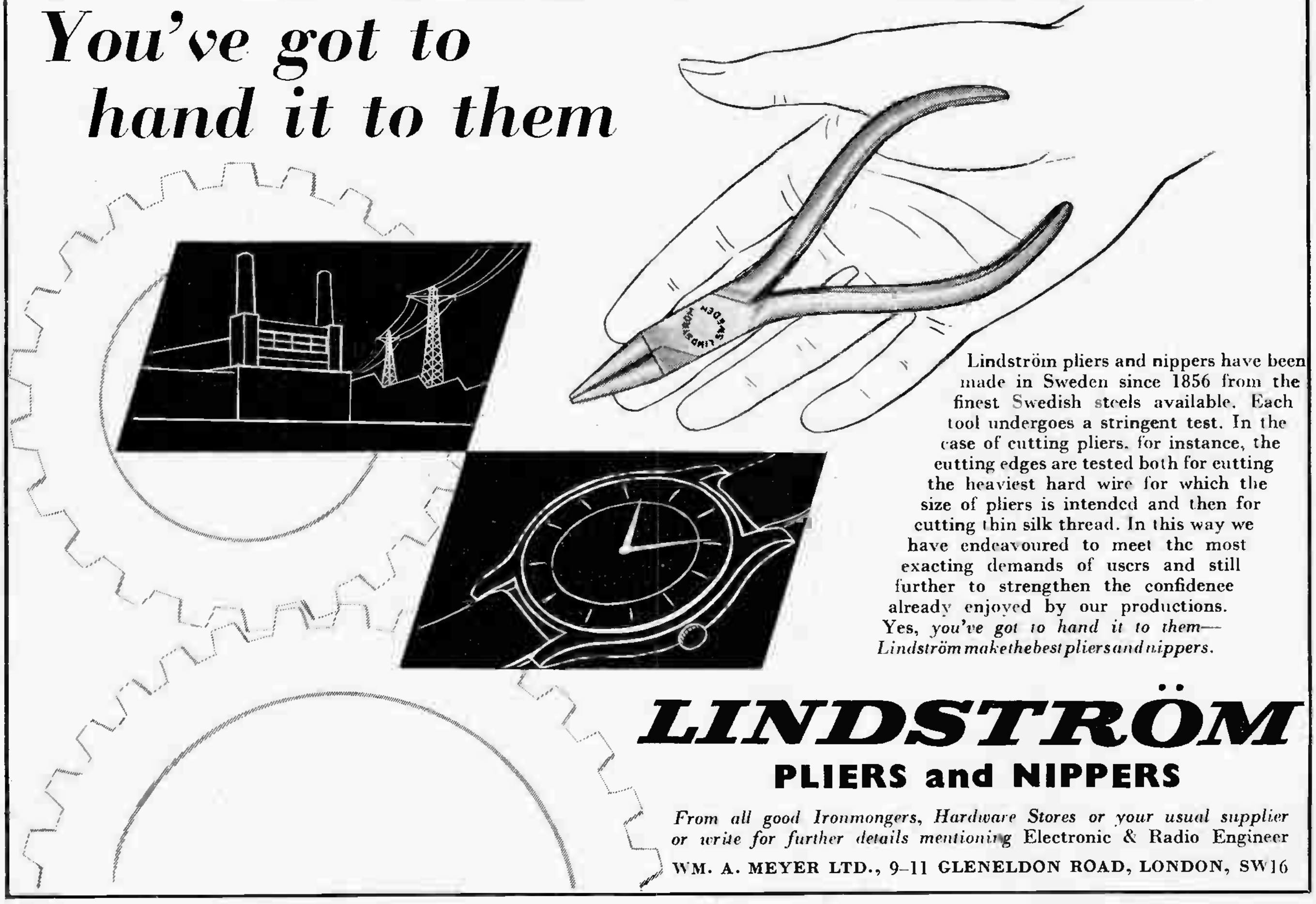

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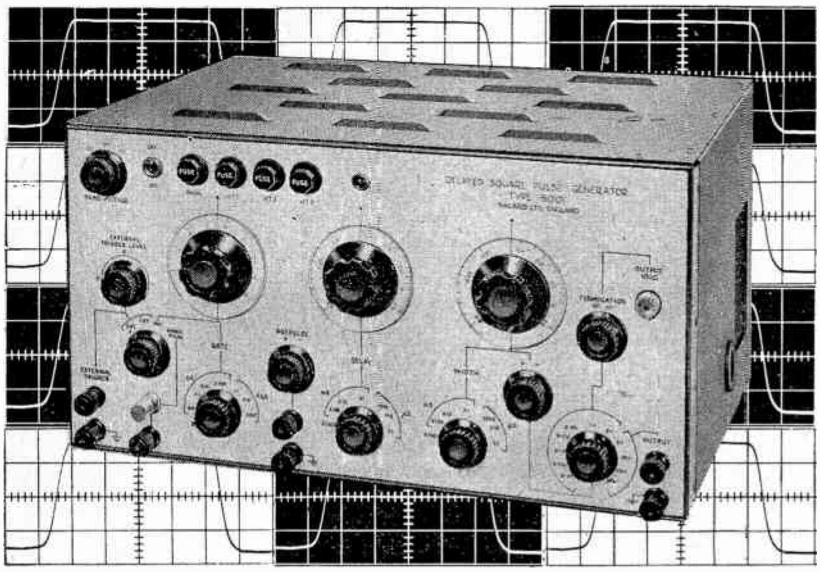

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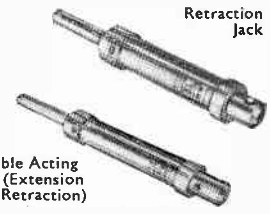

