ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Thermistors

Bifilar-T Trap at Audio Frequencies

Nomogram for Air-Gap Design

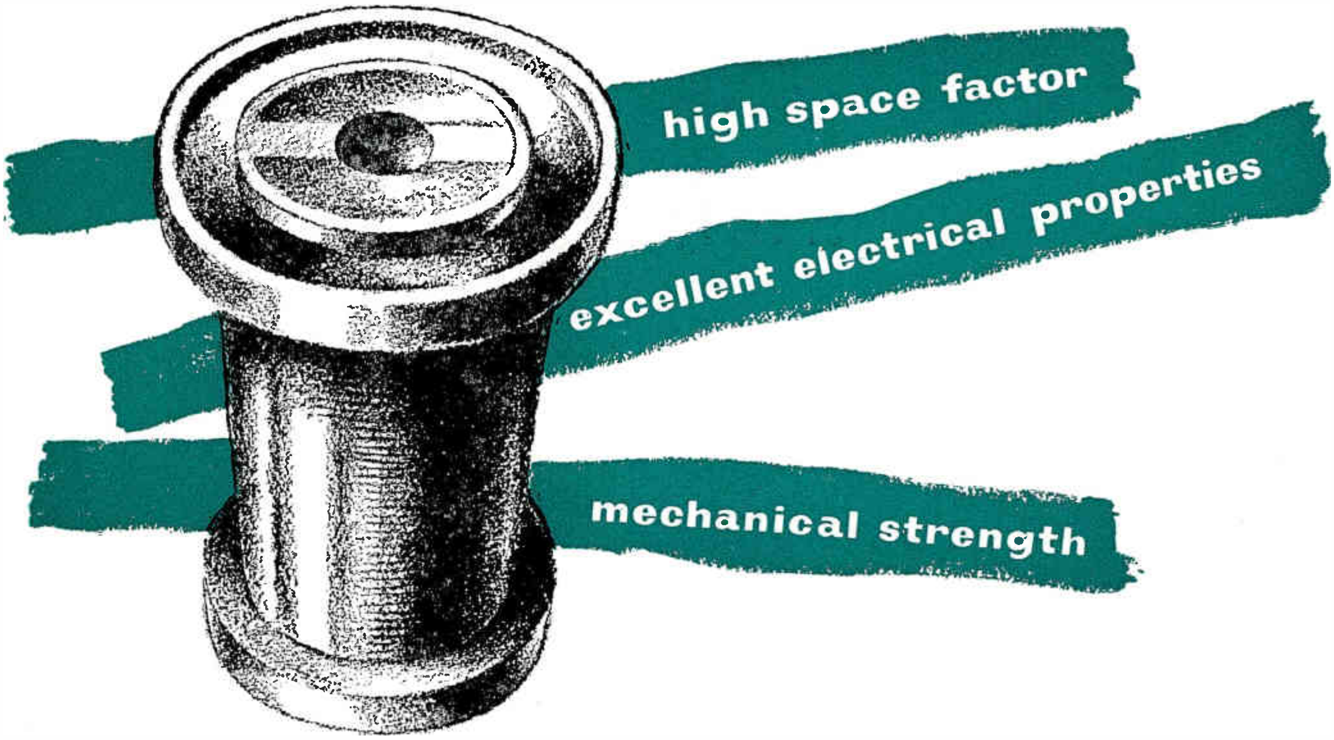
Transmission-Line Discontinuities

Three shillings
and sixpence

JULY 1958 Vol 35 *new series* No 7

3 BIG FACTORS

where small windings are concerned



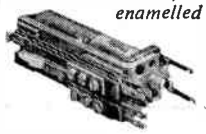
These are three good reasons why BICC Enamelled Winding Wires are ideal for small windings. In addition, coverings are flexible, strongly adhesive to conductors and resistant to heat and damp.

For full details, write for BICC Publication No. 303. We will be glad to send you a copy.

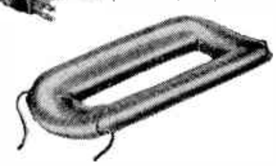
Experienced engineers of the BICC Technical Advisory Service are always ready to assist you in the selection of the right winding wire for your purpose.

1 *Coil for Amplivox miniature magnetic earphone wound with* ·001 *inch enamelled wire.*

2 *Telephone Relay Coil wound with* 6,900 *turns of* ·0036 *inch enamelled wire.*



3 *Field Coil for motor car dynamo wound with* 280 *turns of* ·030 *inch enamelled wire.*



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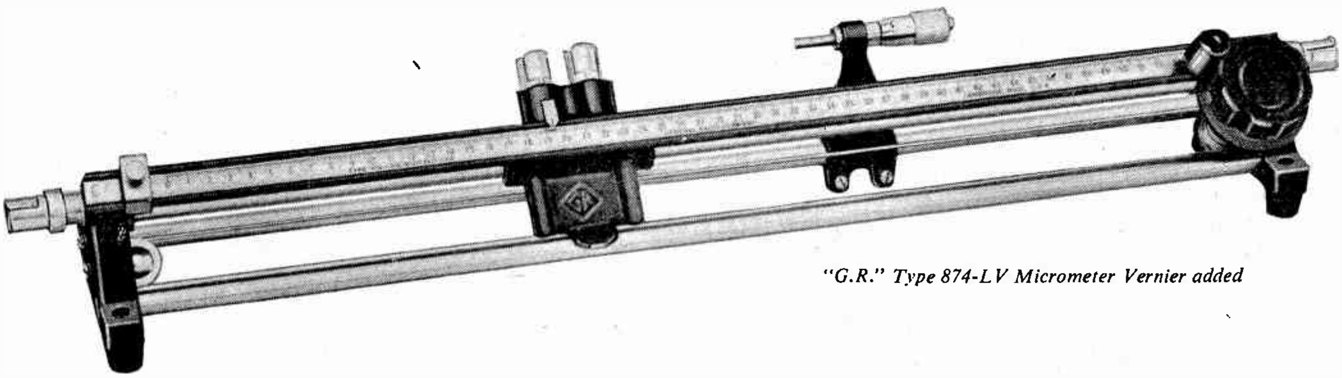
GENERAL RADIO COMPANY

Trade Mark

"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.

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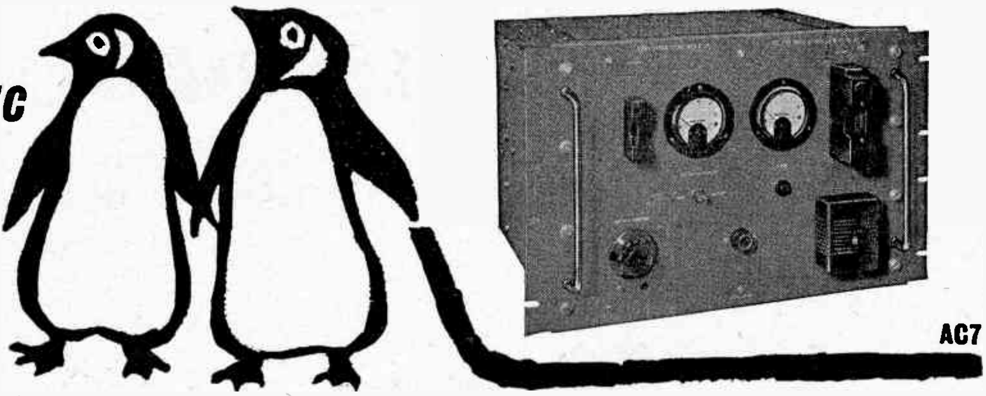
TO THE ANTARCTIC

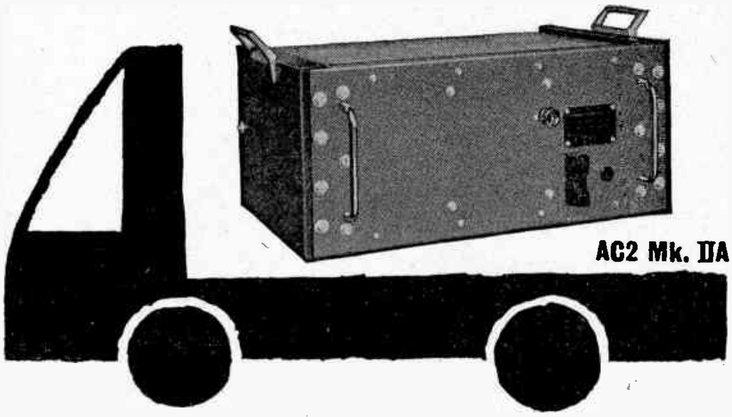


AC7

AC2 Mk. IIB

TO THE TROPICS

TO THE TEMPERATE ZONES



AC2 Mk. IIA

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standard A.C. stabilisers give A.1 service in all climates for ONE price

SERVOMEX a.c. voltage stabilisers are in use in the I.G.Y. programme in the Antarctic and in tropical Nigeria. These are in every way identical with the instruments in common use in this country. By extremely conservative design and by using components selected from the current inter-service approved list wherever possible, a very high degree of reliability is achieved. They will all withstand shock accelerations up to 40 g. These instruments introduce no distortion in the waveform whatever, and are not upset by changes of frequency, power factor, temperature, etc.

AC2 Mk. IIB and IIA

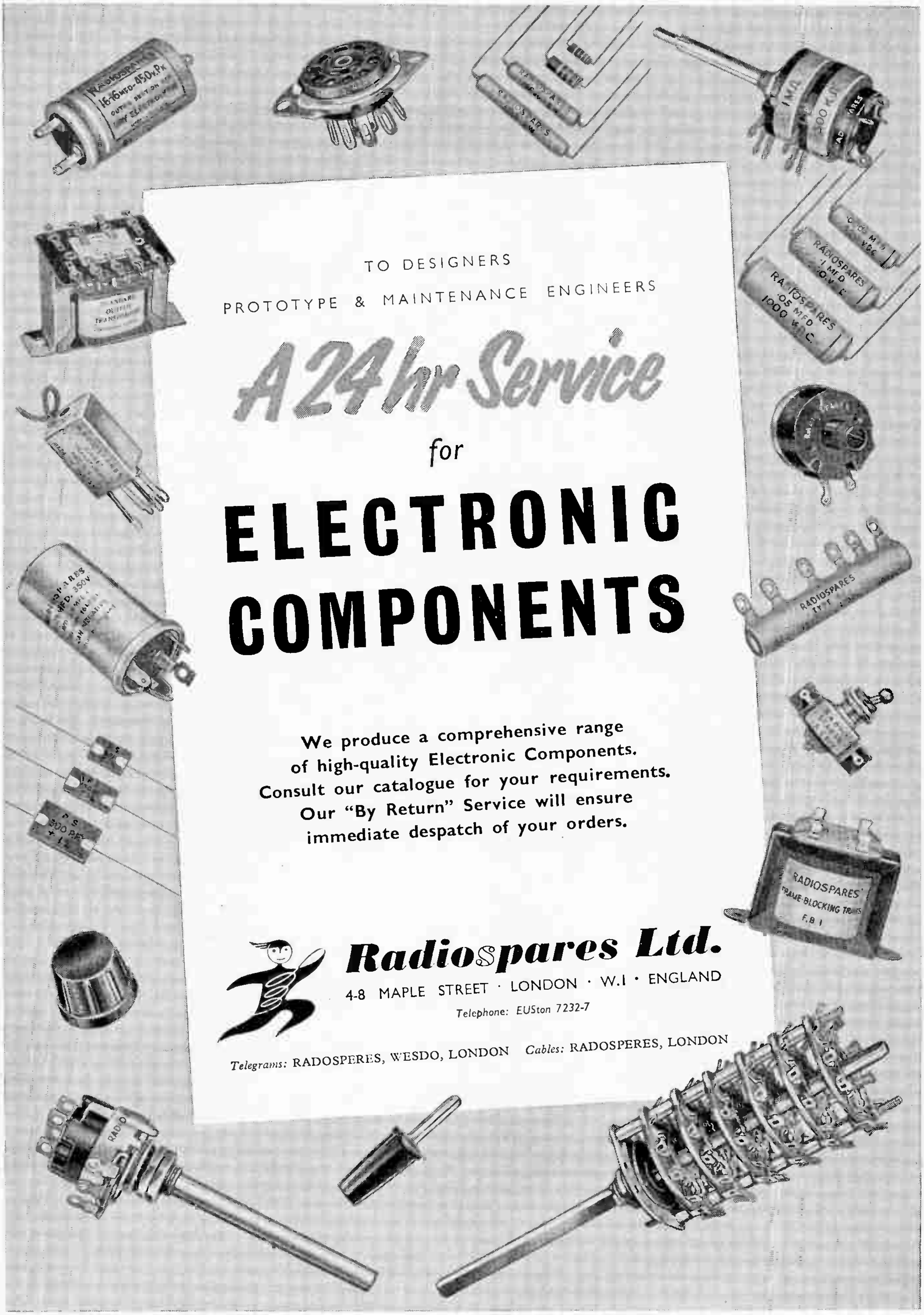
- 0 to 9 amps
- Range minus 17.5% to plus 8.75%
- 15 volts per second

AC7

- 0 to 30 amps
- Range minus 20% to plus 10%
- 12 volts per second

Technical data sheets are available on request

Servomex Controls Limited, Crowborough Hill, Jarvis Brook, Sussex. Crowborough 1247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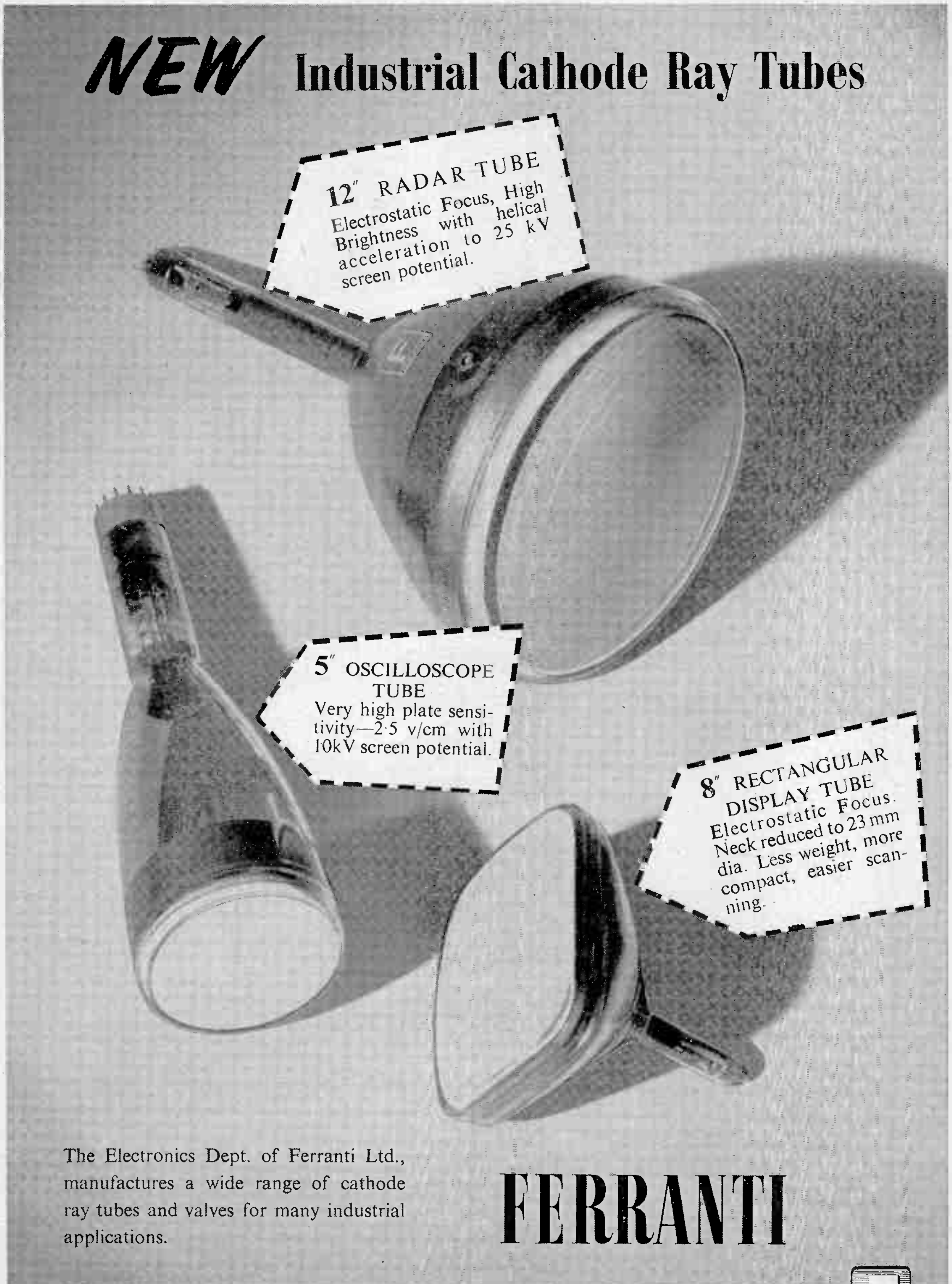

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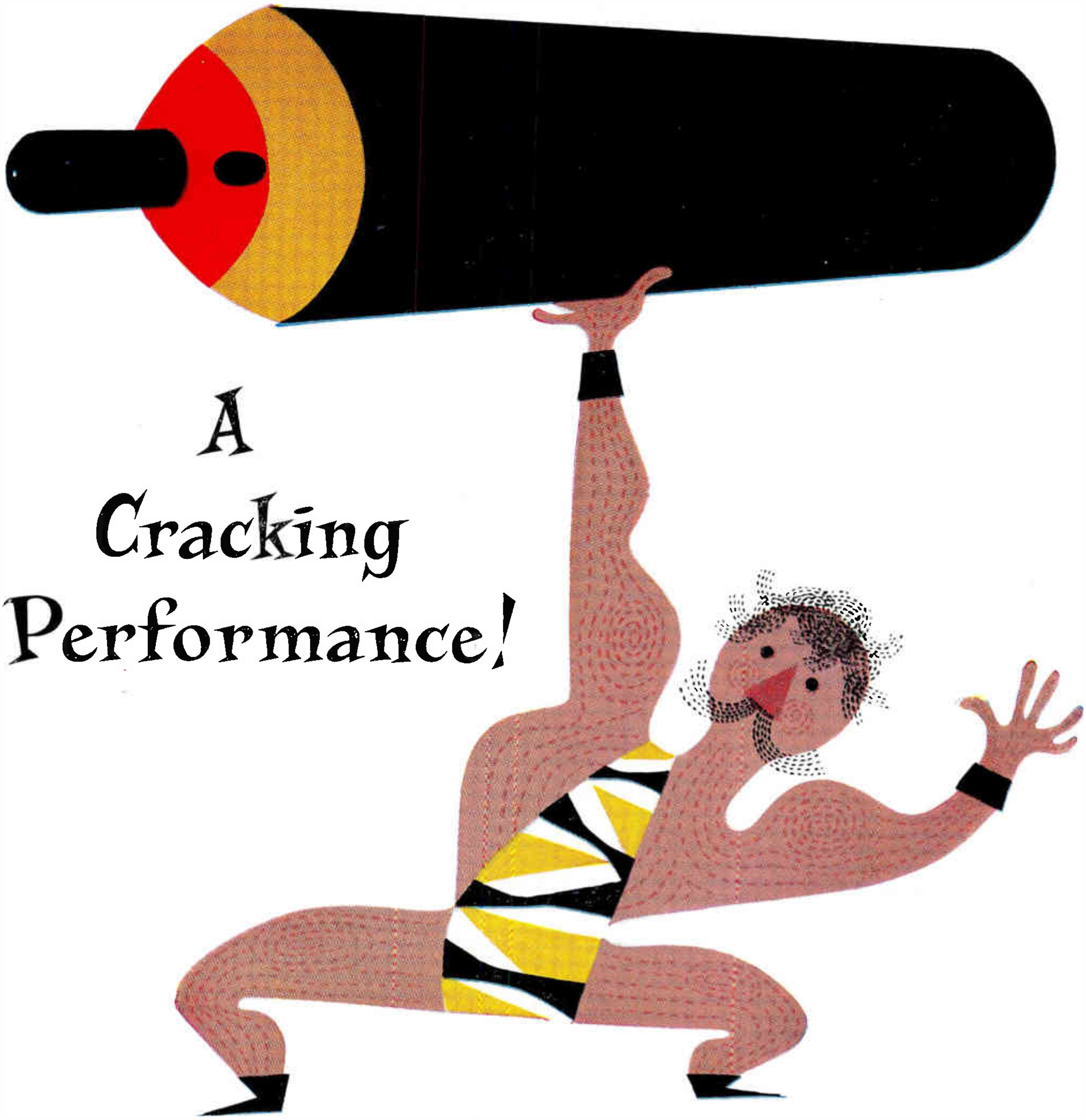
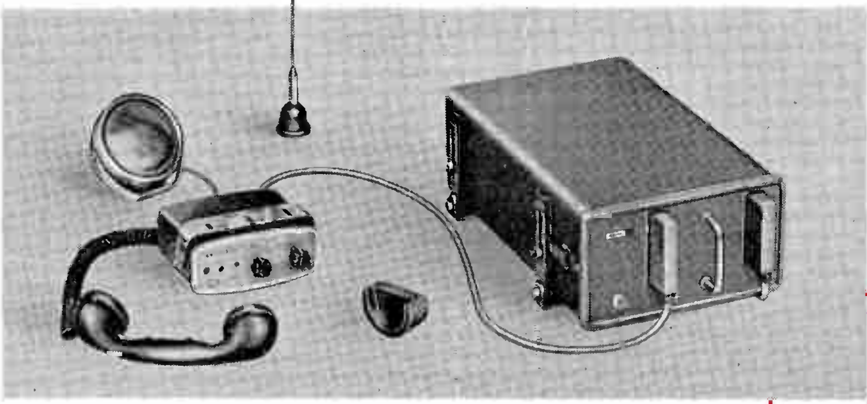
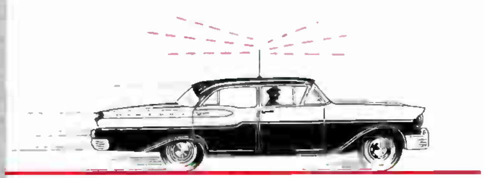


Your voice cuts through with the new

»Stornophone 33«

mobile FM/VHF radiotelephone

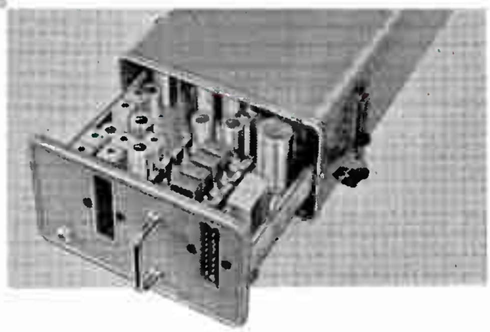




Built for the future, this equipment is a result of our experience from the manufacture of several thousand VHF sets for more than ten years.

"Stornophone 33" offers full 6/12 volts convertibility, extremely low power consumption, full tropicalization, splash- and dust-proof housing, small dimensions and six crystal-controlled channels.

Storno VHF sets are in daily operation in many countries from Northern Greenland in the Arctic to South Africa and South America in the tropics. Everywhere Storno sets are known for their perfect and reliable performance. Storno is represented in thirty countries all over the world. Kindly apply for further information.



- Directly interchangeable between 6 and 12 volt D.C. with any polarity
- Exceeds the American RETMA specifications
- Suitable for use in any climate
- Completely splash- and dust-proof housing
- Single unit drawer construction. Small dimensions: 6 × 11 × 18 in.
- 15 watt RF output.
- Dynamic microphone with transistor pre-amplifier
- Available with up to six crystal-controlled channels
- Extremely high stability
- Low power consumption: at 6 volts: stand-by 5.5 amps, transmit 17 amps.
- One full watt AF output delivered to a 5 in. speaker of high efficiency.

Storno

Manufacturers of Radio Communication Equipment
Div. of The Great Northern Telegraph Co. Ltd.
(*Established* 1869)

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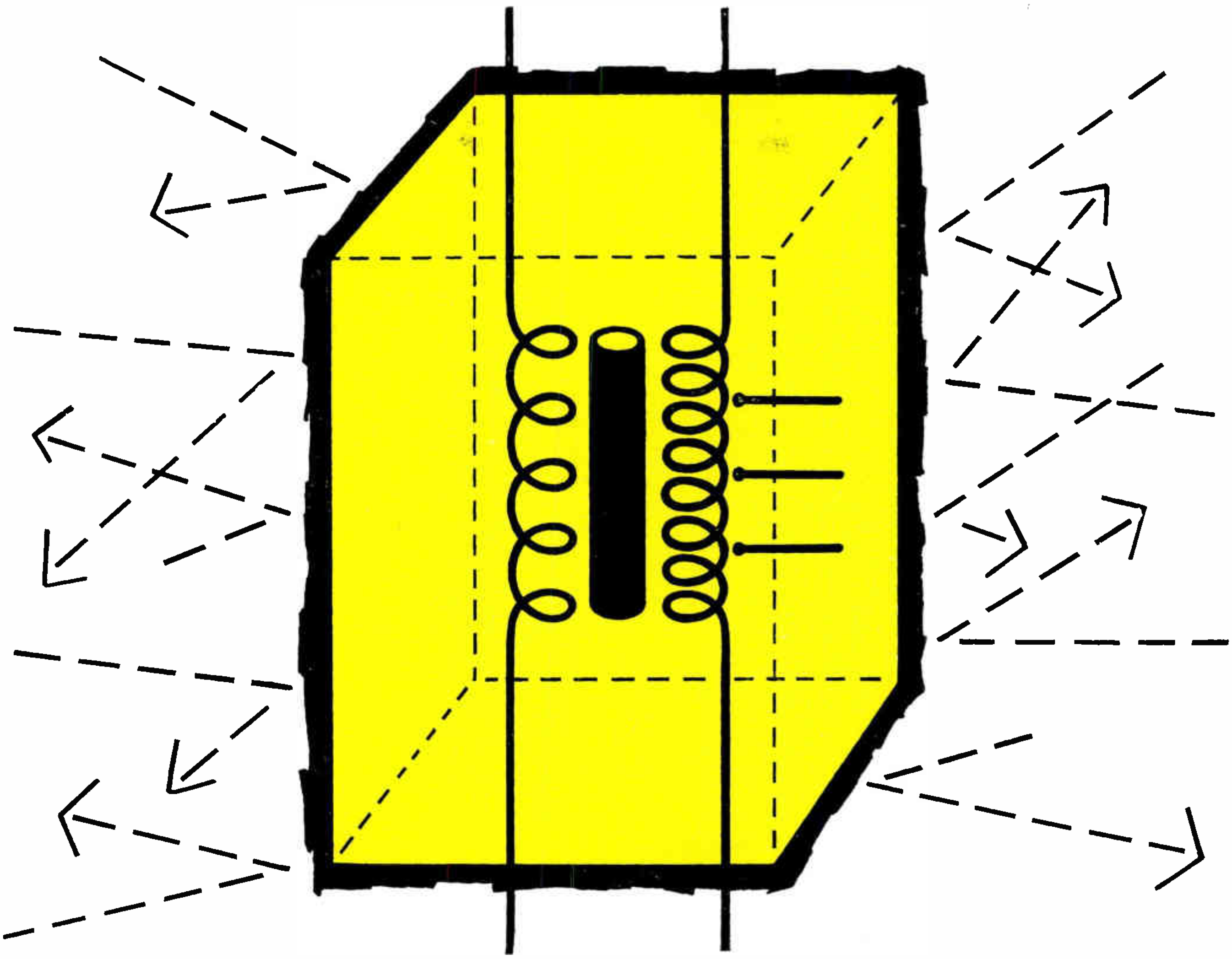


PARMEKO

ATLANTIC SERIES

STANDARD SMOOTHING CHOKES

Catalogue Number	D.C. Current Milliamps	Inductance Henries	Approx. D.C. Resis. Ohms.	Model Size
P-2772	10	50	2000	9000/39
P-2773	25	10	520	9000/39
P-2774	25	25	1500	9000/39
P-2775	25	50	1200	9000/41
P-2776	50	10	700	9000/39
P-2777	50	20	850	9000/40
P-2778	50	50	1200	9000/49
P-2779	75	10	230	9000/41
P-2780	75	15	450	9000/41
P-2781	75	20	500	9000/49
P-2782	100	10	290	9000/41
P-2783	120	10	240	9000/49
P-2784	120	15	280	9000/49
P-2785	120	20	300	9000/57
P-2786	150	1·5	115	9000/41
P-2787	180	5	140	9000/49
P-2788	180	10	190	9000/57
P-2789	180	15	260	9000/57
P-2790	180	20	280	9000/65
P-2791	250	1·5	45	9000/41
P-2792	250	2·5	70	9000/49
P-2793	250	5	120	9000/49
P-2794	250	10	140	9000/65
P-2795	250	20	180	9000/73
P-2796	350	10	100	9000/73
P-2797	500	2·5	32	9000/65
P-2798	500	5	50	9000/73

[illegible]

Standard Mains and Filament Transformers (see tables below) have all primaries wound 10-0-200-220-240V. 50 c/s; an electrostatic shield is fitted between primary and secondary windings on all models; the DC rectified current is quoted for full wave rectifier windings with condenser input filter.

STANDARD MAINS TRANSFORMERS

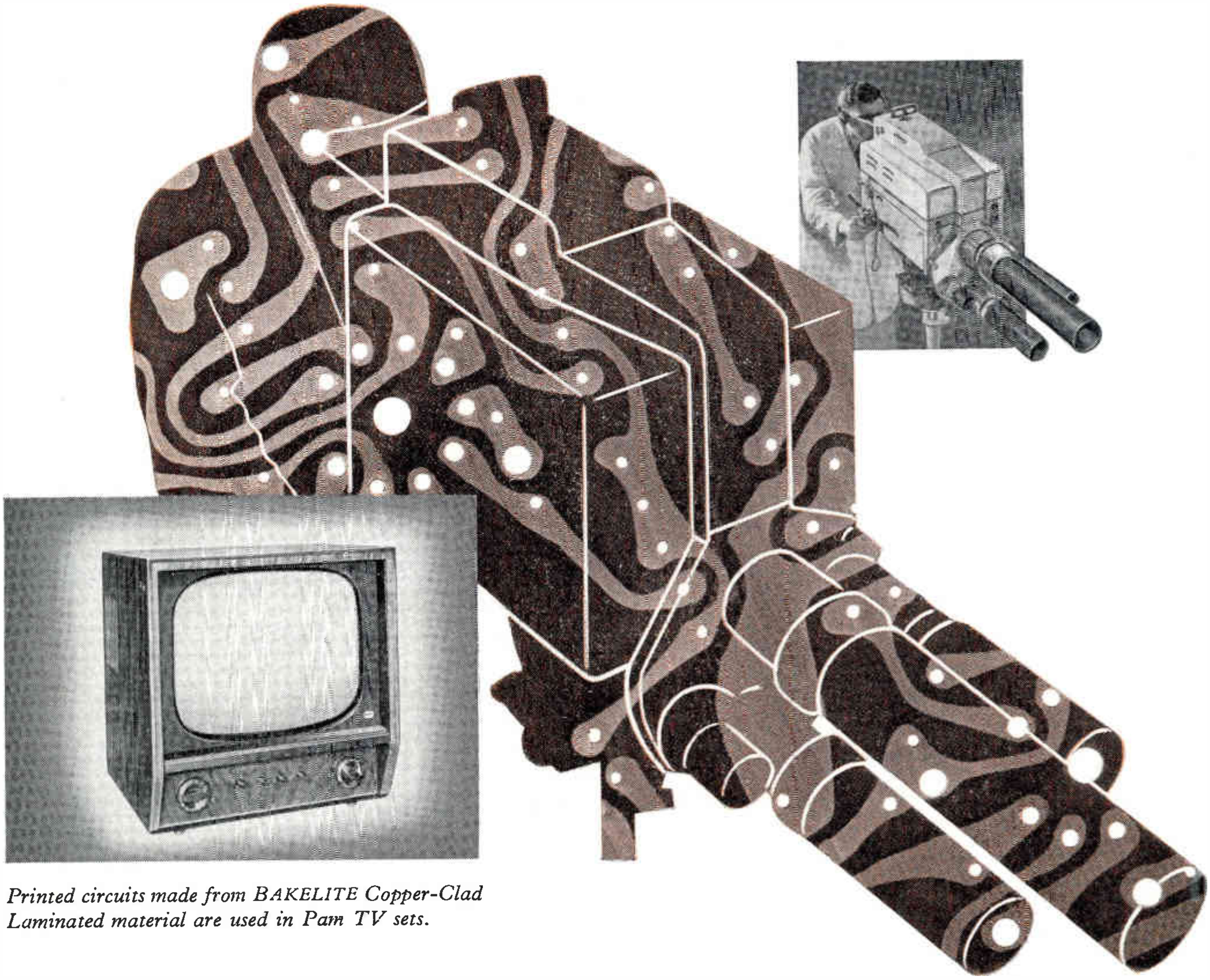
Ref. No.	ELECTRICAL SPECIFICATION						Model Size
	Secondary 1		Sec. 2	Sec. 3	Sec. 4	Sec. 5	
P-2749	150–0–150 V.	25 mA	6·3 V. 0·6 A.	6·3 V. 0·6 A.	—	—	9000/41
P-2750	150–0–150 V	50 mA	6·3 V. 0·6 A.	6·3 V. 1 A.	—	—	9000/49
P-2751	150–0–150 V.	75 mA	6·3 V. 1 A.	6·3 V. 2 A.	6·3 V. 1 A.	—	9000/57
P-2752	200–0–200 V.	75 mA	6·3 V. 1 A.	6·3 V. 2 A.	6·3 V. 1 A.	—	9000/57
P-2753	250–0–250 V.	75 mA	0/5/6·3 V. 2·5 A.	6·3 V. 3 A.	6·3 V. 2 A.	—	9000/65
P-2754	250–0–250 V.	100 mA	0/5/6·3 V. 2·5 A.	6·3 V. 3 A.	6·3 V. 2 A.	—	9000/65
P-2755	250–200–0–200–250	50 mA	6·3 V. 1 A.	6·3 V. 1·5 A.	—	—	9000/49
P-2756	300–250–0–250–300 V.	180 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 3 A.	—	9000/73
P-2757	350–300–0–300–350 V.	75 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 3 A.	—	9000/73
P-2758	350–300–0–300–350 V.	100 mA	0/5/6·3 V. 2 A.	6·3 V. 3 A.	6·3 V. 1 A.	—	9000/65
P-2759	350–300–0–300–350 V.	120 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 4 A.	—	9000/73
P-2760	350–300–0–300–350 V.	180 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 4 A.	—	9000/73
P-2761	350–300–0–300–350 V.	250 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 4 A.	6·3 V. 4 A.	9000/81
P-2762	450–400–0–400–450 V.	120 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 5 A.	—	9000/73
P-2763	450–400–0–400–450 V.	180 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 4 A.	9000/81
P-2764	450–400–0–400–450 V.	250 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 4 A.	6·3 V. 4 A.	9000/81
P-2765	550–500–0–500–550 V.	120 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 5 A.	—	9000/81
P-2766	550–500–0–500–550 V.	200 mA	0/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 5 A.	—	9000/81

STANDARD FILAMENT TRANSFORMERS

Ref. No.	ELECTRICAL SPECIFICATION						Model Size
	Secondary 1		Sec. 2	Sec. 3	Sec. 4	Sec. 5	
P-2767	6·3 V.	1 A.	—	—	—	—	9000/39
P-2768	0–4–5–6·3 V.	2 A.	—	—	—	—	9000/40
P-2769	0–4–5–6·3 V.	3 A.	—	—	—	—	9000/41
P-2770	0–4–5–6·3 V.	2 A.	0/4/5/6·3 V. 2 A.	—	—	—	9000/41
P-2771	0–4–5–6·3 V.	3 A.	0/4/5/6·3 V. 3 A.	6·3 V. 3 A.	6·3 V. 3 A.	—	9000/65

* DESIGN: Complies with BSS2214
* CONSTRUCTION: Steel encased, compound filled
* DIMENSIONS: Plan and fixing to RCL215
* HUMIDITY: Category H2 or better
* TERMINALS: Patented design insulators, layout to RCL215
* MOUNTING: Upright or inverted all models
* FINISH: Grey hammer, stoved enamel

PARMEKO LIMITED
PERCY ROAD . LEICESTER
England



no rungs missing in the ladder of our range

of

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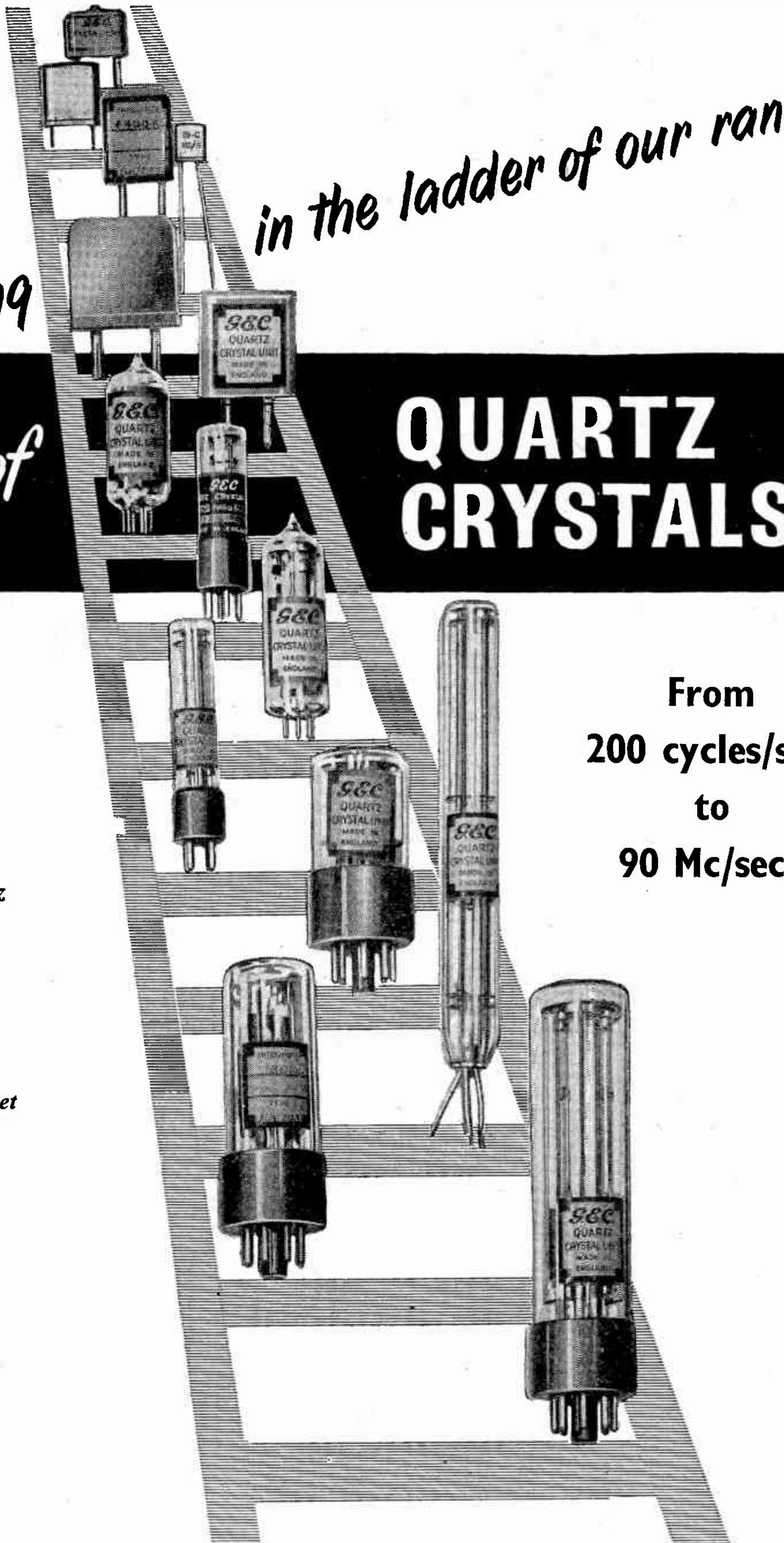

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For long term stability and unfailing activity, G.E.C. Quartz Crystal Units provide the basis for reliable communications systems.

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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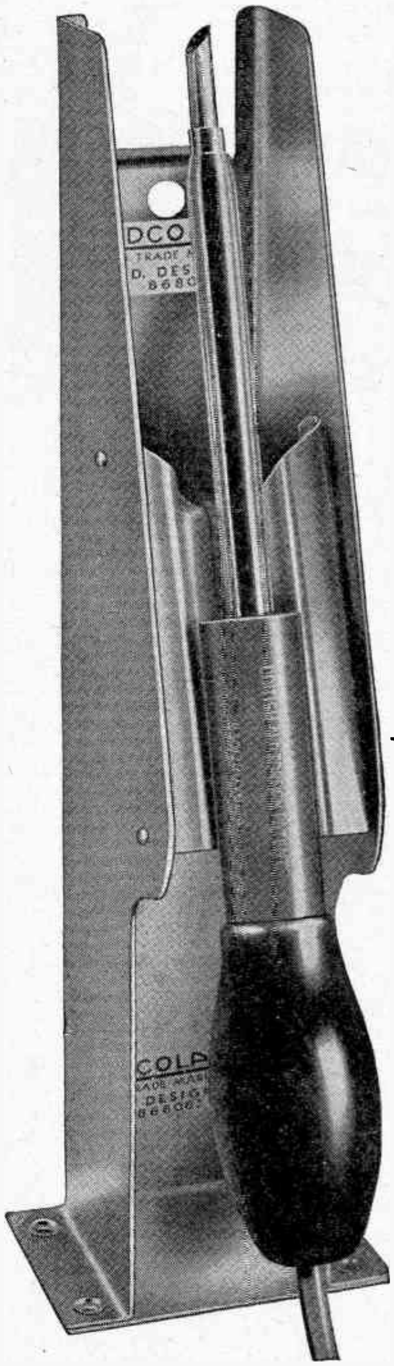

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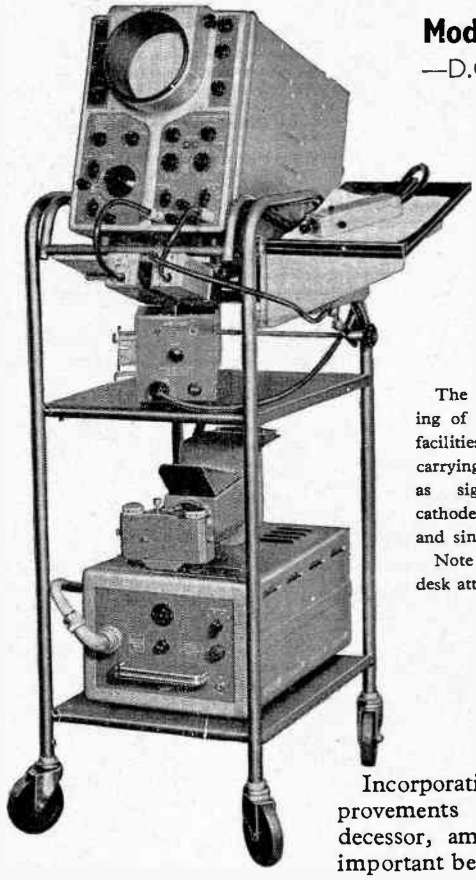
A new wide band

Two channel

OSCILLOSCOPE

Model DE 103 A

—D.C. to 10 Mc/s



The new trolley mounting of this instrument has facilities for conveniently carrying accessories, such as signal delay lines, cathode follower probe and single shot camera.

Note the useful writing desk attachment.

Incorporating many improvements over its predecessor, among the most important being:—

1. **The standard of accuracy with which the displays on the two channels can be compared with each other.**
2. **The linearity of the time base, making possible time measurements within 2% accuracy.**
3. **Internal synchronising facilities provided.**
4. **Greater accuracy of the voltage calibration.**
5. **Higher sensitivity without loss of band width.**

In its present form it provides unique facilities for the accurate display and measurement of two high frequency signals simultaneously, and is a well tried, reliable instrument.

Having two signal channels which are free from any interaction, and which do not depend on beam switching techniques, the two signals may be aligned for comparison in any desired way with confidence that their waveform details are strictly relative, under either single stroke, triggered, or repetitive sweep conditions.

In effect the design provides two first-class wide band oscilloscopes combined in one unit.

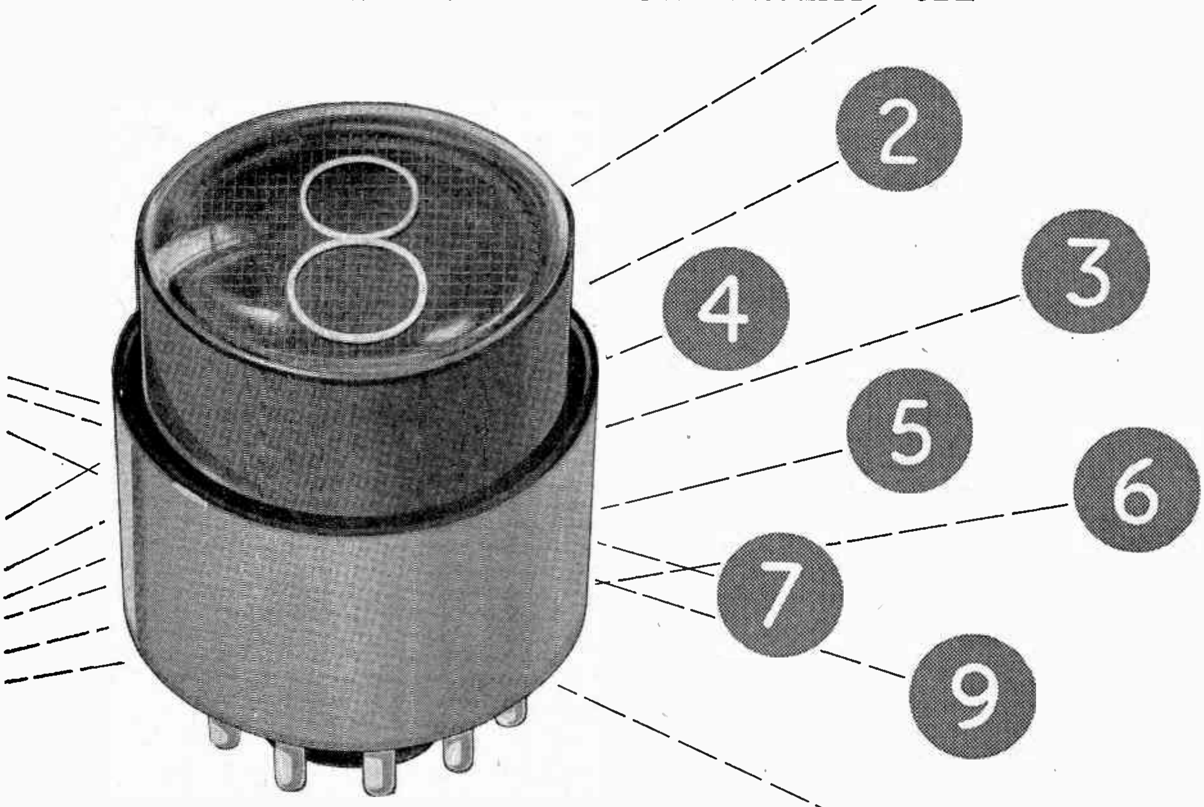
Ask for data sheet No. 5570 for full details.