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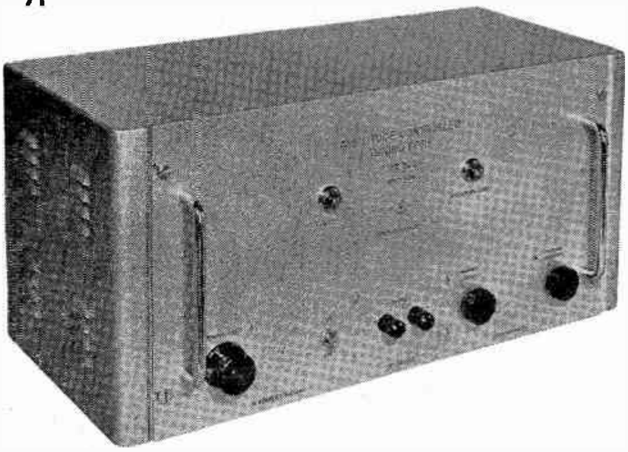

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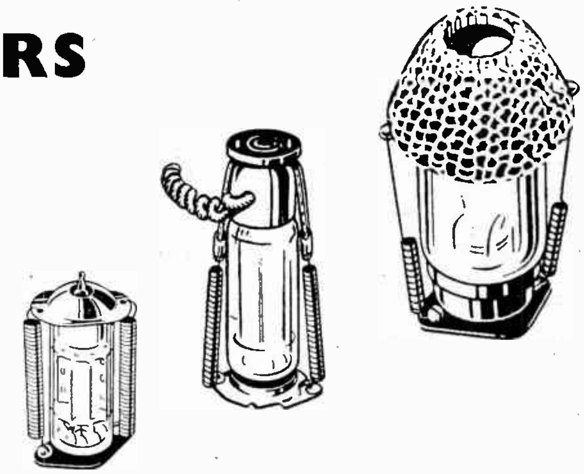