NAGARD LTD

18 Avenue Road, Belmont, Surrey. VIGilant 9161/2

The Nodistron

A NEW NUMERICAL DISPLAY TUBE



G10/200E

This new Numerical Indicator tube is designed for use in any direct read out equipment. It is a cold cathode device, the cathodes being shaped in the form of numerals and brought out to separate pins on a duodecal base. The application of approximately 200 volts between any specified cathode and the common anode causes a glow discharge in the form of any individual numeral from 0 to 9.

Standard Telephones and Cables Limited

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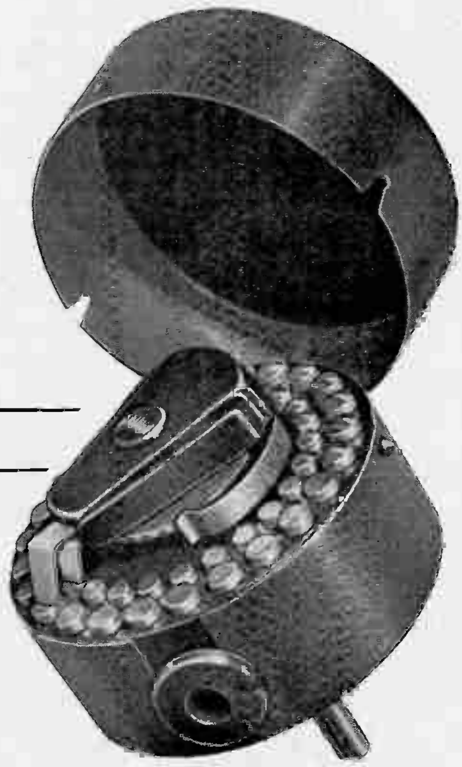


INTRODUCING THE NEW

SULLIVAN & GRIFFITHS

PRECISION BEAT TONE OSCILLATOR

TYPE 10A

0–16,000 c/s

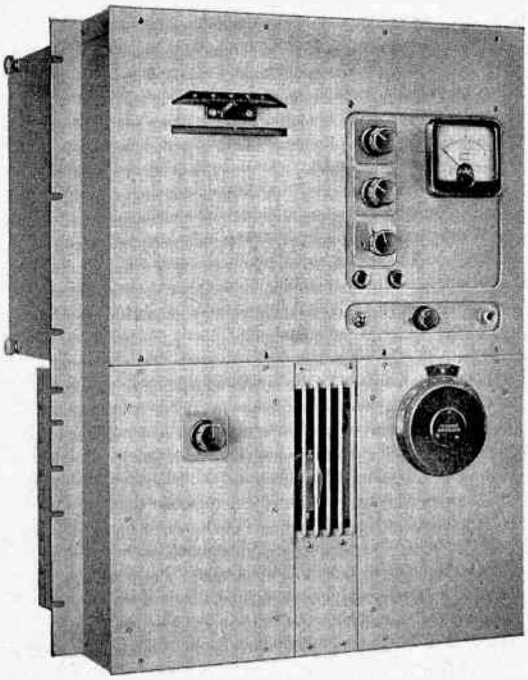
Frequency Accuracy 0.2% ± 0.5 c/s; considerably better below 100 c/s

MAINS OPERATED

A NEW VERSION OF THE WELL-PROVED **RYALL-SULLIVAN** OSCILLATOR IN WHICH THE BEST FEATURES OF THE ORIGINAL DESIGN HAVE BEEN RETAINED, VIZ. HIGH STABILITY OSCILLATORS, PERFECTION OF C.C.I. CALIBRATED SCALE, ETC.

In this redesigned oscillator the high-frequency circuits are contained in a single small unit. Similarly the output amplifier and power supply are contained in a second unit.

These units are easily removable for servicing and, furthermore, interchangeable units are readily available if required.



FREQUENCY RANGE 0–16,000 c/s

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

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

FREQUENCY INTERPOLATING SCALE

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

ALTERNATIVE OUTPUTS ARE PROVIDED:

OUTPUT NO. 1

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

OUTPUT NO. 2

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

OUTPUT LEVEL

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

HARMONIC CONTENT

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

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

MAINS HUM (50 c/s and 100 c/s)	– 70 dB for all output voltages above 4 V. 700 μV with output controls at zero.
NOISE	– 70 dB for all output voltages above 4 V. 700 μV with output controls at zero.
HIGH FREQUENCY (leakage from oscillators) ...	– 70 dB for all output voltages above 4 V. 10 μV with output controls at zero.

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incorporating WIRELESS ENGINEER

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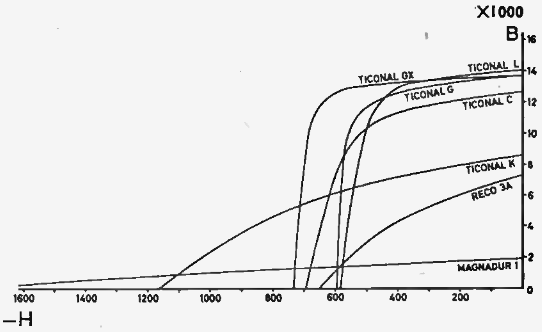
The Choice of Magnet Materials

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

The Mullard range of permanent magnet materials meets the majority of present day magnet requirements. To assist designers in choosing the correct magnet material for a given purpose, some of the advantages of these materials and a few of their applications, are given below:

'Ticonal' GX has the highest energy content of the commercially available permanent magnet materials. It is difficult and expensive to manufacture, and should only be used where minimum size and weight are extremely important with cost a



secondary factor. These magnets are predominantly anisotropic and are only manufactured in block and cylindrical form, with a straight magnetic axis.

'Ticonal' G represents the best compromise between performance and cost of any permanent magnet material available. It is comparatively easy to manufacture into any required shape and gives a high and uniform performance. It is strongly recommended for general purposes and is used throughout the world for the majority of permanent magnet applications.

'Ticonal' L is primarily intended for use in loudspeaker assemblies. A permanent magnet of higher B_d value than is obtainable from 'Ticonal' G has certain technical advantages.

'Ticonal' C has slightly lower magnetic properties than 'Ticonal' G but with a higher coercive force; this material is particularly useful in motors, dynamos and other dynamic applications.

'Ticonal' K is a relatively new material and only certain shapes are available at present. Its magnetic characteristics facilitate the use of very short magnets. It has an exceptionally high coercive force and is recommended in the design of moving coil instrument circuits where a magnet forms the inner core of the system.

'Reco' 3A is isotropic (non-directional) and can therefore be magnetised equally in any direction. It is of comparatively low performance but is particularly suitable for multi-polar applications or inexpensive magnets where high performance is not essential.

'Magnadur' 1 is one of the new ceramic magnets. As it contains no cobalt or nickel it is therefore inexpensive to manufacture. It has an unusually high coercive force and is particularly useful for applications where the magnet is subjected to alternating fields of any frequency.

Principal Characteristics of Mullard Permanent Magnets

MAGNET	BH (max) x 10^6	B_r	H_c	B_d	H_d
'Ticonal' GX	7.5	13,500	720	12,000	625
'Ticonal' G	5.7	13,480	583	11,000	520
'Ticonal' L	5.4	13,500	575	12,000	450
'Ticonal' C	5.0	12,500	680	9,620	520
'Ticonal' K	4.0	9,000	1,300	5,000	800
'Reco' 3A	1.7	7,200	645	4,350	390
'Magnadur' I	0.95	2,000	1,750	950	1,000

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

Mullard

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

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

MC265

VOLUME 35 NUMBER 7 JULY 1958

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Technical Authorship

THIS is the title of a pamphlet issued by the City and Guilds of London Institute which contains the syllabus of a new course. This course is in two sections. The first deals with various kinds of technical literature and publications, technical illustration, methods of reproduction, and similar matters. The second covers the English language, grammar, words, the collection of information and its classification, preparation of lay-outs, proof-reading and writing.

The course is of 600 hours, of which it is recommended that not more than 125 hours should be spent on the first section. Further, it is considered that no more than 20% of the time allocated to the second section should be spent on lectures. It is envisaged that the bulk of the time will be spent on exercises in technical writing.

This is, of course, as it should be, for one can only learn how to write by writing. It does, however, throw a great responsibility upon the teachers, for the student will benefit from writing only if his work receives detailed constructive criticism from his teacher. Proper criticism is quite difficult and very time consuming. There is, too, the difficulty of achieving a uniform standard of criticism among a number of teachers. But this is a difficulty common to all arts.

The syllabus strikes us as a very good one and we particularly welcome the stress placed on writing. We do consider, however, that the title of the course is a misleading one. To us and a great many other people a technical author is one who writes a technical book or an article for a journal such as this.

The course is clearly intended, however, for people who will be employed by manufacturing firms to prepare handbooks, sales literature and the like. In industry such people are usually called technical writers. Although we do not think this is an ideal name, it is a good deal nearer the truth than 'technical author'.

Thermistors

A REVIEW OF THEIR PROPERTIES AND APPLICATIONS

By K. R. Patrick, A.M.I.E.E.*

The term 'thermistor' has been coined from the words 'thermal resistor' to describe a class of resistor having a high temperature coefficient. In practice, the thermistors available today all have negative temperature coefficients which, at room temperatures at any rate, are some ten times higher than that of copper. Unfortunately the temperature coefficient is not constant, as in the case of metallic conductors, but varies with temperature. This is dealt with in more detail below.

By the exercise of a little ingenuity and by taking advantage of the widely differing forms in which they are available, thermistors can be used in a varied and increasing number of applications, many of which have been the subject of articles in the technical press during the last few years. The purpose of this review is to bring together in a compact and, it is hoped, easily digestible form notes on types available in this country and their mode of operation, together with some typical applications. The method of manufacture and the physical theory of their operation are also briefly mentioned; finally, a list of references is appended which may be of use to those who wish to pursue any particular application in more detail.

* *Mullard, Limited.*

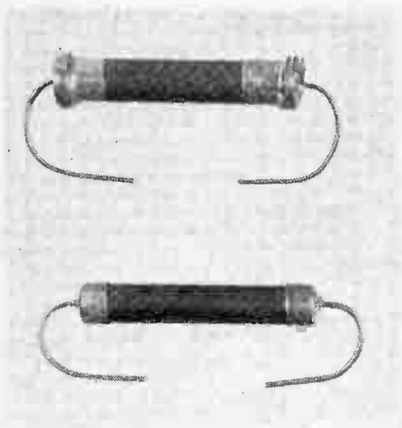
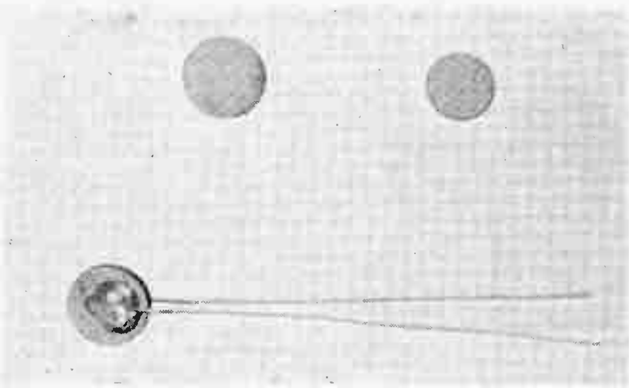


Fig. 1. *Typical thermistors: (left) rod-types, showing soldered wire and pressed-cap methods of making connection; (below) disc-types with silvered faces*



Thermistors are sometimes referred to as non-linear resistors but, as we shall see, the conduction in a thermistor is purely electronic and hence they may be used with both direct and alternating currents. The current which flows through the thermistor is proportional to the e.m.f. applied across it, provided that the temperature is kept constant, and to this extent a thermistor is a linear resistor. Of course, if the current is sufficiently large and is allowed to flow long enough to raise the temperature of the thermistor so that its resistance changes, then the r.m.s. voltage/current ratio changes also; but after all, this is true of all resistors. Certainly thermistors have no voltage coefficient of resistance as such, and must therefore be distinguished from voltage-dependent resistors in which the non-linear voltage/current relationship is not dependent on the effects of temperature.

One curious result of the change of resistance with current flow is that under favourable conditions and at very low frequencies a thermistor will exhibit the characteristics of a reactance[1].

Manufacture of Thermistors

Thermistors are manufactured in three standard forms:

(a) wire-terminated rods resembling composition-type resistors in appearance (Fig. 1),

(b) discs or slabs, the flat surfaces of which are metal-coated, usually with silver (Fig. 1), and

(c) miniature types based on a 'bead' about half a millimetre in diameter which is then usually supported in a glass mount. There are a number of these, each designed for a special application (Figs. 2 and 3).

Thermistors are made from a mixture of various semiconducting oxides which are sintered together to become complex uniphase materials of the spinel type. By varying the choice and proportion of oxides used, control is obtained over the resistivity of the resultant mass.

The oxide mixture is made up to a paste which may

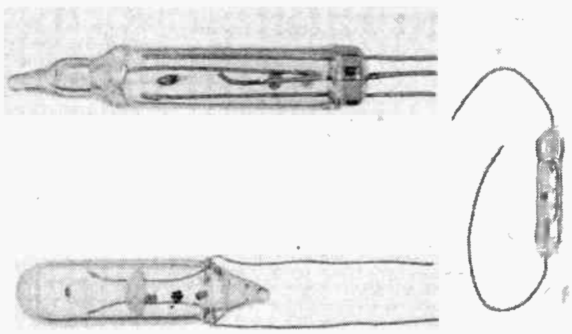


Fig. 2. *Bead thermistors in glass mounts. Top left: indirectly-heated type (see also Fig. 12); bottom left: in evacuated envelope to reduce time constant to a minimum; right: for self-heating applications.*

then be extruded in the form of rods or pressed into discs. These are then fired in kilns at temperatures in the region of 1100°C. During the firing process considerable shrinkage occurs (which may result in distortion) and a final grinding operation therefore frequently follows.

The next stage in the making of rod and disc types is to attach a connecting wire or other means of making an electrical connection. In the case of rod types the ends may be coated with graphite and a metal cap (to which a connecting wire is welded) forced over them. However, a preferable method is to spray the end of the thermistor with molten metal (copper or zinc) and then solder a pigtail of tinned-copper wire to this surface. Two typical rod thermistors are shown in Fig. 1.

Disc thermistors (also illustrated in Fig. 1) are primarily intended for temperature measurement, and thus with these it is common practice to bond a layer of silver to the flat surfaces using one of the commercially-available silver pastes in a low-temperature firing technique. Wires may then be soldered to the surfaces or the entire disc may be soldered to the metalwork of the equipment whose temperature it is desired to measure. When soldering to these silvered surfaces a silver-bearing solder should be used*, although 60/40 tin-lead solder may be employed if care is taken to see that the solder does not remain molten for a sufficient time for the silver surface to be dissolved.

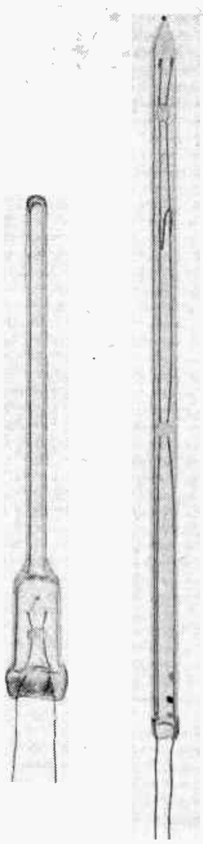


Fig. 3. *Special glass mounts for bead thermistors. Left: for gas analysis and vacuum measurement; right: probe for temperature measurement, the bead being sealed in the tip at the top*

The procedure for making the 'bead'-type thermistor is to place drops of the oxide paste on two parallel platinum alloy wires of some 0·002 in. diameter [Fig. 4 (a)]. This is then fired and given a thin coat of vitreous enamel, after which alternate wires are nipped off [Fig. 4 (b)]. These wires are then welded to heavier connecting wires and the whole element may then be mounted as required, a variety of such mountings being shown in Figs. 2 and 3. More will be said later about their suitability for various applications.

Theory of Operation

The reader will be familiar with the idea that conduction in an electrical conductor (such as a copper wire) is by means of free electrons which are comparatively loosely attached to the molecules of which the conductor is composed. These molecules are in a constant state of agitation, moving rapidly about in a random fashion and colliding with one another. The free electrons have the ability to detach themselves from the one molecule and move to another so that there is a constant interchange of electrons going on all the time. The speed at which the molecules move about and, consequently, the rate at which collisions take place is proportional to the temperature of the conductor, but the number of electrons available for conduction is independent of temperature. When an electric field is applied across the ends of the wire the electrons commence to migrate in the general direction of the field and an electric current is said to flow. The effect of increasing temperature is to

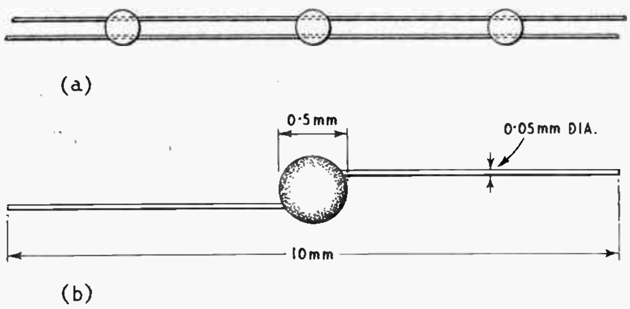


Fig. 4. *Diagram illustrating the manufacture of bead-type thermistor elements. (a) Drops of ceramic paste on wires; (b) alternate wires removed*

reduce the mobility of the electrons *in any particular direction* due to the increasing rate at which collisions between molecules take place. Thus the resistance of the conductor *increases* with temperature in an approximately linear fashion leading to the familiar relationship:

$$R = R_0(1 + \alpha T)$$

We define α as the temperature coefficient and

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

In the case of the semiconductor the number of

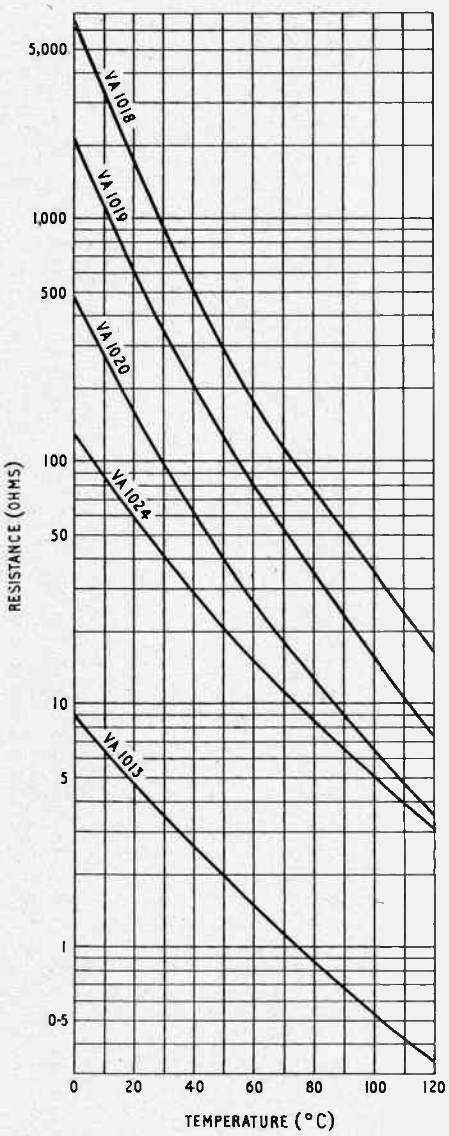
* e.g., Fry's Metal Foundries '290 Alloy'.

electrons available for conduction is no longer a constant but increases exponentially with temperature, owing, as it were, to the molecules 'unlocking' as their thermal energy increases. Conversely, as the temperature decreases the molecules 'lock-up' again and the electron supply diminishes. This effect far outweighs the decrease in electron mobility and, consequently, a thermistor exhibits a large non-linear *negative*-temperature characteristic which may be approximately expressed as:

$$R = A \exp. (B/T)$$

where A is a constant largely dependent on the shape of the thermistor, B is a constant determined by its material, and T is expressed in degrees absolute. There is, however, a temperature below which a thermistor ceases to adhere to this law, the normal working range of the materials commonly encountered being within $-100°C$ to $+300°C$. (Soldered connections and other similar considerations will, of course, impose restrictions within this range.) A recent American article[2] has described thermistors for operating at very low temperatures, but these are not commercially available in this country.

Fig. 5. *Family of resistance/temperature curves for some typical disc thermistors.* (*Negligible current flowing*)



Thermistor Characteristics

Basic Equations

The relationship mentioned above reduces to:

$$R_1 = R_2 \exp. \left(\frac{B}{T_1} - \frac{B}{T_2}\right)$$

where R_1 is the resistance at an absolute temp. T_1 (°K)
R_2 is the resistance at an absolute temp. T_2 (°K)
B is a constant for a given thermistor.

A more useful form of this equation is:

$$\log_{10}R_1 = \log_{10}R_2 + \frac{B}{2{\cdot}303}\left(\frac{T_2 - T_1}{T_1 T_2}\right) \qquad (1)$$

The temperature coefficient is given by:

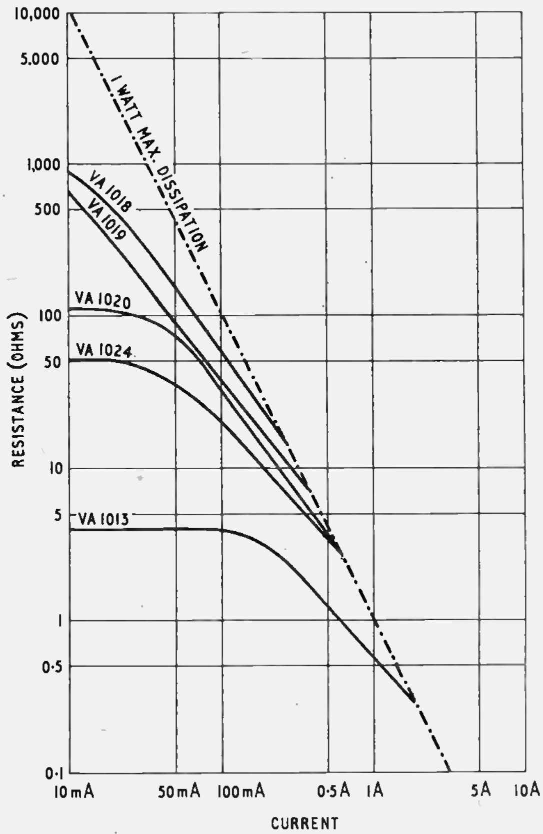
$$\alpha = -\frac{B}{T^2} \qquad (2)$$

The value of B usually lies between 2000°K and 5000°K and thus it will be seen that α is negative and it usually lies between about $2\frac{1}{4}\%$ and 6% per °C at 25°C, falling rapidly with increasing temperature. (In general, as between different thermistor materials and configurations, B increases with resistivity and hence the higher the 'cold' resistance, the greater B will be.)

Thermistor Data

In order to design thermistor circuits two characteristic curves are required; that relating temperature to

Fig. 6. *Family of resistance/current curves for the same disc thermistors as in Fig.* 5. (*Ambient temperature,* 25°*C*)



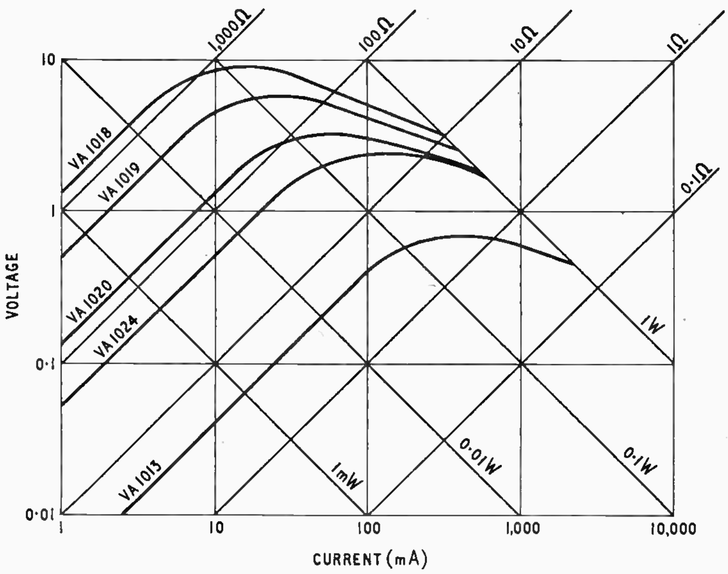


Fig. 7. *Family of voltage/current curves constructed from Fig.* 6

resistance and that relating current to resistance. Some typical curves are shown in Figs. 5 and 6. For some applications, as we shall see below, it is useful to construct from these curves the curve relating voltage to current as shown in Fig. 7, from which it will be seen that for any given thermistor there is a maximum voltage which can be obtained across it.

All these curves are based on Equ. (1) above, the two parameters being:

1. The value of resistance at 25°C.
2. The value of B at 25°C.

The former is subject to a tolerance which may be as much as $\pm 33\frac{1}{3}\%$ for certain rod thermistors although is customarily $\pm 10\%$ or 20% for disc and bead types. The value of B is approximately constant and is also subject to a tolerance, usually of $\pm 5\%$ or 10%.

It is important to remember the effect of these tolerances on the performance of circuits containing thermistors. Narrower tolerances can usually be obtained only by selection, which involves considerable expense. The effect of the tolerance on B should be particularly noted. It can be shown[3] that the advantages of choosing a narrower tolerance on the resistance at 25°C disappear if the working temperature range is large.

The manufacturers of thermistors will usually also supply a curve showing the variations of resistance with time,

(a) when the thermistor is connected in a given circuit, and
(b) when the thermistor, being hot, is disconnected from that circuit.

These curves have only very limited use. In the case of (a) the characteristics of the remainder of the circuit and in the case of both (a) and (b) the physical situation in which the thermistor is placed, affect the resistance/time characteristic so greatly that for all ordinary purposes accurate information can only be obtained by experiment under actual design conditions. Fig. 8 shows the warming and cooling characteristics of a typical rod thermistor under stated conditions.

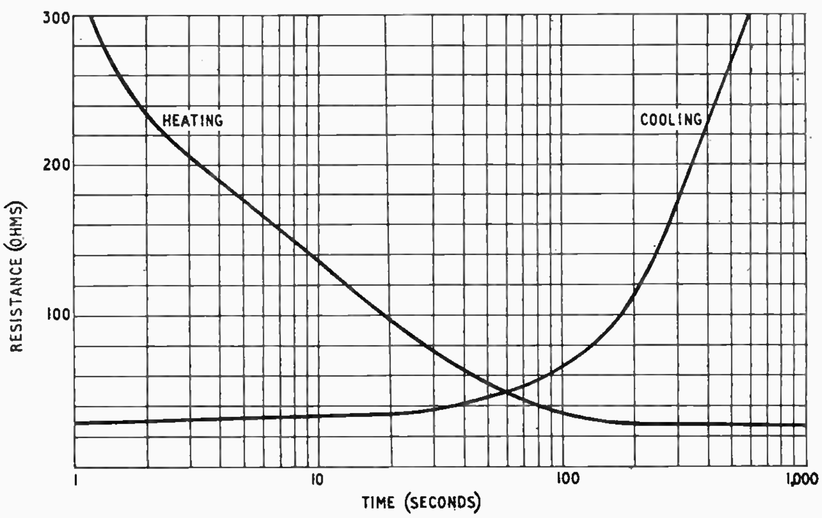
Sometimes the cooling property of the thermistor is described by the manufacturer in terms of some such phrase as 'recovery time', which has been defined by one manufacturer as the time taken by the thermistor to reach half the value of its resistance at 25°C, while cooling in still air of that temperature. Again, the 'dissipation constant' is defined as the power required to raise the thermistor's temperature by 1°C. It is quite obvious that this varies widely according to the ambient temperature so, here again, the information is of little more use than a guide when comparing two different thermistors for a given application. For bead-type thermistors, typical values are:

recovery time = 5 sec.
dissipation constant = 0·25 mW/°C.

By mounting the thermistor bead in an evacuated envelope the dissipation constant may be reduced to about a half or a third of the figure quoted. Naturally the vacuum envelope is considerably larger than the normal type and both are illustrated in Fig. 2.

A factor which is of considerable interest to the user is the stability of thermistor characteristics. Generally speaking, it will be found that if a thermistor is heated for a considerable time and then brought back to the

Fig. 8. *Variation of resistance with time. Thermistor connected in series with a typical string of valve heaters* (0·3 *A rating*). *Note that cooling time is considerably greater than heating time, which is effectively* 200 *seconds*



datum temperature (which is usually 20°C or 25°C, according to manufacturer) the resistance has increased slightly. The increase obtained approaches asymptotically a maximum figure in the neighbourhood of 1%. The effective drift experienced by the user can be reduced substantially by ageing the thermistor at a temperature of about 150°C for a period of about 200 hours. Indeed, thermistors intended for temperature measurement are usually aged for some period of time by the manufacturer. Naturally, since this is an expensive process, it is not carried out when extreme stability is not required[4].

Typical Values

Rod thermistors are available in sizes from about 10 mm long by 2 mm in diameter (corresponding to about 0·5 W maximum dissipation) to 40 mm long by

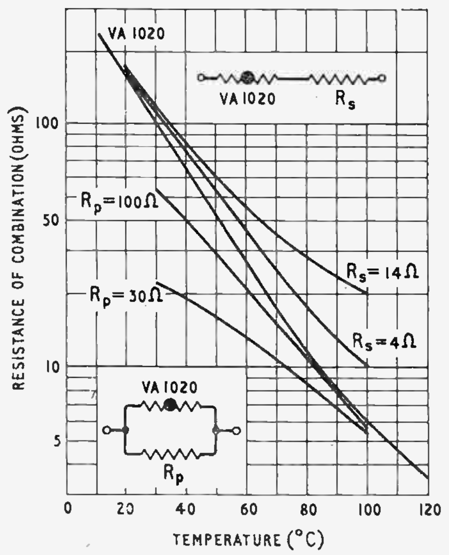


Fig. 9. *Family of curves showing effect of series or parallel resistance on resistance/temperature relationships of a typical disc thermistor*

15 mm diameter (over 5 W maximum dissipation). Resistance values at 25°C vary from some hundreds of ohms to a few hundred kilohms. These thermistors are mainly intended for 'current' applications and the maximum current-carrying capacity for the largest size is about 1 A.

The surface areas of disc and slab thermistors are of the order of 0·5 to 1·5 sq. cm and wattage ratings vary respectively from about 0·5 to 1·5 W. Resistance values at 25°C range from about 1 Ω to a few kilohms.

Bead thermistors, unlike the other types, are usually of a constant nominal physical size and hence the same wattage rating of about 50 mW. Resistance values offered range from a few hundred ohms to 2 MΩ.

In all the above statements the wattage rating is that obtained when the thermistor is mounted in still air.

Shaping the Characteristics

By connecting fixed resistors in series and/or parallel with the thermistor its resistance/temperature (and, of course, resistance/current) characteristics may be altered. Fig. 9 shows a family of curves obtained in this way. Although it is possible to derive algebraic expressions for calculating a suitable shaping network[5] the author has frequently found that a graphical computation, based on a common-sense approach to the particular conditions obtaining, gives satisfactory results quickly and a clearer insight into the operating processes[6].

It should be noted that two thermistors should not be operated in parallel if any degree of self-heating is likely to occur. Owing to the variations in characteristics between two nominally identical thermistors, an unbalanced sharing of the current will certainly occur, with unpredictable and perhaps disastrous results.

Mechanical Properties

As we have seen, thermistors are unglazed ceramics and they possess the characteristic properties of hardness, brittleness and some degree of porosity. Very sudden heating, due to an unreasonably heavy current-surge, may give rise to thermal-shock cracks, but otherwise their use should present no difficulties in normal domestic and industrial situations.

Where excessive moisture is likely to be present, however, protection from the atmosphere is necessary, both to prevent absorption of water by the thermistor itself and in the case of disc types to prevent oxidation of the silver film. Some manufacturers supply discs coated with lacquer; alternatively, the user may employ any good quality lacquer (e.g., impregnating varnish) or potting resin. In the latter case, however, care must be taken to see that the resin when setting does not 'seize' the silver film and strip it from the thermistor body. Silicone grease applied to the thermistor before immersing it in the resin will counteract this. Sometimes it is convenient to immerse the thermistor in a fluid, although this will, of course, alter the resistance/current characteristics. A good quality transformer oil or, ideally, silicone fluid is suitable for this purpose. (See Fig. 11.)

Applications

Thermistor applications fall into three clearly-defined groups:

(a) Those in which the thermistor is virtually used as a resistance thermometer, either to measure temperature or to compensate for the effect of temperature on other components.

(b) Those which make use of the heating effect of the current passing through the thermistor to vary its resistance and, hence, control that current or the potential difference across the thermistor.

(c) An extension of the second type of application in which use is made of the *time* taken for the thermistor to heat or cool so that the element of time is introduced into the control function.

Where cost is a consideration, for example in radio and television circuits and in normal industrial uses, disc or rod-type thermistors are invariably used; the former for the first group of applications, and the latter for the remaining two. The correct type of bead thermistor, on the other hand, may be used with advantage in all three applications where the special properties resulting from its small size are required and its considerably greater cost can be justified.

Temperature Measurement

The advantages of thermistors as temperature-measuring elements are that they can be made as small as thermocouples; the large resistance value obviates difficulties arising from the resistance of connecting leads; the large temperature coefficient makes for high sensitivity. Against these, they possess the disadvantage of a non-linear relationship between temperature and resistance and they require calibration on account of the tolerances on resistance and *B* factor already mentioned. Any type of thermistor may be used for temperature measurement although disc types are particularly suitable, and, in the bead range, the thermometer type shown in Fig. 3. The latter type can usually be supplied in lengths down to about 1 in. but the standard length is some 3 or 4 in. This form is particularly suited to biological work[7].

The thermistor may be used in a simple circuit which measures the current flow from a constant-voltage source (a method which has obvious limitations as regards accuracy) or it may be used in bridge circuits. An example of the former class is shown in Fig. 10 which is the temperature-sensing element of a car-radiator temperature indicator. In this instance the use of a double-element ammeter compensates for changes in supply voltage. A number of bridges have been described[8, 9, 10, 11] in which a thermistor is used in both arms of the bridge, the second being employed for compensation of changes in ambient temperature. A bridge of this type can be employed for measuring temperature differentials and, for this purpose, the provision of thermistors in matched pairs will obviously be a great advantage, but some manufacturers are unwilling to do this. In any case, the cost is very much increased, because matching can only be done by a process of selection.

Temperature Compensation

Thermistors may be used for compensating for the positive temperature coefficient of copper-wire coils. By reason of the much higher temperature coefficient of the thermistor, the thermistor resistance can usually be made some 10%–15% of that of the copper component. One such application is for the compensation of television deflector coils. This may be attempted in two ways; either a high resistance rod thermistor is mounted near the coil former and is then connected in an appropriate point of the drive circuit, or low resistance disc thermistors can be connected in series with the coil itself.

If thermistors are used for compensating measuring instruments the difficulty occasioned by their resistance tolerances becomes more acute. This can sometimes be countered by arranging for the thermistor to be shunted with a variable resistor, the value of which is adjusted during calibration.

'Self-Heating' Applications

Applications in this class fall into three groups: those which use the thermistor as an electrical circuit element whose resistance varies with current; those which measure the rate of conduction of heat away from the thermistor; and those which employ the thermistor to measure energy.

In the first class are circuits for voltage control[12] and, hence, of amplifier gain control[13, 14]. The basic principle of the voltage-control circuit is that when the thermistor is operating near the peak of the curves shown in Fig. 4 a wide variation of current flowing through it will produce only a small change in the voltage appearing across it. Starting from this point, a number of circuits have been developed for providing quite close voltage control over wide current variations with, in some cases, thermistor compensation for changes of ambient temperature[12].

At very low frequencies the thermistor will exhibit reactive effects[1] due to the time-lag between the flow of current and the heating effect spreading throughout the thermistor body. This phenomenon can be utilized in low-frequency oscillators.

Finally, thermistors may be employed as high resistance shunts across, for example, radio-receiver dial lamps. When the lamp is connected in series with the valve-heater chain the receiver is rendered inoperative if the lamp fails. If the lamp is shunted by a thermistor, however, and it fails, the thermistor will carry the full load current, heat up and its resistance will fall to a value which can be selected to be approximately equal to that of the dial lamp.

In the second class, we have a number of ingenious applications which are not really specific to thermistors but in which thermistors can often be used with great advantage because of their small size. That is to say, these are applications for bead-type thermistors[15]. Heat is supplied to the thermistor by means of an electric current and the rate at which it is conducted away is an indication of the thermal conductivity of the medium in which the thermistor is placed. A special glass mounting (Fig. 3) has been designed for use in vacuum measurement and for gas analysis. (The glass tube is sealed on to the equipment containing the gas.) Two thermistors connected in a bridge can be employed for liquid flow measurements, and thermistors arranged at varying heights in a tank can be employed as liquid level indicators in so far as when the thermistor is covered by the liquid its resistance will not fall to such a low value as when it is held in the open air. Thermistors used in this way may be rather too sensitive to liquid splash and other short-term effects, which can be counteracted by connecting the thermistor in series with the coil of a simple thermal delay switch (such as a bimetal type) to give the necessary time-constant to the circuit. The arrangement is illustrated in Fig. 11.

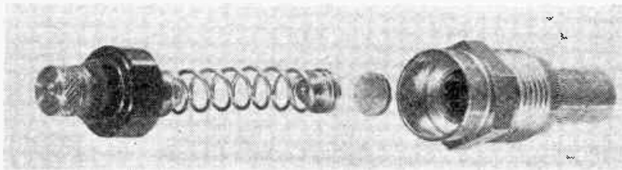


Fig. 10. *Exploded view of a car-radiator temperature-sensing element using a disc thermistor (Courtesy, Smith's Motor Accessories, Ltd.)*

Bead thermistors are particularly suitable as a.c./d.c. comparators and may be used in high-frequency bridges and for power measurement at very high

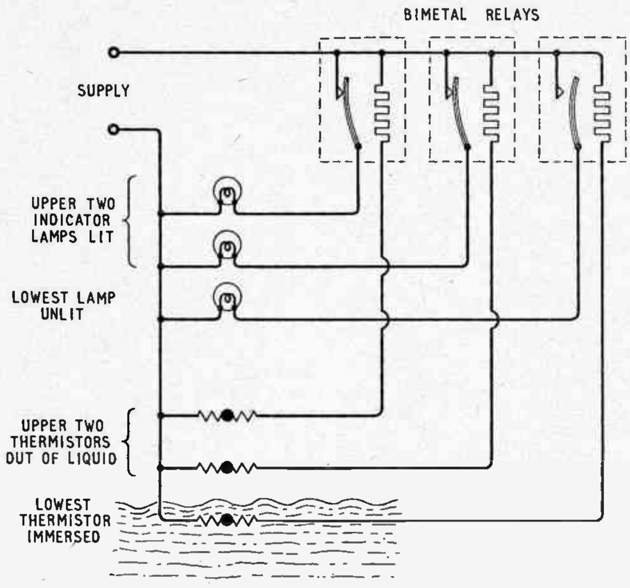


Fig. 11. *A simple liquid level indicator. When the thermistor is immersed in the liquid the heat generated by the current passing through it is conducted away. Hence the resistance remains high and the current is insufficient to actuate the relay. The function of the relay is to increase the time-constant of operation and hence prevent the lamp from indicating when the thermistor is only momentarily immersed or splashed*

frequencies generally. A number of articles have appeared in the literature on this subject[16, 17, 18]. A similar application is in the measurement of microwave energy[19], and as a radiometer[20, 21].

Indirectly-Heated Bead Thermistors

Thermistors may, of course, be heated by a separate heating element connected in a different electrical circuit, thus providing a four-terminal device. A special bead thermistor is made in which the heater is incorporated in the same glass envelope as the bead with which it is in intimate contact, and this is illustrated in Fig. 2, typical characteristics being shown in Fig. 12. One application of this type of thermistor is in amplifier gain control[14]. The heater is supplied with current proportional to the output of the amplifier, while the thermistor element is connected in the input network in such a way that when the output rises the thermistor resistance decreases and reduces the amplifier input voltage.

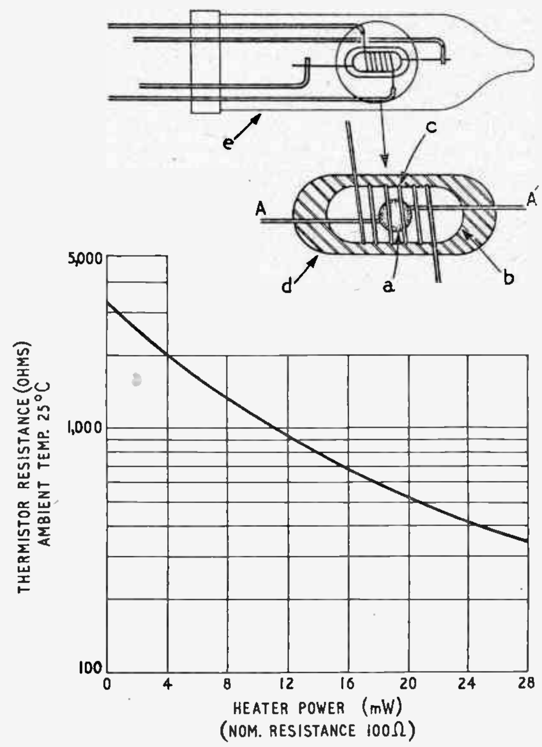
Time Delay Circuits

By far the widest use of rod-type thermistors is for the limitation of surge currents in valve-heater chains. Modern practice in the design of, for example, television receivers is to connect all the heaters in series together with a small dropping resistor and to put the chain thus formed straight across the supply voltage. The number of valves involved in the modern receivers is such that the dropping resistor is often required to drop as little as 8–20 V. When this chain is first switched on, the heaters being cold, the surge of current which flows is considerably in excess of the normal value. Moreover, since the time taken for some valves to heat up to the normal running temperature is different from that of others, a voltage in excess of the maximum permitted by the manufacturer may appear across certain valve heaters with consequent risk of early failure.

If a thermistor is employed as the dropping resistor in the chain the surge current can be limited to a very much lower value. The time taken for the resistance of the thermistor to fall to the value required for normal running conditions depends, amongst other things, on the latter's thermal capacity and the conditions governing the rate at which heat can be lost. The former can be increased by adding to the thermistor mixture non-conducting ceramic material to increase its thermal mass. Occasions may arise when, with the thermistor available to the designer, it either heats up too quickly for his purpose on the one hand, or never reaches the required low resistance value on the other.

The first difficulty can often be overcome by shunting the thermistor with a fixed resistor. At first it may seem to be paradoxical that the surge current can be reduced in this way; however, it must be borne in mind that the 'cold' resistance of the thermistor is usually very much higher than is required for surge-limiting purposes. Suppose, for example, that the thermistor is shunted with a fixed resistor having a value equal to its cold resistance. This will halve the initial resistance, but the initial current will still be well within the maximum permitted. The power dissipated in the thermistor, however, has been reduced to a quarter of that which was obtained in the unshunted case and consequently the initial rate of heating is reduced by the same factor. Hence the net effect is to decrease the rate at which the thermistor

Fig. 12. *Typical characteristics and* (*inset*) *diagram of internal construction of an indirectly-heated bead thermistor. a = bead thermistor with leads A and A' embedded in glass bead, b, round which is wound the heating element c. The combination is then fused in glass, d, and supported in an evacuated envelope, e*



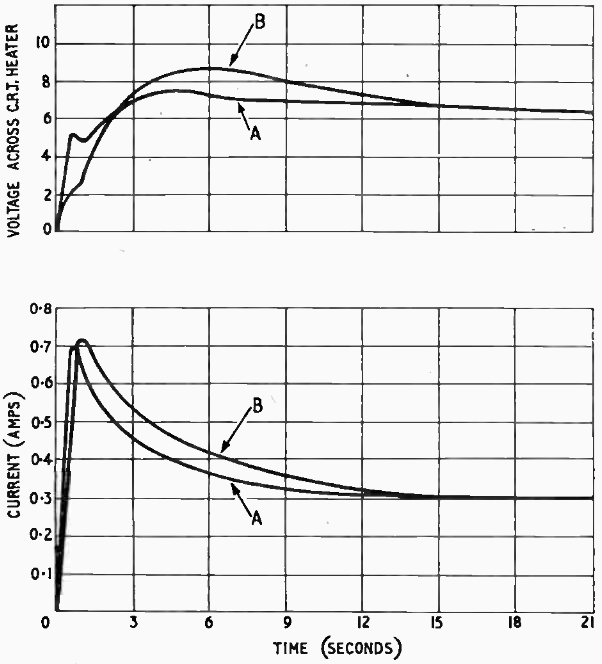


Fig. 13. *The variation in voltage across, and current through, a c.r.t. heater connected in series with a typical string of valves and a thermistor. Curves (A) relate to a normal rod thermistor shunted with a resistance of* 130 Ω; *curves (B) relate to a thermistor of the equivalent 'shuntless' type*

resistance falls. The effect of various thermistors and shunt resistors is shown in Figs. 8 and 13.

The case of the thermistor whose resistance never falls to the value required can be accommodated by increasing, artificially or otherwise, its ambient temperature and hence reducing its heat-losing capacity. For example, the thermistor can be lagged with a piece of glass-fabric sleeving or it may be possible to place it very close to a component which itself becomes hot. As has been stated previously, there is a maximum temperature permissible for all thermistors and in the case of rod types it is usually about 120°C, and this alone governs the minimum resistance obtainable. It does not matter whether the temperature of the thermistor is raised to its maximum by virtue of the heat generated by the current which flows or by ambient conditions or by a combination of both.

Projector lamps of the smaller powers (150 W, 250 W) such as are used in 'home projectors' are said by some users to suffer from the effects of the current surge which occurs when they are first switched on. Thermistors can be used for the protection of such lamps and special low-resistance types have been developed for this purpose.

A thermistor placed in series with a relay coil can be used to delay its time of operation[22]. 'Slugging' the relay by means of copper shorting rings produces accurate delays but of duration less than one second. Resistor-capacitor networks can extend this delay to several seconds, but these become elaborate and bulky for longer delays. A thermistor placed in series with the coil can produce delays from a few seconds to several minutes. However, this method has its limitations; first, there is liable to be a considerable variation in the delay time obtained with different samples of the same type of thermistor because of the tolerance on thermistor properties which has been previously described. Secondly, due account must be taken of the possible effects of a change in ambient temperature. Thirdly, it must be remembered that the delay obtained with successive operations will be reduced if the thermistor has not had time to cool completely following the previous operation. In order to provide maximum cooling time it is advisable to arrange for an auxiliary pair of contacts to short out the thermistor once the relay has actually closed. This also makes for more definite operation.

Acknowledgements

The author wishes to thank the Directors of Mullard, Limited, for permission to publish this article. He takes no credit for any of the devices and circuits described, but is indebted to his many colleagues for a great deal of advice which has been freely given over a period of years. His especial thanks are due to Mr. B. A. Evans, of Mullard Research Laboratories, for constructive criticism in the preparation of the article.