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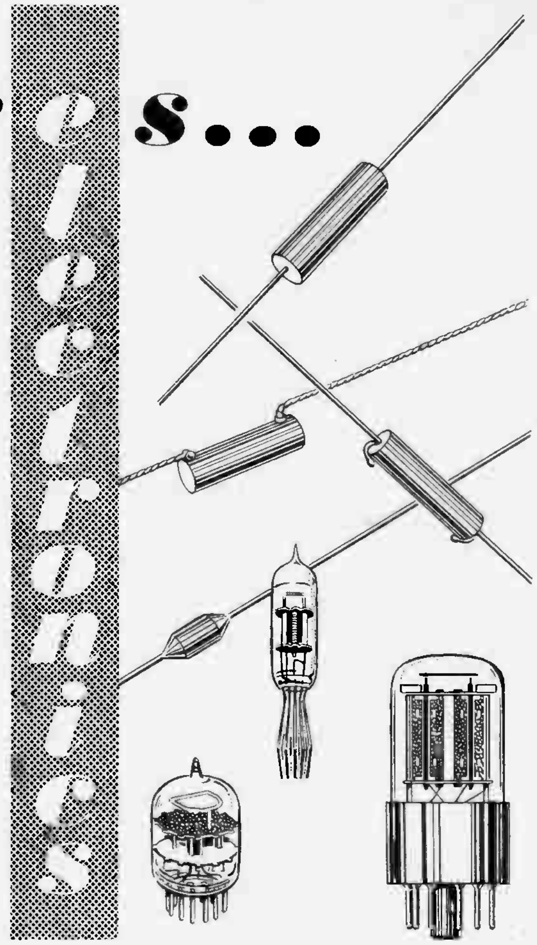

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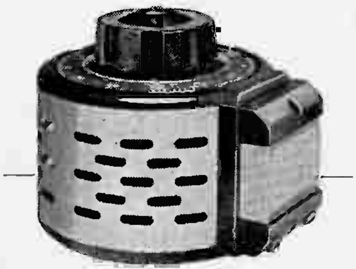

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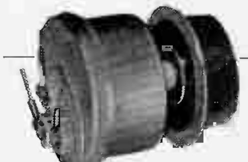

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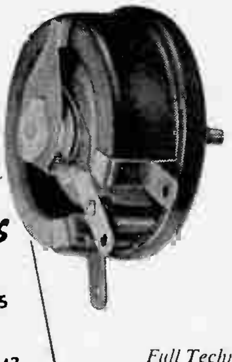

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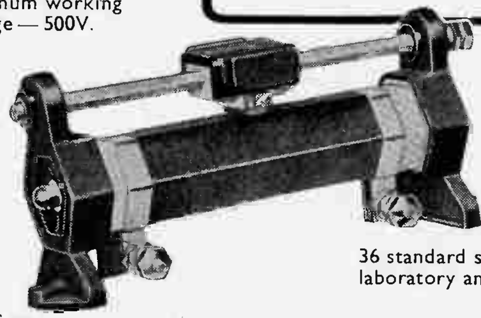