REFERENCES

1 "Reactive Effects in Thermistors at Very Low Frequencies." F. J. Hyde. *Brit. Commun. Electronics*, Jan., 1957, Vol. 4, No. 1, pp. 16–17.
2 "Low Resistance Thermistors as Ultra-Cold Thermometers." H. B. Sachse. *Electronic Ind. Tele-Tech.*, January, 1957, Vol. 16, No. 1, p. 55.
3 "N.T.C. Resistors." *Matronics* (Technical Information Bulletin, N.V. Philips), August, 1954, No. 7.
4 "The Stability of Thermistors." A. Beck. *J. Sci. Instrum.*, January, 1956, Vol. 33, pp. 16–17.
5 "Shaping of the Characteristics of Temperature Sensitive Elements." E. Keonjian and J. S. Schaffner. *Trans. Amer. Instn elect. Engrs*, September, 1954, Vol. 73, Pt. 1, pp. 396–400.
6 "Designing Thermistor Temperature-Correcting Networks Graphically." Bennett. *Control Engng*, November, 1955, pp. 66–68.
7 "The Thermistor in Biological Research." B. L. Andrew. *Electronic Engng*, September, 1947, Vol. 19, p. 288.
8 "Temperature Controller." Noltingk and Snelling. *J. sci. Instrum.*, 1953, Vol. 20, pp. 349–51.
9 "A Simple Direct-Reading Thermistor Bridge." J. Swift. *Proc. Instn Radio Engrs, Australia*, 1956, Vol. 17, pp. 341–45.
10 "Oven Control with Thermistors." Koerner. *Bell Lab. Rec.*, 1953, Vol. 31, pp. 115–118.
11 "Direct-Reading Thermistor Bridge." Pearson and Benson. *Electronic Eng.*, 1953, Vol. 25, pp. 51–57.
12 "Voltage Regulating Circuits with Temperature Compensating Thermistors." Dietrich. *Elektronische Rundschau*, November, 1956, Vol. 10, No. 11, pp. 302–5.
13 "A Thermistor Device as Automatic Gain Control in Lamp–Photocell Transducer Systems." A. M. Hardie. *J. sci. Instrum.*, February, 1957, Vol. 34, pp. 58–62.
14 "A.G.C. by Means of Miniature N.T.C. Resistors with Heating Element." Schounten. *Electronic Application Bulletin* (N.V. Philips), 1951, Vol. 12, No. 2, pp. 33–37.
15 "New Thermistor Applications." Hubelbank. *Electronic Design*, March, 1955, p. 32.
16 "Load Measurement by Means of Thermistors at High Frequency." H. Groll. *Fernmeldetech. Z.*, 1952, Vol. 5, pp. 522–27.
17 "Application of Thermistors to Bridges for Measuring Low Conductance at High Frequencies." Soldi. *Alta Frequenza*, 1952, Vol. 21, pp. 243–259.
18 "A New Thermistor Power Measuring Unit for 9–20 cm wavelengths." H. Rieck and F. Panniger. *Nachrichtentechnik*, March, 1957, Vol. 7, No. 3, pp. 101–4.
19 "Microwave Power Measurement." R. N. Greinheimer. Chapter 3 in "Technique of Microwave Measurements," M.I.T. Radiation Laboratory Series, Vol. 11, pp. 79–220. McGraw-Hill.
20 "Two Portable Thermistor Radiometers." W. H. Ward. *J. sci. Instrum.*, August, 1957, Vol. 34, pp. 317–321.
21 "Measurement of Power at a Wavelength of 3 cm by Thermistors and Bolometers." J. A. Lane. *Proc. Instn elect. Engrs*, Part B, 1955, Vol. 102, p. 819.
22 "Relay Time-Delay Circuits with N.T.C. Resistors." Linau and Siefert. *Matronics*, April, 1957, No. 12, pp. 216–222.

ADDITIONAL REFERENCES

"The Properties and Applications of Thermistors." *Onde élect*, 1953, Vol. 33, pp. 497–509 and 578–90.
"Conductivity of Semi-Conductors and Thermistors." N'Guyen and Suchet. *Onde élect*, 1951, Vol. 31, pp. 473–89.
"Characteristics and Applications of . . . Thermistors." Howes. *Proc. Instn Radio Engrs, Aust.*, 1952, Vol 13, pp. 123–31. Also *J. Brit. Instn Radio Engrs* April, 1953, p. 28.
"Semi-Conductors with large Negative Temperature Coefficients." N'Guyen and Suchet. *Ann. Radioélect.*, 1950, Vol. 5, pp. 155–67.
A Study of Thermistor Circuits. Ekelhof and Kihlberg. *Chalmers Tekn. Hogsk. Handl.*
(a) "Small-Signal Behaviour of Directly Heated Thermistors." 1955, No. 169, 43 pp.
(b) "Large-Signal Behaviour of Directly Heated Thermistors." 1954, No. 185, 30 pp.
(c) "Theory of the Thermistor as an Electric Circuit Element." 1954, No. 142, 35 pp.
"Electron Microscope Study of Sintered Thermistor Flakes." *Bell Lab. Rec.*, 1952, Vol. 30, pp. 285–7.

Nomogram for Air-Gap Design

By A. C. Sim*

SUMMARY. *An air-gap is necessary in the iron core of a choke or transformer if its inductance is to be reasonably controlled or stable. A nomogram is provided which relates the actual air-gap to its effective value for normal types of laminations, based on a theory due to Partridge.*

A simple but adequate estimation of the effect of fringing around an air-gap in a core has been fully described by G. F. Partridge[1] and, it appears, subsequently forgotten. His formula has been found to give adequate accuracy and a practical application of it is described here.

It is well known that owing to fringing effects the reluctance R_g of an air-gap in a choke or transformer core is always less than that implied by the length of the gap. If this fringing is ignored in the design of an iron-cored inductor the resulting inductance will be too large, and removal of laminations to reduce it may increase the flux density to an over-heating level. The procedure to be described gives, within a wide practical range, an estimation sufficiently close to eliminate this difficulty.

Partridge found that for a core having a single central air-gap and a winding of N turns, the inductance is

$$L = \frac{4\pi N^2 10^{-9}}{(R_c + R_g)} \text{ henrys}$$

where $R_c = l/\mu A$ is the reluctance of the core and R_g is the reluctance of the air-gap, l being the average magnetic path length, A the cross-section of the core leg and μ the permeability of the core. His approximate theory resulted in the formula

$$1/R_g \approx [A/g + \tfrac{1}{2}P \log_e (S/g)]$$

where g is the actual air-gap length, P is the perimeter of the core leg and S the length of the leg.

Let $\gamma = g/l$ be the actual ratio of air-gap length to total magnetic-path length, and γ_e be the effective air-gap ratio so that

$$L = \frac{4\pi N^2 10^{-9}}{l(1/\mu + \gamma_e)} \text{ henrys.}$$

Then it follows that

$$\gamma_e = \frac{A}{[A/\gamma + \tfrac{1}{2}Pl \log_e (S/\gamma l)]}$$

Nomogram

Consider the application of this formula to a normal type of core as illustrated, having a centre leg of width a and a stack depth of b with a window of height h.

If the perimeter is P_1 at the centre leg and P_0 at the outer legs, it is not difficult to agree that the effective value of perimeter P for a gap in all three legs is $P \approx \frac{1}{2}(P_1 + P_0) = \frac{1}{2}(2a + 3b)$.

* *Standard Telecommunication Laboratories Ltd.*

Let s be the stacking factor of the core, usually $1 \cdot 1$, equal to the actual cross-section divided by the iron cross-section. Then with γ_e the effective gap ratio required in design (and therefore known) and γ the actual gap ratio necessary in practice, define the following quantities:

Case 1. *Gap in Centre Leg Only.*

$$G = (a + b)\, l/2ab$$
$$H = 2\cdot 3\, G \log_{10} (l/h)$$
$$X = 1/s\gamma_e$$
$$Y = 1/\gamma = l/g$$

Case 2. *Gap in All Legs.*

$$G = (2a + 3b)\, l/4ab$$
$$H = 2\cdot 3\, G \log_{10} (l/h)$$
$$X = 2/s\gamma_e$$
$$Y = 2/\gamma = 2l/g$$

The equation is then simply $X \approx (Y + H + G \log_e Y)$ and, from Partridge's results, the errors should not exceed 10% if (h/l) lies between $0\cdot 25$ and $0\cdot 5$. The nomogram presented represents this equation and enables Y to be found from X, G, and H.

Application

Calculate X, G, and H, using a value of effective gap ratio which gives the desired inductance. Join the appropriate value of H to that of X, this line intersecting the reference line R at a particular point. This point on R must then be joined to the appropriate point on G, and the intersection of this second line with Y determines the actual gap ratio necessary to achieve this effective gap.

As an illustration, consider a core made from Sankey No. 4 laminations stacked $1\frac{1}{2}$ inches deep. For such a core the appropriate constants are (in inches):—

$a = 0\cdot 94$; $b = 1\cdot 5$; $l = 8\cdot 0$; $h = 2\cdot 3$; $s = 1\cdot 1$
$h/l = 0\cdot 29$; $2\cdot 3 \log_{10} 0\cdot 29 = -1\cdot 24$; $\gamma_e = 0.06$ (say)

1. *Centre Leg Gap Only.*

$G = 7\cdot 0 \qquad H = 8\cdot 6 \qquad X = 120$

Joining $H = 8\cdot 6$ to $X = 120$ on the scale X_2, noting the intersection on R and joining this point to $G = 7\cdot 0$ gives $Y = 83$ from scale Y_2 (which must be used with X_2). From this, to obtain an effective air gap of $0\cdot 06$ in. an actual gap of about $\frac{1}{5}$ in. is necessary ($\gamma \approx 0\cdot 03$).

2. *Gap in All Legs.*

$G = 9\cdot 1 \qquad H = 11\cdot 3 \qquad X = 240$

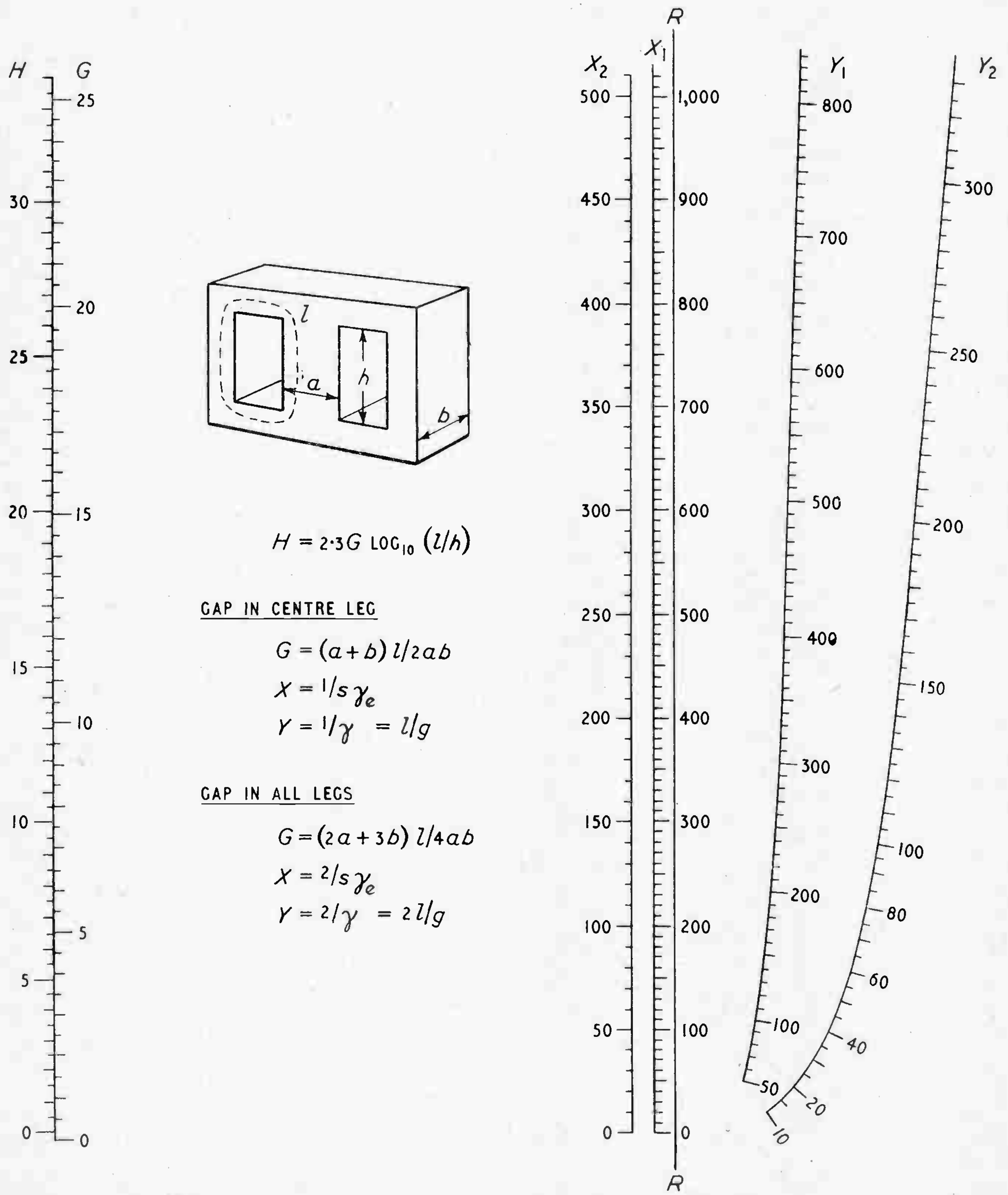
Joining $H = 11 \cdot 3$ to $X = 240$ on X_2 (or X_1), noting the intersection point on R and joining this point to $G = 9 \cdot 1$ shows on Y_2 (or Y_1) the value 185. In this case γ is $0 \cdot 011$ and the actual gap necessary is reduced to about $0 \cdot 043$ in. by spacing all three legs.

It can be seen that if the gap were reduced to $0 \cdot 03$ in. (ignoring fringing), Y would increase to about 280 and from the nomogram (using Y_2 and X_2), X would be increased from 240 to 340 making a large reduction in the effective air-gap. Thus a gap of as little as $0 \cdot 010$ in. introduces fringing which should not be ignored.

REFERENCES

[1] G. F. Partridge, "The Inductance of Iron-Cored Coils having an Air-Gap". *Phil. Mag.*, October 1936, Vol. 22.

Nomogram for air-gap design



THE ROENTGEN, THE REP, AND THE REM

Like so many other people, I spend a certain amount of time helping the local Civil Defence training—chiefly, it must be admitted, in seeing that the meters are serviceable, and reporting with a frenzied efficiency the first signs of a desiccator turning pink. During the past few years, I have had, if only on this account, to keep fairly up to date with some aspects of nuclear physics, and learned to think in terms of doses, dose-rates, and decay. More enterprising types may spend their leisure leading double lives; I simply happen to find half-lives a pleasanter distraction. But it doesn't do to be too smug. It was rather a shock to my complacency to read that Sir John Cockcroft, speaking of the accomplishments of a Minister of the Crown, had said, "He does of course have to understand the language of science; he must know the difference between reps and rems." Which, I was horrified to realize, was certainly more than I myself, torrential pourer-forth of scientific language though I am, could say that I did. The abbreviations for "roentgen-equivalent, physical", and "roentgen-equivalent, man" are of course well-known, but outside the direct scope of γ-ray dosimetry as I have met it—you will soon see why. I wondered how many professional scientists, taken off their guard, could give you the answer; it isn't in the latest "Kaye and Laby", so I guessed not so many. Samuel Glasstone's "Sourcebook on Atomic Energy" (1952), and the International Labour Office report on "The Protection of Workers from Ionizing Radiations" (1955) had the information. I am not ashamed to admit that I had to look it up in this way—after all, in these days of specialization there must be many scientists outside the radio world who don't know the difference between a bel and a dragon, and associate "Q" solely with tales of Troy town.* Radiation monitoring is itself so widespread nowadays that I thought it might be worth while to say a little of the principles behind it, and pass on the revelation of the rep and the rem; it will at least help you to feel on equal terms next time you meet your Member of Parliament.

Ionizing Radiations

The sole purpose of dosimetry is to calculate the total effect of a certain quantity of energy when it is absorbed, in a particular way, by the human body. From this point of view, radiations can be classified, in order of importance as:

(*i*) *Penetrating photon radiation.* X-rays and γ-radiation. X-rays are electromagnetic radiation, of wavelength lying between about 10^{-10} and 10^{-7} cm, emitted from high-voltage electron tubes of all types—including electron microscopes and cathode-ray tubes. The accelerating voltage across the tube determines the quantum energy of the radiation, and hence its penetrating power; while the tube current determines its intensity. γ-rays are shorter wavelength electromagnetic waves which originate from nuclear changes in natural or artificially-produced radioisotopes. Both X-rays and γ-rays penetrate great distances in air, which they ionize sparsely along the whole path; they also give rise to secondary radiation and scattered radiation in the materials through which they pass.

(*ii*) *Charged corpuscular radiation.* α-particles and β-particles. These come only from radio-active isotopes. The α-particles are helium nuclei (He^{++}), and the β-particles fast-moving electrons. Both have very short ranges in air, which they ionize, and consequently are a less general radiation hazard than X- and γ-rays; they cannot as a rule do much harm unless the source is very close to the body, or absorbed inside it.

(*iii*) *Neutrons.* These, being uncharged, have no direct ionizing effect; but they interact with the nuclei of atoms, the result being a radioactive change leading to the emission of γ-rays, α-particles, and β-particles. While these constitute a very serious hazard to people working in the neighbourhood of nuclear reactors, or experimenting with neutron-emitting sources, the number of people likely to be exposed to them is very limited. Fast neutrons, with energies of several MeV, produce one class of nuclear changes. Neutrons are slowed down (moderated) during their passage through substances rich in light-element atoms, losing energy on colliding with, say, hydrogen nuclei just as a billiard-ball does on striking another at rest. Slow neutrons ('thermal neutrons'), with energies of the order 0·025 eV, produce another class of change—the (n, γ) reaction, or radiative capture, in which the neutron is absorbed by a nucleus and a γ-ray photon is emitted.

The Roentgen

The roentgen (r) is a unit for measuring quantities of X-rays *using their ionizing effect*. The experimental basis for it is the saturation ionization current measured between the plates of a parallel-plate air capacitor, the total charge of either sign being obtained from the

* Of course I wrong them. Outside the delectable Duchy and the wireless world, most people would recognize this as the symbol for the MeVs involved in a nuclear change.

product current × time. It was defined in 1928 as "The dose which, when the secondary electrons are fully utilized and the wall effects of the chamber avoided, produces in 1 c.c. of atmospheric air at normal temperature and pressure such a degree of conductivity that one electrostatic unit of charge is measured at saturation current". The reason for specifying what is really the mass of air is that the ionization depends on the number of electrons that are available to absorb energy from the beam—that is, on the atomic number Z of the absorbing atoms (taken as 7·7 for air) and the number of atoms present. The definition was later amended, when the need arose, to include γ-rays; the term "quantity of radiation" was then used.

Dose, or quantity in this sense, is not the same as the total number of quanta in the beam of rays, though it must depend both on this and also on the penetrating power, hardness, or quantum energy—that is, on the frequency. The more penetrating the radiation, the smaller is the percentage of the total quanta absorbed, and the greater the total number of quanta needed to provide a given dose. Further, it must not be confused with the ionization itself. The roentgen is not the quantity of energy trapped by one particular c.c. of air; it is the quantity of radiation which will liberate that quantity of energy in *every* c.c. of air in its path—or at least in the early stages of it, before absorption has reduced it appreciably.

The only problem to be faced in adopting the roentgen as a unit for γ-ray doses as well was that of relating it to the usual radioactive standard γ-ray source of those days. Once this purely technical difficulty was overcome, and it was established that the dose at a given distance from a shielded one-milligram radium source could be measured by the ionization chamber and compared with X-ray doses, the job was done. The revised definition of 1937 said:

"The roentgen shall be the quantity of X- or of γ-radiation such that the associated corpuscular emission per 0·001293 gm of air produces, in air, one electrostatic unit of quantity of electricity of either sign."

You will probably find the earlier definition in most reference books, except Glasstone. This one was taken from a review article written in 1938 by W. V. Mayneord, in the Physical Society's *Report on Progress in Physics* for that year, where there is a most interesting account of the development of radiation-dose measurement. The only difference is the explicit statement of a mass of air, instead of a recipe for working out that mass; but this is clearly preferable, because the absorption of a beam of such rays depends directly on the mass of material (or better, number of atoms; or better still, number of electrons) presented to it as a target. Further, when the effect on substances other than air (for example, human tissue) is to be computed, the masses (not the volumes) of these materials are used.

Energy Converted

Consider now the effect of one roentgen on one gram of air. The charge on a singly-charged ion of either sign is $4{\cdot}8 \times 10^{-10}$ e.s.u.—the electronic charge. The number of electronic charges in one electrostatic unit is the reciprocal of this, that is, $2{\cdot}08 \times 10^9$. The roentgen liberates this number of ion-pairs in 0·001293 gm of air, and so in one gram liberates $2{\cdot}08 \times 10^9/0{\cdot}001293 = 1{\cdot}61 \times 10^{12}$ pairs. The energy required to produce a pair of ions in air is about 32·5 eV, so the energy given to one gram of air by one roentgen is $1{\cdot}61 \times 10^{12} \times 32{\cdot}5 = 5{\cdot}23 \times 10^{13}$ eV. Remembering that one electron-volt is approximately $1{\cdot}6 \times 10^{-12}$ ergs, this works out to about 83 ergs.

This is really getting somewhere—almost approaching an absolute measurement of dosage in ergs per gram. But only in ergs per gram of air, unfortunately. It is found that one roentgen of X-rays liberates more nearly 90 ergs per gram in human soft tissue, and a great deal more in bone. Mayneord gives figures in his article for the energy release per roentgen per *electron* (readily convertible to per gram if the effective atomic weight A and atomic number Z are known) for different materials. For high-energy γ-rays, there is little to choose between air and substances such as water, fat, and tissue. For X-rays of wavelength between 2×10^{-9} and 10^{-8} cm, the energy conversion in water and tissue is about 70% of that in air, per electron, while in carbon and fats it is less than half as much; it does not appear to vary with the wavelength in a given substance over this range. It is evident that, though air does not quite match the body tissues and water in its energy-absorbing power, the correspondence is close enough for the roentgen to be ideal for its purpose.

The Rep (Roentgen-Equivalent-Physical)

The roentgen is a measure of quantity of *electromagnetic* radiation. This radiation both ionizes air, and liberates energy on the same scale in body tissue. Other ionizing radiations, α-particles and β-particles, protons, and also the indirectly ionizing neutrons, produce the same kind of bodily effects which are, however, less directly correlated with their ionizing effect in air. In order to be able to measure doses of corpuscular radiation, the rep was defined as the quantity of radiation of this kind which upon absorption by the body tissue is accompanied by the gain of 83 ergs of energy per gram of tissue. Two points call for comment here. The first is, that the *electromagnetic* energy didn't have to be all absorbed—the roentgen just fell on the gram of air (or tissue) and gave up 83 of the large number of ergs it might be carrying. Actually, α-particles and β-particles must necessarily be absorbed, for they do not travel far even in air; I rather feel that, as far as neutrons are concerned, there is a loophole or ambiguity in the definition —for it doesn't seem certain whether it refers to the quantity of neutrons which will give 83 ergs from the fraction of the beam absorbed, or whether it means the quantity which all have to be absorbed to do this. However, I may be quibbling; 83 ergs absorbed per gram is the physical result of a roentgen, and any quantity of corpuscular radiation which gives up 83 ergs per gram is a rep. The second point is, that actually the figure for soft tissue in the case of X-rays is well above 83 ergs, as has been mentioned. This is, however, not very far out for the purpose of this kind of measurement.

The Rem (Roentgen-Equivalent, Man)

This should be roentgen-equivalent, biological really. It is the quantity of radiation which produces the same biological damage in man as that resulting from the absorption of 1 roentgen of X-rays or γ-rays. Here again the point about "absorption" crops up, but it really means "action" in the case of electromagnetic radiation. The reason why this unit was needed is that different kinds of radiation which produce the same physical release of energy may have very different biological effects; neutrons and α-particles, for example, are much more damaging than X-rays. Glasstone gives the provisional relationships of Table 1 between the roentgen, the rep, and the rem, for different types of radiation.