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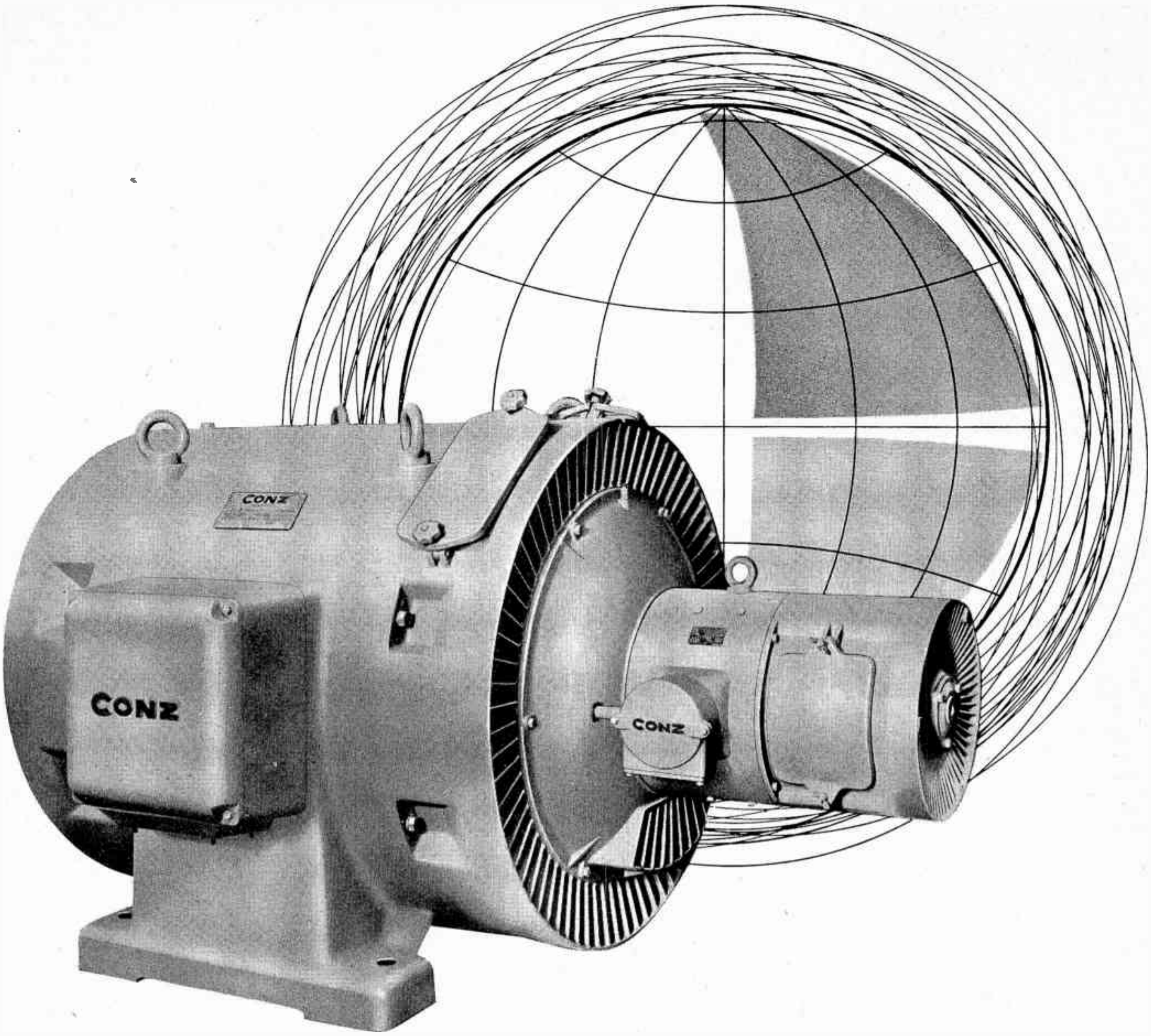

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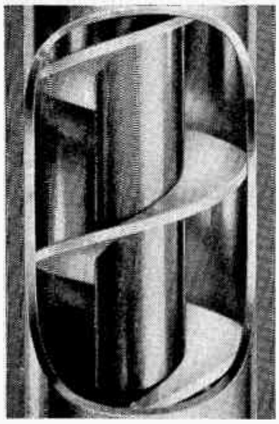

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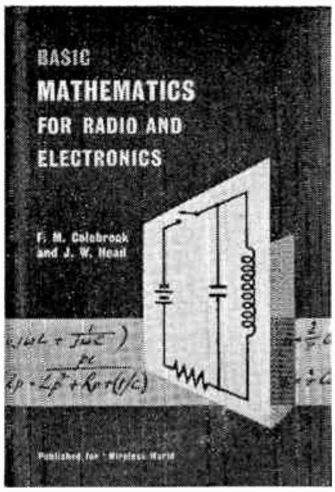