The need for this independent unit, quite apart from the differential effect shown in the last column, becomes clear when the time-factor is taken into account. The physical (or rep) effect is simply proportional to dose-rate $\times$ time. The biological effect is certainly not a straightforward time-integral—a low dose-rate in rems/hour spread over a long time is less damaging than the same number of rems absorbed as a high dose-rate for a small time. Also, the equivalence table, rough as it is, enables an assessment to be made of the result of exposures, simultaneously or successively, to different kinds of radiation.

TABLE 1

Type of radiation	roentgen	rep	rem
X-rays, γ-rays	1	1	1
β-rays	—	1	1
Fast neutrons	—	1	10
Thermal neutrons	—	1	5
α-particles	—	1	10–20

In the case of α-particles, it is assumed that the source is within the body.

Well, I hope I have cleared the vocabulary part of this up; you will find all the further information that you need in the references I gave at the start. To the awkward questioner who begins to wonder where the curie comes in, I can only say that it doesn't in this connection; and the measurement of the strengths of radioactive sources is quite a different story which might be taken up a little later on.

Bifilar-T Trap

AUDIO-FREQUENCY APPLICATIONS

By A. Hendry and A. G. McIntosh*

SUMMARY. *Circuits utilizing the bifilar-T trap in the audio-frequency range are described. These include a circuit for the rejection of harmonics, particularly the second, of 1000 c/s signals; a tuned audio-frequency amplifier of a few hundred cycles bandwidth, with high attenuation in a very narrow band at the centre of the passband; and a narrow-band feedback amplifier. Formulae are given to permit a designer to modify the performance of the basic bifilar-T circuit.*

The bifilar-T trap[1] is primarily used in television-receiver intermediate-frequency amplifiers, in which it is used to trap unwanted signals such as adjacent-channel sound carriers and adjacent-channel picture carriers. In these applications it is very suitable because of its high rejection capabilities, which can be obtained without critical adjustments or the use of precision components. Since very large rejections can be obtained with components of moderate Q, equivalent performance should be obtainable at any frequency for which components of moderate Q can be constructed. With the use of ferrite pot cores, the useful frequency range of the trap covers the audio-frequency range, thus the following work was undertaken in order to evaluate the performance of the trap at audio frequencies, and to develop a circuit suitable for rejecting harmonics of the 1-kc/s signals used in an infra-red detector amplifier.

* *National Research Council, Ottawa, Canada.*

Principles of Operation

The circuit of the bifilar-T trap is given in Fig. 1. The two halves of the bifilar coil, each of inductance L, are very closely coupled so as to provide an inductance of $4L$, as indicated. The operation of the circuit can

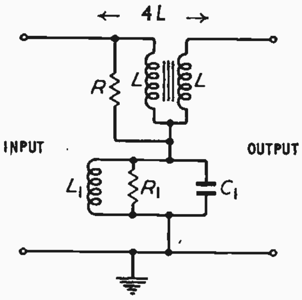


Fig. 1. *Bifilar-T trap circuit*

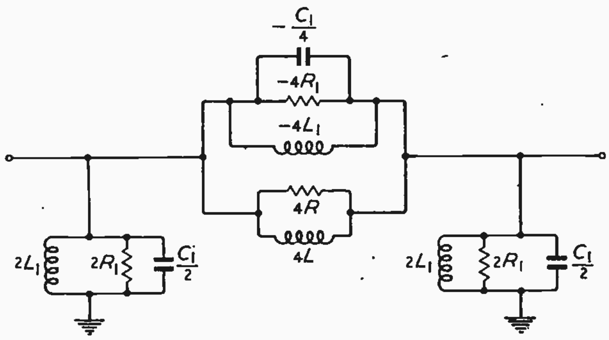


Fig. 2. *Equivalent circuit of the bifilar-T trap*

best be described with reference to its equivalent circuit, given in Fig. 2. The derivation of this equivalent circuit from the actual circuit, using star-delta transforms, is fully described in reference 1, and will not be repeated here.

Note first that the series arm is self-resonant at a frequency given by

$$f_0 = \frac{1}{2\pi\sqrt{\left(\frac{-4LL_1}{L-L_1}\right)\left(\frac{-C_1}{4}\right)}}$$

providing that L is greater than L_1. Note also that, subject to the approximations made in deriving the equivalent circuit of Fig. 2, the impedance of the series arm can be made infinite if $R = R_1$. (The resistances R and R_1 include the effects of losses in L and L_1.) In practice, very high impedances can be obtained, resulting in very high attenuations at f_0. The choice of $R = R_1$ is most easily made by using a rheostat for R, and adjusting it to obtain maximum rejection at f_0. In the derivation of the equivalent circuit shown, only the circuit elements indicated in Fig. 1 were considered. Capacitive coupling between the two halves of the bifilar coil, leakage reactance of the bifilar coil, and stray capacitance across the input and output of the circuit have been neglected.

The resonant frequency of the shunt arms is

$$f_1 = \frac{1}{2\pi\sqrt{L_1C_1}}$$

at which frequency there will be high impedance in the shunt arms, and therefore a maximum in the response curve, while

$$f_0 = \frac{1}{2\pi\sqrt{\frac{LL_1}{L-L_1}\cdot C_1}}$$

at which there will be rejection. Thus

$$\frac{f_1}{f_0} = \sqrt{\frac{L}{L-L_1}}.$$

If we let $L/L_1 = \alpha$ (which must be greater than 1) then

$$\frac{f_1}{f_0} = \sqrt{\frac{\alpha}{\alpha-1}}$$

It may be seen that f_1 is always higher than f_0, except in the limiting case of $L \gg L_1$, when f_1/f_0 approaches 1.

The frequency response of a filter such as that of Fig. 1 will always resemble the curve of Fig. 3; i.e., peak response at a higher frequency than the notch frequency, as long as no external elements are added and the component values are such that stray capacitances may be neglected in comparison to $C_1/2$.

Modification of Performance Using External Elements

Referring again to the equivalent circuit of Fig. 2, it can be seen that if capacitances are added in parallel with $C_1/2$ at the input and output of the filter, the resonant frequency of the shunt arms will be lowered. Thus the frequency of maximum response can be adjusted to any value lower than f_1.

To analyse the effect of added shunt capacitance, let shunt capacitors C_s be added as shown in Fig. 4. The

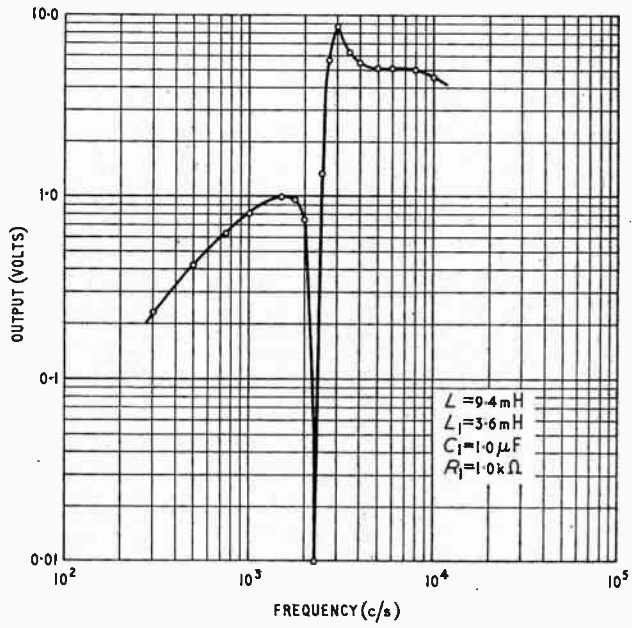


Fig. 3. *Frequency response of bifilar-T trap*

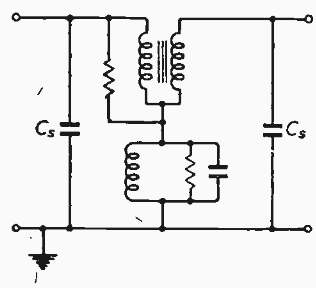


Fig. 4. *Bifilar-T trap modified by shunt capacitors*

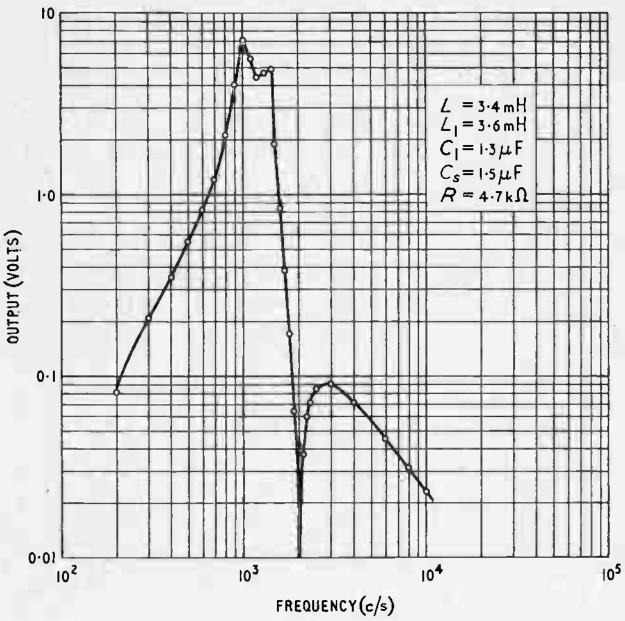


Fig. 5. *Frequency response of modified bifilar-T trap*

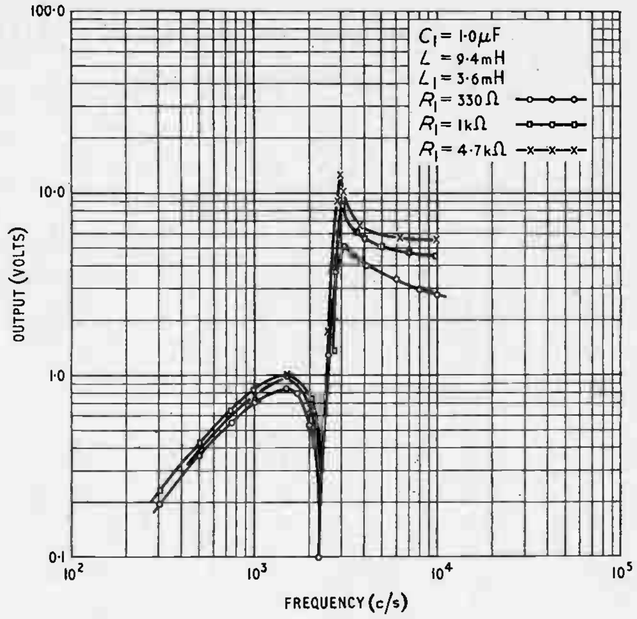


Fig. 6. *Frequency response of bifilar-T trap as function of* R_1

resonant frequency of the shunt arms is now given by

$$f_1' = \frac{1}{2\pi\sqrt{2L_1\left(\frac{C_1}{2} + C_s\right)}}$$

$$\text{thus } \frac{f_1'}{f_0} = \sqrt{\frac{\frac{L}{L - L_1}}{1 + 2\frac{C_s}{C_1}}}$$

$$\text{or } \frac{f_1'}{f_0} = \sqrt{\frac{\alpha}{(\alpha - 1)(1 + 2\beta)}}$$

where $\beta = C_s/C_1$. In the special case of $f_1' = f_0$,

$$C_s = \frac{1}{2(\alpha - 1)} C_1, \text{ or } \beta = \frac{1}{2(\alpha - 1)}.$$

Similarly, if a shunt inductor L_s is added at both input and output,

$$\frac{f_1'}{f_0} = \sqrt{\frac{\alpha}{\left(\alpha - 1\right)\left(1 + \frac{\gamma}{2}\right)}} \text{ where } \gamma = \frac{L_s}{L_1}.$$

Also, if both shunt inductors and shunt capacitors are added,

$$\frac{f_1'}{f_0} = \sqrt{\frac{\alpha}{\left(\alpha - 1\right)\left(1 + \frac{\gamma}{2}\right)\left(1 + 2\beta\right)}}.$$

If dissimilar capacitors or inductors are added at the input and output, stagger-tuning can be achieved. The design of such a network is exactly the same as that of a high-frequency staggered pair amplifier, and will not be discussed here.[2]

If a capacitor is added between the high-potential input and output terminals (i.e., across the parallel circuit of $-4LL_1/(L - L_1)$ and $-C_1/4$ in the equivalent circuit of Fig. 2), the notch frequency can be varied independently of the resonant frequency of the two shunt arms. If the added capacitance is C_t, then the notch frequency is changed to f_0' such that

$$\frac{f_0'}{f_0} = \sqrt{\frac{C_1}{C_1 - 4C_t}}.$$

Note that the addition of capacitance to this resonant circuit raises the resonant frequency. This is due to the negative capacitance in the equivalent circuit. Note also that the added capacitance C_t must not exceed $C_1/4$, otherwise the notch will not appear at a real frequency. Capacitive coupling between the two halves of the bifilar coil will have an effect similar to that produced by C_t; i.e., it will raise the notch frequency.

Similarly, an inductance L_t 'across the top' will change the notch frequency as follows:

$$\frac{f_0'}{f_0} = \sqrt{\frac{L_t}{L_t - \frac{4LL_1}{L - L_1}}}$$

L_t must not be smaller than $\frac{4LL_1}{L - L_1}$.

Design of Bifilar-T Traps

The usual design problem is to provide rejection at some frequency f_0, and to provide maximum gain at or in the region of some other frequency f_1. The capacitors should be much larger than any stray capacitance involved, unless stray capacitance can be held constant. One of the first choices which must be made is a value for $\alpha = L/L_1$. This choice does not appear to be critical. Values between 2 and 6 seem to be quite satisfactory but, presumably, higher values could be used if it is desired to have f_1 near (and above) f_0. (For $\alpha = 6, f_1/f_0 = 6/5$ with no additional shunt elements.) If it is desired to have f_1 lower than f_0, external shunt elements may be used, their magnitudes

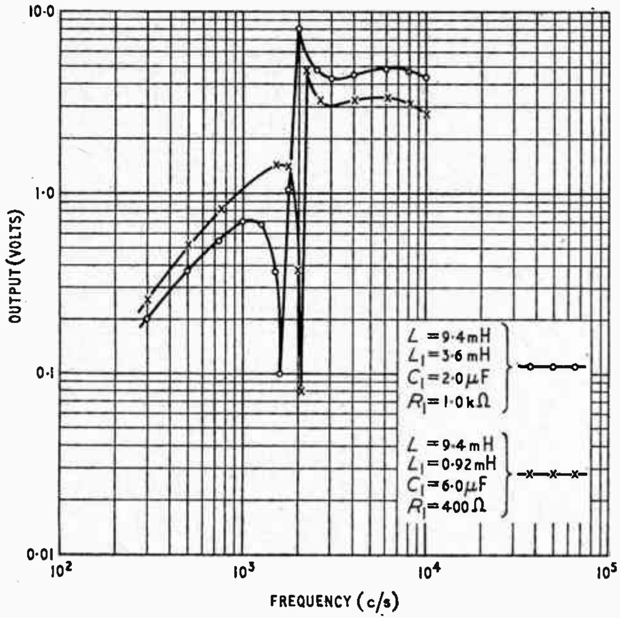


Fig. 7. *Frequency response of various bifilar-T traps*

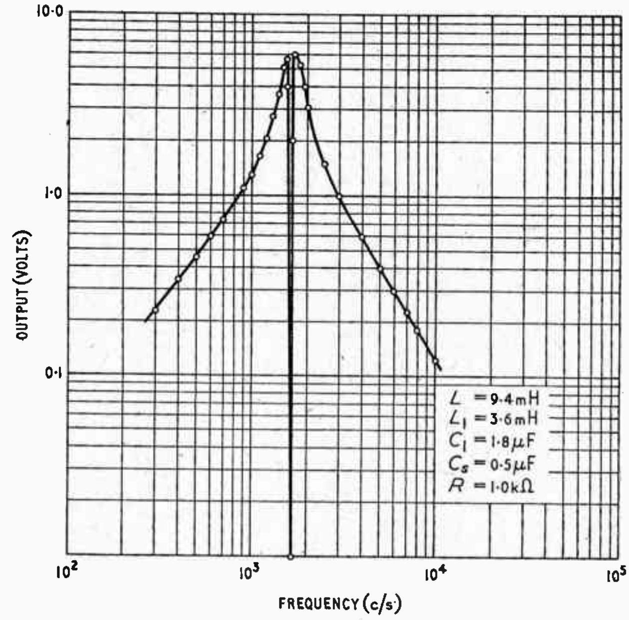


Fig. 8. *Symmetrical response trap*

calculated using formulae listed above. The impedance of the filter is approximately R_1 at f_1, and the upper limit of the value of R_1 is determined by the losses in coil L_1. It is therefore often advantageous to maintain $\omega L_1 Q_1$ as high as possible. (Q_1 is the Q of coil L_1.)

Several different circuits were tried to evaluate the performance of the trap at audio frequencies, and a specific circuit was designed to pass 1-kc/s signals, and reject 2-kc/s signals, for use in an infra-red detector amplifier. Initially, circuits were assembled using inductors on hand. Later, when some experience was gained in the use of the circuit, some specific designs were made. To illustrate the design of a trap, the calculations for the 1-kc/s pass, 2-kc/s rejection filter are outlined:

Inductors on hand were:

(a) Bifilar coil $L = 9 \cdot 4$ mH each leg

$(4L = 37 \cdot 5 \text{ mH})$

(b) Trap coil $L_1 = 3 \cdot 6$ mH.

thus $\alpha = L/L_1 = 2 \cdot 6$.

For rejection at $\omega_0 = 2\pi f_0$,

$$\omega_0^2 \frac{LL_1}{L - L_1} C_1 = 1$$

whence $C_1 = 1 \cdot 09\ \mu\text{F}$ for $f_0 = 2$ kc/s.

For $f_1' = 1$ kc/s, using shunt capacitors C_s

$$\frac{f_1'}{f_0} = \sqrt{\frac{\alpha}{(\alpha - 1)(1 + 2\beta)}}$$

whence $\beta = 2 \cdot 75$, and $C_s = 3 \mu\text{F}$.

No resistance was added across the coil L_1, thus R_1 was effectively $\omega_1 L_1 Q_1$, or approximately $1200\,\Omega$. A $0 - 10{,}000\ \Omega$ rheostat was connected across one half of the bifilar coil, and adjusted for maximum rejection at the notch frequency, at which setting its value was approximately 4700 Ω. (The remaining resistance shunting to reduce this to approximately 1200 Ω

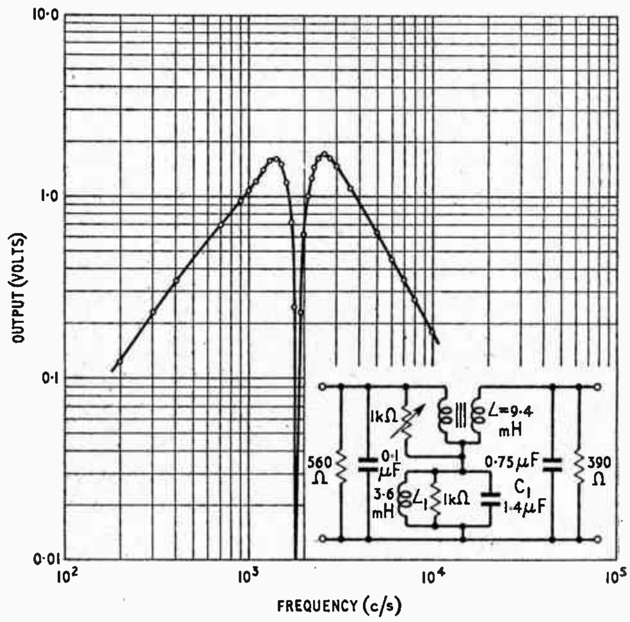


Fig. 9. *Frequency response of staggered-pair amplifier*

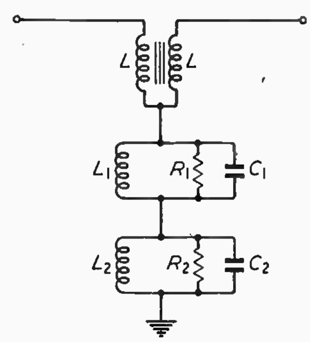


Fig. 10. *Two-trap bifilar-T circuit*

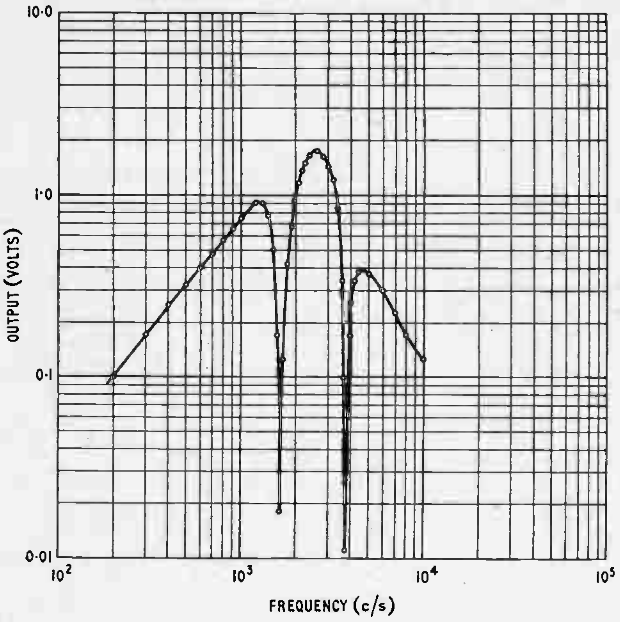


Fig. 11. *Frequency response of two-trap circuit*

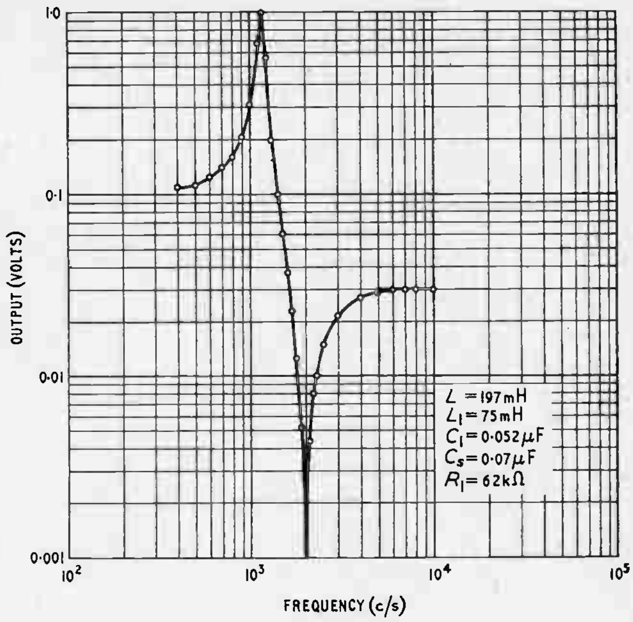


Fig. 12. *Frequency response of high-impedance trap*

presumably is provided by coil losses in the bifilar winding.)

Modification of the values calculated above was necessary in order to obtain satisfactory performance. The value of C_1 had to be increased to $1 \cdot 3\,\mu F$ to lower the notch frequency from 2300 c/s to 2000 c/s, the notch frequency being higher than calculated due to capacitive coupling in the bifilar coil. Also, C_s was reduced to $1 \cdot 5\,\mu F$ in order to obtain a satisfactory pass-band shape in the region of 1000 c/s. Fig. 5 gives the response curve of this trap.

Experimental Results

In addition to the trap designed specifically for the infra-red detector amplifier, the traps described below were constructed and tested.

(*a*) *Unmodified Traps.*

Several of these were tested. No external elements were added, so all are of the form illustrated in Fig. 1, and all exhibit peak responses above the notch frequency. Figs. 6 and 7 give response curves and component values.

(*b*) *Trap with* $f_0 = f_1$.

By suitable selection of components, the notch frequency and the resonant frequency of the shunt arms of Fig. 2 can be made equal. Since the bandwidth of the notch is a great deal less than that of the main response, a symmetrical, peaked response curve with a narrow rejection notch at mid-frequency should be obtained. (See Fig. 8 for a typical response curve.)