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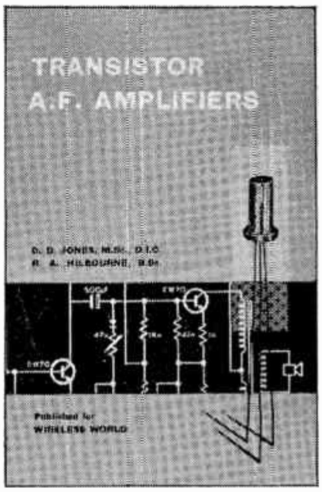

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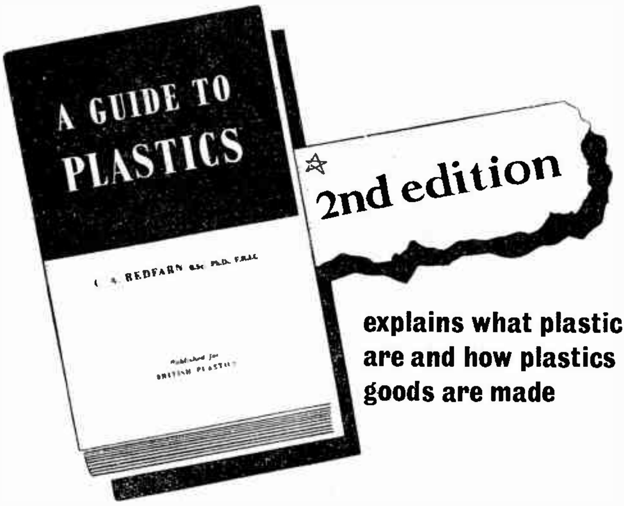

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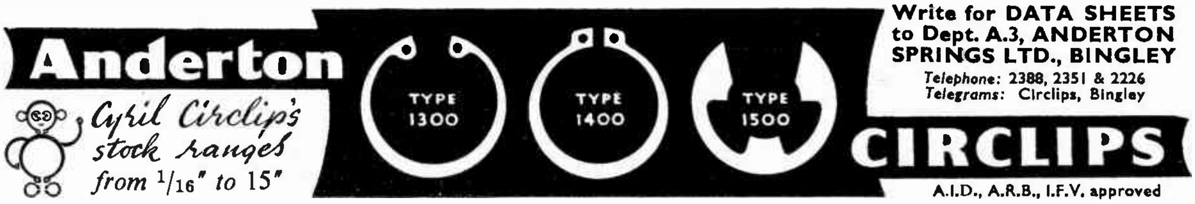
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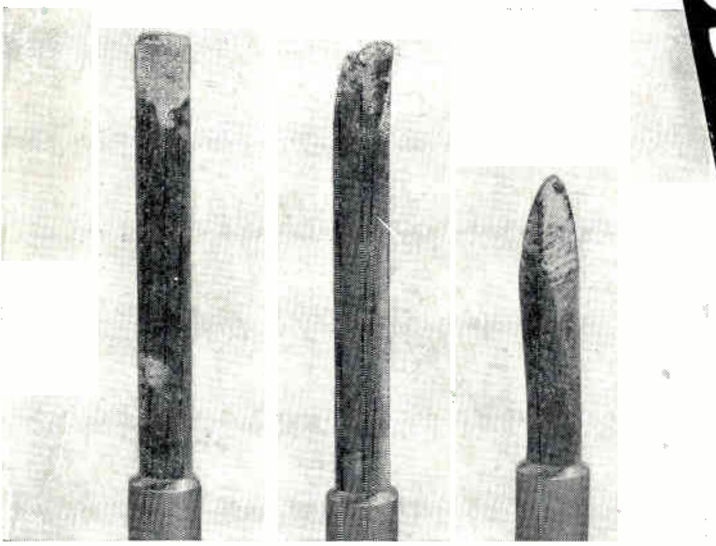

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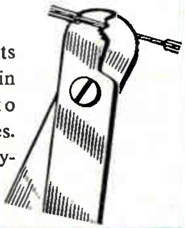

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