(*c*) *Staggered-Pair Amplifier.*

As mentioned, the resonant frequencies of the input and output shunt arms may be staggered by the addition of unequal shunting capacitors, using conventional design procedure. Fig. 9 gives the response curve and circuit of a staggered-pair amplifier.

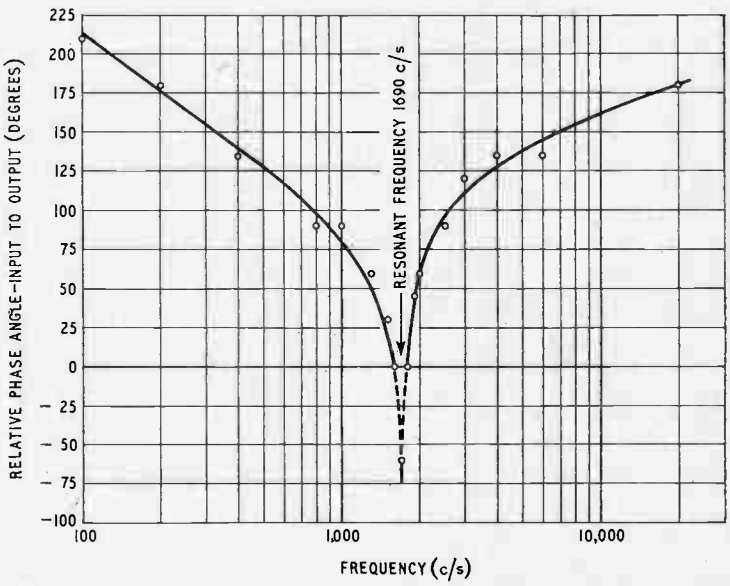
(*d*) *Two-Trap Circuit.*

A second trap in series with the first was added as indicated in Fig. 10. Two rejection notches were obtained, with high gain between the notch frequencies, and also fairly high gain both above the high frequency notch and below the low frequency notch. Symmetry is quite difficult to achieve with fixed-tuned coils. A typical response curve is given in Fig. 11.

(*e*) *High-Impedance Trap.*

All traps previously described have impedances of the order of 1000 Ω. A network of considerably higher impedance was constructed using the following

Fig. 13. *Approximate phase response of bifilar-T trap*



components:

$L = 197{\cdot}5$ mH

$(4L = 790$ mH)

$L_1 = 75$ mH.

$C_1 = {\cdot}05\ \mu F$

$C_s = {\cdot}07\ \mu F$

$R \approx 60$ kΩ

This network has an impedance of the order of 24 kΩ at 1000 c/s, and therefore introduces less loss when inserted in interstage circuits. The response curve of this network, when driven from a low impedance source, is given in Fig. 12.

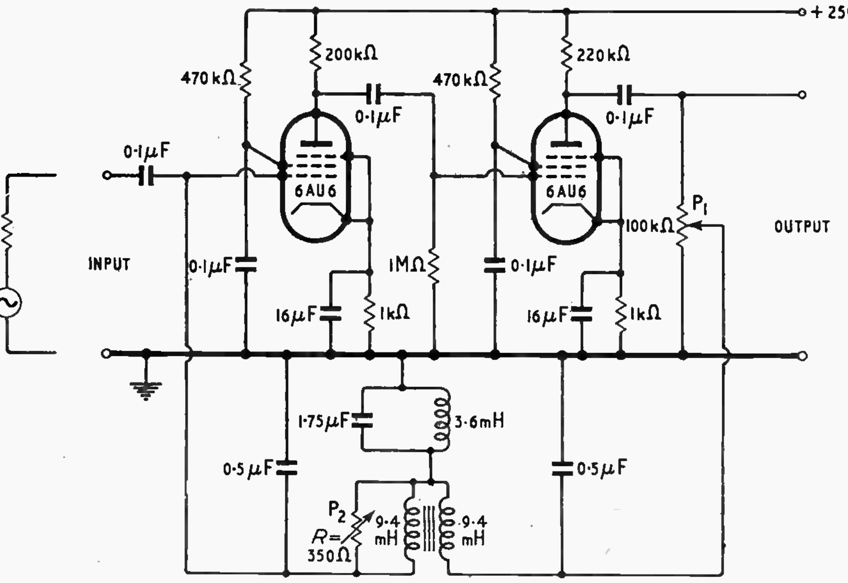


Fig. 14. *Narrow-band feedback amplifier*

Use of Traps in Feedback Amplifiers

The extremely high rejection and steep sides of the rejection notch suggest that an exceedingly narrow-band audio-frequency amplifier could be constructed by inserting the filter in the feedback path of a highly degenerative feedback amplifier. High gain would then be obtained only very near the notch frequency.

One of the traps with $f_0 = f_1$ described above was inserted in the feedback paths of several amplifiers. Considerable difficulty was experienced in obtaining a stable amplifier, however, because of the widely varying phase angle of the impedance of the trap. A very rough estimate of phase angle is given in Fig. 13. These points plotted were estimated using the Lissajous figures observed on an oscilloscope when the input and output voltages of the filter were applied to the horizontal and vertical deflection amplifiers, respectively.

However, one stable feedback amplifier was constructed, and a bandwidth between half-power points of approximately 30 c/s was obtained, with a centre frequency of 1700 c/s. The bandwidth of this amplifier 20 dB down from the peak was only 100 c/s wide, illustrating the extreme selectivity of the circuit. At frequencies one octave above and one octave below the centre frequency, the response was down by 35 dB. The circuit of this amplifier is given in Fig. 14, and the response curve obtained with it is shown in Fig. 15.

The sharpness of the frequency-response curve can be varied somewhat by suitable adjustment of potentiometers P_1 and P_2 (see Fig. 14). As the slider of P_1 is moved towards the output terminal end of the potentiometer, increasing the feedback, the gain of the amplifier decreases, and the bandwidth also decreases. The bandwidth may also be adjusted somewhat by variation of P_2. At any given setting of P_1, the bandwidth may be decreased by reducing R, the value of the resistance introduced by P_2. However, at very low, or zero values of R, the amplifier tends to oscillate, particularly at high feedback settings.

Conclusions

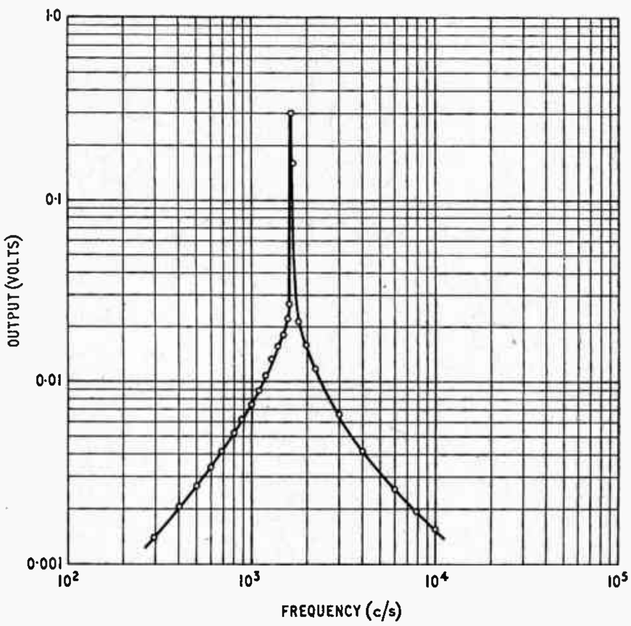
The suitability of the bifilar-T trap as a selective filter for some audio-frequency applications has been demonstrated. Equations and design procedures have been given to permit the design of filters such as those described herein with a minimum of 'cut and try'. A filter suitable for use in an infra-red detector amplifier has been designed. It is capable of attenuating the second harmonic of the 1,000 c/s signals used by approximately 70 dB. The performance of an extremely narrow (30 c/s) audio-frequency amplifier has been described.

REFERENCES

[1] "The Bifilar-T Trap". Editorials, *Wireless Engineer*, 1956, Vol. 33, Nos. 4 & 5, pp. 77 and 105.

[2] Valley and Wallman, "Vacuum Tube Amplifiers", M.I.T. Radiation Laboratory Series, Vol. 18, p. 176 *et seq.* McGraw-Hill, 1948.

Fig. 15. *Narrow-band feedback amplifier*



Far-Field Radiation of a Cheese Aerial

COMPARISON OF MEASURED PATTERN WITH THAT COMPUTED FROM APERTURE DISTRIBUTION

By R. F. Kyle*

SUMMARY. *This article describes the measurement of the aperture distributions of an X-band cheese aerial and the method of computing the radiation pattern from these figures. The computed pattern is compared with one measured in the field.*

The radiation pattern of a linear aerial can be obtained from the phase and amplitude distributions across the aperture by means of a Fourier transformation. If the distributions can be described by simple analytical expressions then, in many cases, the resultant integral in the transformation can be performed by elementary methods. In other cases where the phase is constant, and the amplitude can be represented as a combination of simple analytical functions, the resultant radiation pattern is obtained by adding together the radiation patterns produced by the simple functions. However, when the amplitude and phase vary in a complex way across the aperture, the radiation pattern can only be determined by numerical methods. This note describes work carried out on an X-band cheese aerial in which the phase and amplitude across the aperture were measured, and the radiation pattern was computed with the help of a high-speed digital computer. The resultant pattern was then compared with one obtained by measurement in the field.

* *Admiralty Signal and Radar Establishment*

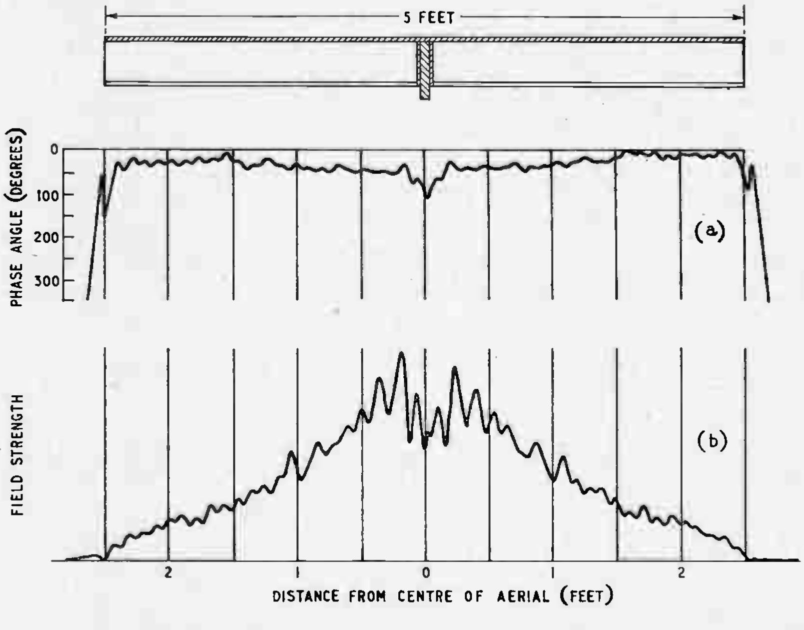


Fig. 1. *Phase (a) and amplitude (b) distributions across aperture of cheese aerial*

Phase and Amplitude Measurements

The aerial under test was a horizontally-polarized cheese with an aperture 5 ft. x 3 in., fed by a hoghorn placed at the focus. The aerial was mounted so that it looked into free space, and the phase and amplitude distribution across the aperture measured with an Elliott phase and amplitude plotter. The probe, which consists of a piece of circular waveguide one inch in diameter, was moved along the centre line of the aperture and approximately $2\frac{1}{2}$ inches in front of it. The plots are shown in Fig. 1, and it will be seen that there are rapid variations in amplitude, particularly in the neighbourhood of the central hoghorn. The effect is similar to the optical case of the diffraction pattern on each side of a narrow obstacle.

It is assumed that, with an aperture of many wavelengths, the distortion of the field caused by the probe is negligible and that the variations in phase and amplitude across the aperture are not so rapid that the integrating effect of the width of the probe causes an error. The degree to which these assumptions are valid can be deduced from the final comparison between the computed and measured radiation patterns. The chief precaution which must be taken in this experiment is to ensure that the probe runs parallel with the aperture. If it does not run parallel, a spurious phase shift will occur across the aperture, leading to a squint in the computed radiation pattern. The accuracy of the field amplitude as recorded is within 2% and the possible phase error is 2½°.

The vertical distributions across the height of the aperture (see below) were taken by mounting the cheese vertically and using the horizontal traverse for the probe as before, but the actual probe was modified to rectangular guide in order that a bend might be introduced to prevent the vertical part of the probe obstructing the aperture.

Transformation of Phase and Amplitude Distributions

The simplifications necessary to enable one to use the Fourier transformation of the aperture distribution to give the radiation pattern are dealt with in detail in textbooks. The chief practical points are that it is assumed that the field is zero at all points except in the aperture, and that the phase variations across the aperture are not very great. In the case of a linear aperture, such as a cheese aerial, it is also assumed that the vertical aperture distributions are similar at all points. This point was checked by taking a number of vertical cuts, and these are shown in Fig. 2. The three graphs were taken 2, 12, and 24 inches from the centre and show that the shapes are in fact only approximately the same.

The Fourier transformation which is required to give the radiation pattern $G(u)$ is

$$G(u) = \int_{-1}^{+1} \{A(x) \cos \phi(x) + jA(x) \sin \phi(x)\} \{\cos ux + j \sin ux\} \, dx$$

where $u = \dfrac{\pi a}{\lambda} \sin \theta$

$A(x)$ = amplitude distribution
$\phi(x)$ = phase distribution
a = length of aperture

Since the operation is performed on a digital machine the integral becomes a finite summation over the n points taken on the distribution

$$C(u) = \sum_n \{A_n(x) \cos \phi_n(x) \cos (ux_n) - A_n(x) \sin \phi(x) \sin (ux_n)\}$$

$$D(u) = \sum_n \{A_n(x) \cos \phi_n(x) \sin (ux_n) + A_n(x) \sin \phi_n(x) \cos (ux_n)\}$$

$$G(u) = \sqrt{C^2(u) + D^2(u)}$$

The number of points to be taken to give an accurate radiation pattern depends upon the frequency of the ripples in the distribution. In this particular case it was decided that 240 points would be adequate, as not only does this give at least 7 points in each wavelength of the ripple, but also the accuracy of the experiment and of reading the records does not justify any more points being taken. A radiation pattern was also computed using 100 points. These patterns are shown in Fig. 3 (b) and (c) where it will be seen that the main difference is in the wide-angle radiation.

The angles at which the radiation pattern is required must also be stated for the computation, and it is necessary to have some idea of the side-lobe structure if it is required to obtain an accurate radiation pattern with the minimum number of computed points.

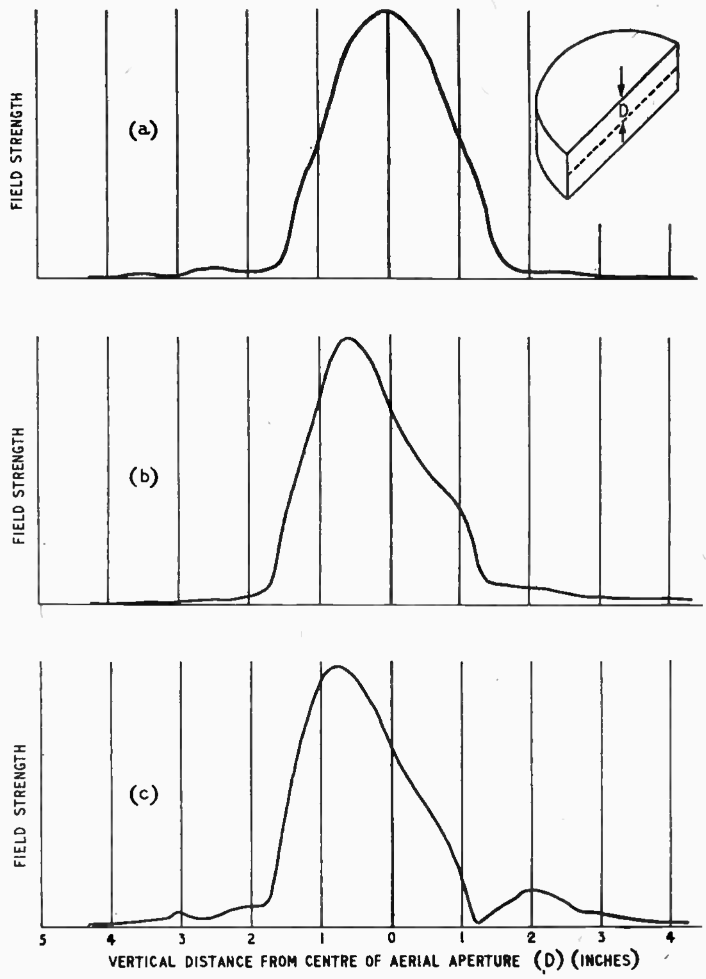


Fig. 2. *Vertical amplitude distributions in aperture of cheese aerial; (a) 2 in. from central feed, (b) 12 in. from central feed; (c) 24 in. from central feed*

As an alternative to decreasing the number of points taken on the distribution, 100 points were read from an amplitude plot which was smoothed out completely except for the central dip, and from which the phase ripples were also eliminated. The radiation pattern obtained in this way [shown in Fig. 3 (d)] can be seen to be very different indeed.

Measurement of Radiation Patterns

The experimental patterns were taken using a 5-ft paraboloid as transmitter on a 60-ft tower, and the cheese aerial on a 40-ft rotating tower at a range of 600 feet. This site should give a radiation pattern in which spurious reflections are negligible and, in fact, once again the agreement between the measured and computed patterns is a measure of the degree of perfection of the site. The patterns were taken using a crystal detector and logarithmic amplifier feeding a pen recorder. The receiving system was calibrated

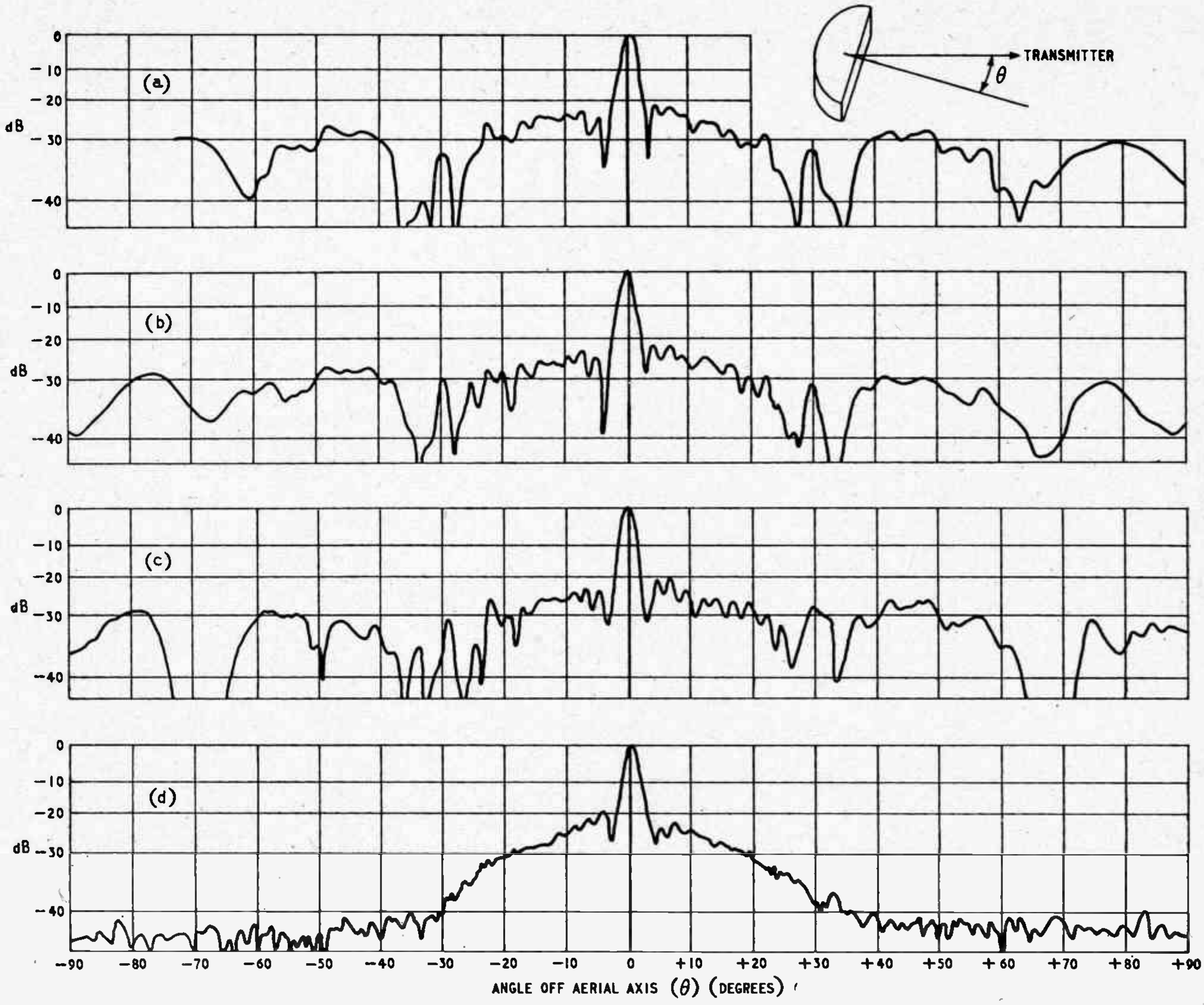


Fig. 3. *Measured and computed radiation of cheese aerial; (a) measured pattern, (b) pattern computed from* 240 *points, (c) pattern computed from* 100 *points, (d) pattern computed from smoothed phase and amplitude plot*

down to 40 dB with an r.f. attenuator after each run. Since a cheese aerial is very frequency-sensitive, it is very necessary to check that the field experiment is done at the same frequency as that at which the phase and amplitude plots were taken.

Comparison of Radiation Patterns

The measured radiation patterns are shown in Fig. 3(a) and can be compared with the computed patterns using 240 points shown in Fig. 3(b) and it can be seen that they are in very good agreement. For any practical purpose the diagrams are equivalent, and it is extremely difficult to give any figure which describes quantitatively the difference between the curves. It is felt that the correlation coefficient is not suitable in this context and that the most satisfactory figure is the average percentage difference between the curves. In this case, taking the figures at degree intervals gives a figure of $4\frac{1}{2}\%$.

Conclusions

These experiments demonstrate that if for any reason it is impracticable to measure a radiation pattern of a linear aerial, it is possible to compute one accurately from the phase and amplitude distributions. It must be emphasized that the case of a circular aperture would be much more difficult, as then a number of phase and amplitude plots would be required and the computation would be more complex.

The true purpose of the phase and amplitude plotter is not to eliminate the taking of radiation patterns of aerials but to assist in tracking down irregularities in such radiation patterns by detailed examination of the aperture distributions.

Acknowledgements

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Transmission-Line Discontinuities

EFFECT OF MULTIPLE REFLECTIONS

By K. W. H. Foulds, Ph.D.*

SUMMARY. *A detailed explanation is given of the effects of reflections from discontinuities in transmission lines. It is shown that under certain circumstances two identical discontinuities may not, in the steady state, upset the matching between a generator and a line. The manner in which the steady-state behaviour arises from the transient behaviour is described.*

Most problems concerned with matching transmission lines are readily solved by using the conventional transmission-line formulae or charts, given in any of the standard books (e.g., Ref. 1). Such methods show that the mismatch due to any obstacle or discontinuity in a transmission line can be cancelled out so far as the generator is concerned by placing a similar obstacle or discontinuity a specified distance away from the first. The physical processes which lead to this result are sometimes not fully appreciated by those using the formulae and charts and for some people the explanation is somewhat as follows. A certain fraction of the wave incident upon the first obstacle is reflected back towards the generator and the rest goes on towards the second obstacle where, in turn, part is reflected and the rest transmitted. Provided that the amplitude of the wave reflected from the first obstacle is small, the amplitude of the wave incident upon the second obstacle is not very different from that incident upon the first and, therefore, the amplitudes of the two reflected waves are also approximately the same. It is only necessary, therefore, to obtain the correct phase relation between the two reflected waves for them almost to cancel and produce an approximately matched transmission line.

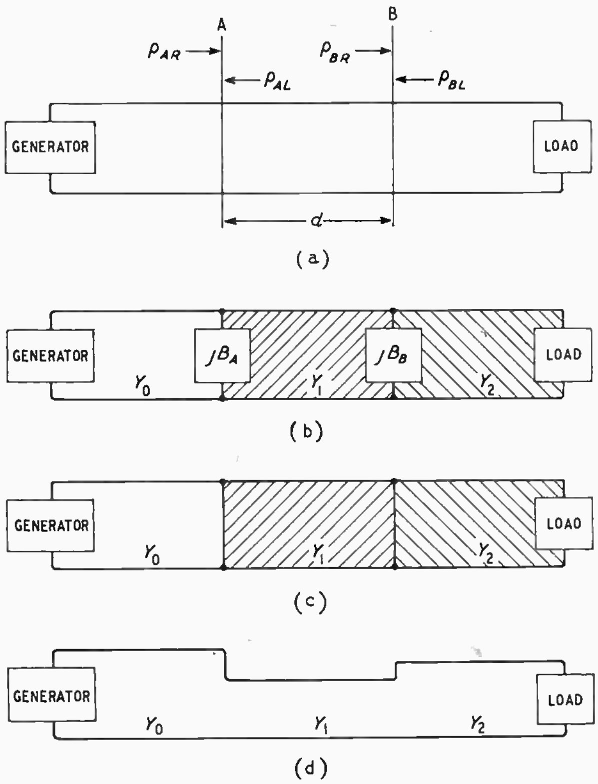
This incomplete reasoning is fairly satisfactory so long as the individual reflected waves are small, but it obviously leads to an absurd situation when the amplitude of the wave reflected from the first obstacle is greater than 50% of the amplitude of the incident wave. It is essential to remember that a fraction of the wave which is reflected from the second obstacle is subsequently reflected by the first obstacle back towards the second obstacle again, and thus a fraction of the wave is partially trapped between the two obstacles. The previous difficulty is resolved when one takes into account the effect of the multiple reflections of the partially-trapped wave.

It is interesting to examine in detail the way in which the fields build up to their steady-state values, and it is of course these steady-state values which correspond to the conventional transmission-line formulae and charts.

Multiple Reflections

Fig. 1 (a) represents a transmission line which is fed by a matched generator to the left and which is terminated in a matched load to the right. We imagine that at planes A and B there are discontinuities in the

Fig. 1. *Transmission lines with two discontinuities. (a) General form, (b) shunt susceptance and change of dielectric at reference planes, (c) change of dielectric at reference planes, (d) change of line dimensions at reference planes*



* *Imperial College of Science and Technology, London.*

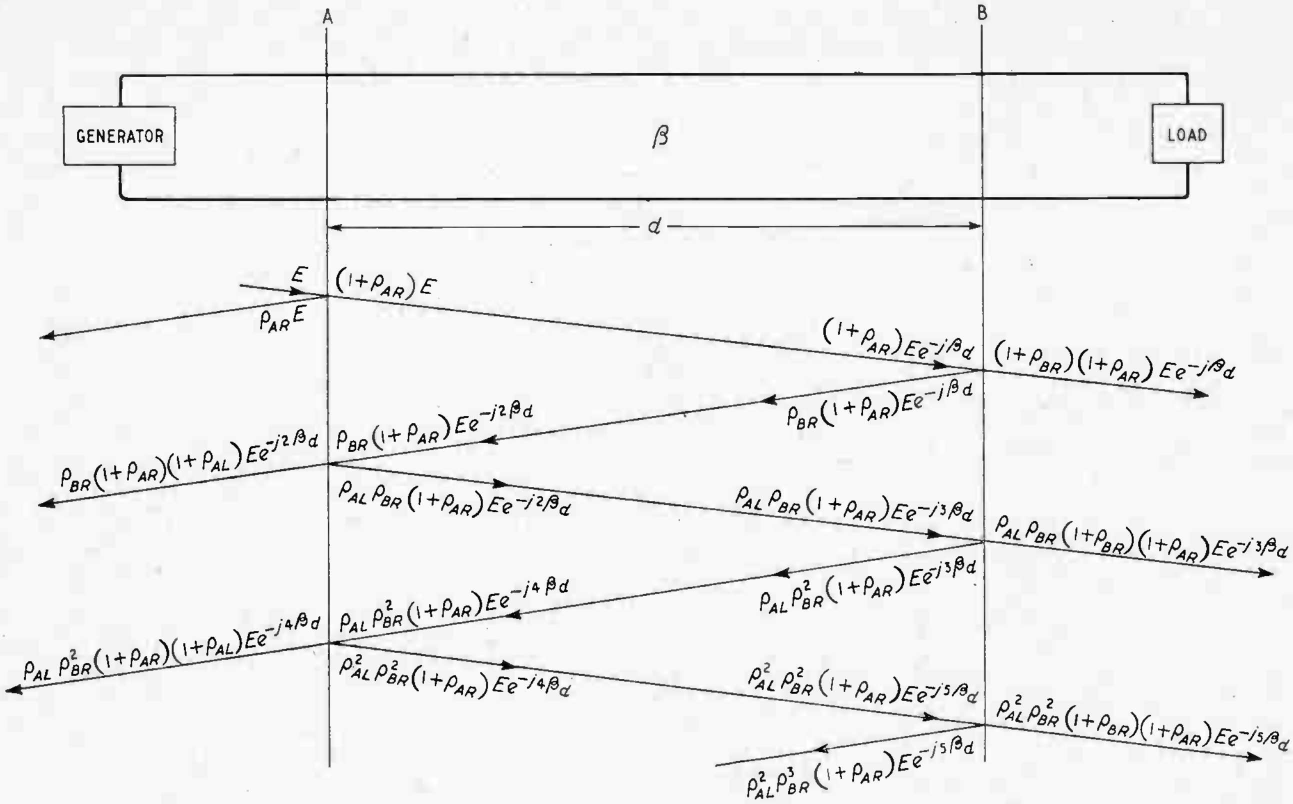


Fig. 2. *Values of incident, reflected and transmitted waves in a transmission line containing two discontinuities*

line such that the reflection coefficients for an incident wave travelling towards the right are ρ_{AR} and ρ_{BR} respectively and, for a wave travelling towards the left, the corresponding values are ρ_{AL} and ρ_{BL}. Fig. 1 (b), 1 (c) and 1 (d) show three possible transmission lines corresponding to the general form shown in Fig. 1 (a).

It is essential to make quite clear what is meant by the values of the reflection coefficients ρ_{AR}, ρ_{BR}, ρ_{AL} and ρ_{BL}. We suppose that the generator is switched on instantaneously at some time t and that the wave then travels towards the right, without reflection until it reaches the plane A. Since the wave-front has not yet travelled past this plane towards the right, the reflection coefficient ρ_{AR} will correspond either to a wave incident upon a load of the form $Y_1 + jB_A$, as in the case of Fig. 1 (b), or to a wave incident upon the interface between two infinite media, as in the cases of Fig. 1 (c) and 1 (d). Similar meanings are to be applied to the values of the other reflection coefficients.

Fig. 2 shows a wave of amplitude E incident upon the reference plane A together with the associated reflected and transmitted waves. The reflected wave $\rho_{AR}E$ travels back towards the generator, setting up a standing-wave pattern, and the transmitted wave $(1 + \rho_{AR})E$ travels to the right until it is partially reflected at plane B. The fraction of the wave transmitted past plane B then travels without further reflection to the load, but the reflected fraction propagates back to reference plane A, where it is again partially reflected. The fraction of this wave which is transmitted past reference plane A propagates back towards the generator, and this contribution may be of any phase with respect to the wave initially reflected from A depending upon ρ_{AR}, ρ_{BR}, ρ_{AL} and βd. Following this procedure, as shown in Fig. 2, it is clear that a fraction of the wave becomes partially trapped between the two reference planes of discontinuity. Whenever the wave-front of this partially-trapped wave is incident upon plane A a further fraction of the wave propagates back towards the generator, and whenever it is incident upon plane B a further fraction is transmitted towards the load.

The first few reflections at the two planes are shown in Fig. 2.

We are interested in the build-up of the field-pattern in the various sections of the line and it is clear that the modulus values of the field at any point change with each transit of the partially-trapped wave. However, before examining in detail this build-up of the field-pattern it is interesting and instructive to obtain the final steady-state values and compare them with those obtained from the standard transmission-line equations and charts. Let us calculate, for example, the conditions necessary for the transmission line to be matched to the generator.

The final reflected electric field in the region between the generator and the first discontinuity is given by the sum of all the individual waves propagating back towards the generator as described in the previous paragraphs. Thus the final reflected electric field in this section is given by

$$R = \rho_{AR}E + \rho_{BR}(1 + \rho_{AR})(1 + \rho_{AL})Ee^{-j2\beta d} + \rho^2{}_{BR}\rho_{AL}(1 + \rho_{AR})(1 + \rho_{AL})Ee^{-j4\beta d} + \ldots$$

and after simplifying, this becomes

$$R = E\left[\rho_{AR} + \frac{\rho_{BR}(1 + \rho_{AR})(1 + \rho_{AL})e^{-j2\beta d}}{1 - \rho_{AL} \cdot \rho_{BR}e^{-j2\beta d}}\right] \quad (1)$$

In order to make the calculations simple let us consider these relations for the simple transmission lines in Fig. 3 (a) and 3 (b).

Dielectric Slab of Thickness *d*

Fig. 3 (a) shows a transmission line of characteristic admittance Y_0, which is real, in which the region between the reference planes A and B is filled with a different dielectric such that the characteristic admittance of this section is now Y_1; this is also real. The propagation constant in this section is β.

Then $\rho_{AR} = \rho_{BL} = \dfrac{Y_0 - Y_1}{Y_0 + Y_1}$

and $\rho_{BR} = \rho_{AL} = \dfrac{Y_1 - Y_0}{Y_1 + Y_0}$

Thus $\rho_{AR} = \rho_{BL} = -\rho_{BR} = -\rho_{AL} = \rho$ say, and this is real. Substituting these values into Equ. (1) leads to

$$R = \rho\left[\frac{1 - e^{-j2\beta d}}{1 - \rho^2 e^{-j2\beta d}}\right]E$$

After rationalizing, this can be expressed in the form

$$R = \frac{2\rho \sin \beta d}{\sqrt{(1 - \rho^2)^2 + (2\rho \sin \beta d)^2}}E$$

The line is matched; i.e., the reflected field is zero, when $\sin \beta d = 0$. This occurs when the dielectric slab is an integral number of half-wavelengths long, the wavelength being that measured in the dielectric.

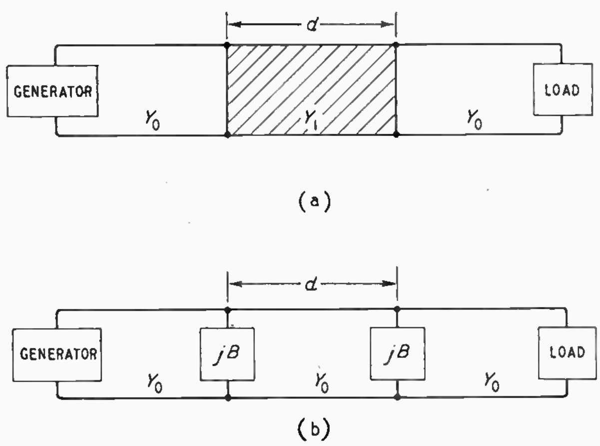
If we deal with the same problem using the conventional transmission-line formulae we know that the input admittance of the line immediately to the load side of plane A is given by

$$Y = Y_1\left[\frac{Y_0 + jY_1 \tan \beta d}{Y_1 + jY_0 \tan \beta d}\right] \quad \cdots \quad (2)$$

and thus the line is matched when $\tan \beta d = 0$; i.e., when the line is again an integral number of half-wavelengths long.

It is worth while pointing out that, although most electrical engineers are more familiar with the transmission-line presentation, most physics books dealing with the topic of half-wavelength slabs, or slices, adopt the reflection method[2].

Fig. 3. *Simple examples of transmission line with two discontinuities; (a) change of dielectric at the reference planes, (b) identical shunt susceptance at the reference planes*



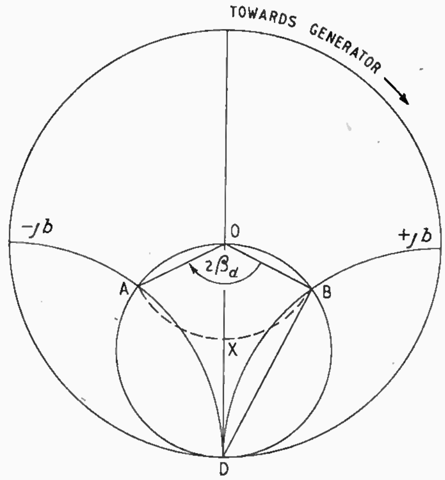


Fig. 4. *Simplified Smith circle admittance diagram*

Two Loss-Free Obstacles Spaced a Distance *d*

Fig. 3 (b) shows a simple transmission line of characteristic admittance Y_0 in which identical loss-free obstacles of reactance jB are placed at the reference planes A and B. The admittances of both the generator and the load are assumed to be Y_0.

As far as an incident wave-front is concerned, the admittance immediately before each reference plane, irrespective of the direction of propagation, is given by $Y = Y_0 + jB$.

Then $\rho_{AR} = \rho_{BR} = \rho_{BL} = \rho_{AL} = \dfrac{Y_0 - Y}{Y_0 + Y}$ and this is complex.

Substitute the value $\rho_{AR} = \rho_{BR} = \rho_{BL} = \rho_{AL} = |\rho|e^{j\phi}$ in Equ. (1) and this leads to the value of the reflected field in the region between the generator and the first discontinuity.

$$R = |\rho|e^{j\phi}[1 + e^{-j2\beta d} + 2|\rho|e^{-j(2\beta d - \phi)}]E$$

For a matched line $R = 0$ and this leads to

$$\beta d = \cos^{-1}|\rho|$$

and $$\phi = \cos^{-1} - |\rho|$$

(It is convenient to observe here that therefore $\phi - \beta d = \pi$.) This shows that two similar obstacles will set up no reflected waves so far as the generator is concerned provided that the electrical length between them βd is given by $\cos^{-1}|\rho|$.

This same result can of course be obtained from the transmission formula, Equ. (2), but in fact it can be deduced much more conveniently using the Smith circle diagram shown in Fig. 4.

The normalized admittance immediately on the generator side of plane B looking towards the load is $1 + jb$ (writing $B/Y_0 = b$), and this corresponds to the point B which lies on the unit conductance circle with the radius OB $= |\rho|$. If we travel towards the generator, i.e., clockwise along the circumference BXA, until the

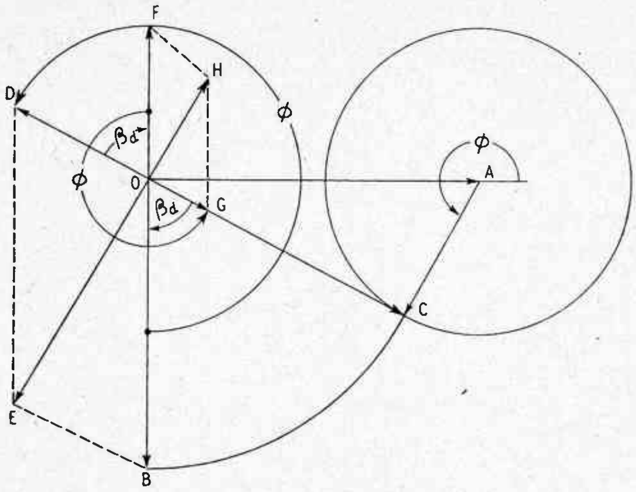


Fig. 5. *Vector diagram of the incident, reflected and transmitted waves*

admittance again lies on the unit conductance circle at A, the normalized admittance at A is $1 - jb$. Thus if we choose this position in the line as our reference plane A and place a susceptance jb there, the input admittance immediately on the generator side of this plane is $(1 - jb) + jb$; i.e., is matched. The angle through which the radius OB has to be rotated is equal to $2\beta d$, and from the triangle OBD we see that $\cos \beta d = OB/OD$. Bearing in mind that $OB = |\rho|$, and OD corresponds to unity reflection coefficient we have $\cos \beta d = |\rho|$ and this, of course, is the same result as before.

Transient Conditions

Let us now consider the way in which the steady-state conditions, which we have just discussed, are obtained. We shall deal with a transmission line containing two loss-free obstacles which have been placed so that under steady-state conditions there is no reflected wave fed back to the generator. The transient analysis is most conveniently carried out by drawing a vector diagram based on the crank diagram principle[1]. This vector diagram, drawn for the specific case of $|\rho| = 0{\cdot}5$, is shown in Fig. 5.

OA represents the original wave incident upon the first obstacle at reference plane A. The first reflected field from this plane is $|\rho|.OAe^{j\phi}$; i.e., is AC, and thus the total field at plane A and, therefore, the field which is subsequently transmitted towards the reference plane B is OC. The field incident upon B at the first transit corresponds to OC retarded through βd and, if we draw the diagram corresponding to final steady-state matched conditions, OB is at right angles to OA because $\phi - \beta d = \pi$. The reflected wave at plane B corresponds to $|\rho|$OB advanced through ϕ and is thus OD. The total field at plane B is the vector sum of OD and OB, and this is OE, the wave which is then transmitted to the load. From the geometry of the diagram

$$\frac{DE}{DO} = \frac{1}{|\rho|} = \frac{OA}{CA} \text{ and } \angle EDO = \angle CAO = \phi - \pi,$$

therefore $\angle DOE = \angle OCA$ and OE is parallel to AC.

The wave reflected from plane B travels towards plane A and is there represented by OD retarded through βd; i.e., OF. This in turn sets up the reflected wave corresponding to OG, which is $|\rho|$OF advanced through ϕ and which clearly lies along OC. The vector sum of these two fields $OF + OG = OH$, which is parallel to AC but is opposite in sense. The vector OH represents the field which is transmitted backwards past plane A towards the generator and will contribute to the standing-wave pattern in the region between generator and plane A. The diagram shows, however, that this contribution nullifies a fraction $(1 - |\rho|^2)$ of the wave which was reflected directly from the obstacle at plane A. The diagram also makes it clear that the wave trapped between the two obstacles which is now making its second transit from planes A to B; i.e., OG,

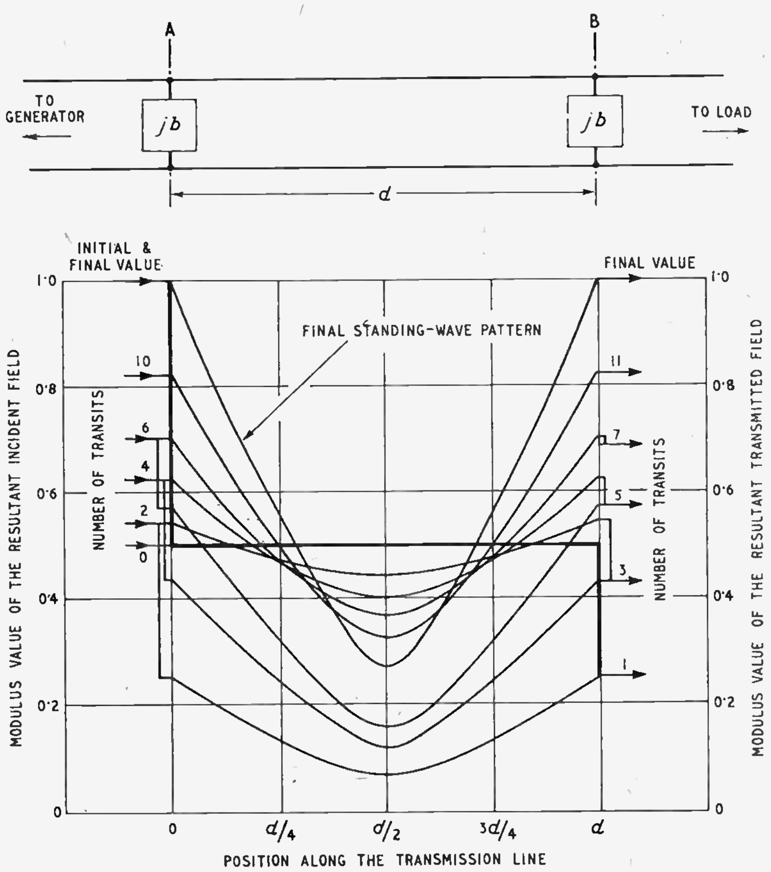


Fig. 6. *Modulus values of the resultant electric field at different positions along the transmission line shown as a function of the number of transits made by the wave partially trapped between the two discontinuities*

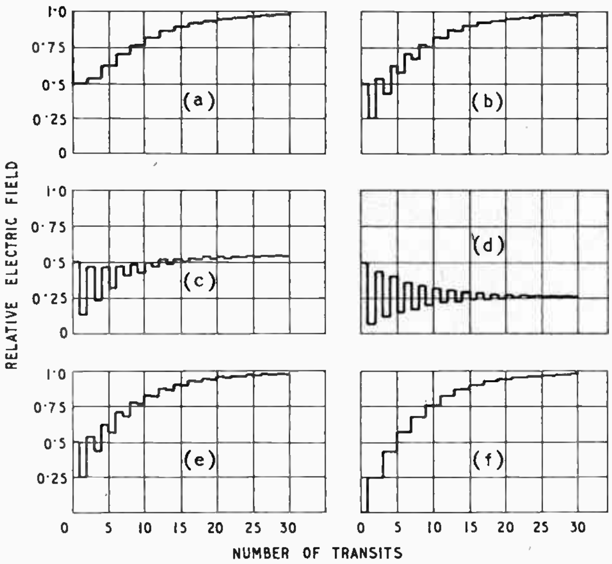


Fig. 7. *Step-by-step variation in the modulus value of the resultant electric field for successive transits of the partially-trapped wave, shown for various positions along the line; (a) immediately to the generator side of discontinuity A, (b) immediately to the load side of discontinuity A, (c) at the 1/4 and 3/4 positions between the two discontinuities, (d) midway between the two discontinuities, (e) immediately to the generator side of discontinuity B (f) immediately to the load side of discontinuity B*

has the same phase as it had on the first transit though, of course, the amplitude is reduced by a factor $|\rho|^2$.

Repeating the argument exactly for this wave leads to the result that the contributions to the field which are transmitted through plane B to the load all have the same phase, and on the vector diagram lie along OE. Similarly the contribution to the field transmitted backwards past plane A towards the generator all have the same phase and in the diagram lie along OH, and that each contribution cancels out a little more of the wave originally reflected from plane A until finally the reflected wave is reduced to zero. At the same time the resultant field transmitted to the load becomes equal to the uniform field which now exists between the generator and the first obstacle. There will, however, be a standing wave existing between the two obstacles.

The multiple transit diagram of Fig. 2 enables the value of the electric field at any point in the transmission line to be calculated at any time during the initial building-up period. Such calculations have been made for a reflection coefficient whose modulus value is 0·86603. The value $|\rho| = 0{\cdot}86603$ implies that 75% of the incident power is reflected at the first incidence upon the first obstacle and 25% of the power is transmitted past plane A towards plane B.

Fig. 6 shows the resultant electric-field distribution in the transmission line corresponding to the various transits of the partially-trapped wave which is reflected from one obstacle back to the other. In order to demonstrate what is happening more clearly, Fig. 7 shows the step-by-step change in the resultant electric field which occurs at seven positions in the transmission line. This shows convincingly the gradual approach to the final steady-state values.

REFERENCES

1 "Microwave Measurements," H. E. M. Barlow and A. L. Cullen. (Constable.)
2 "A Treatise of Light," by R. A. Houston. Longmans (London)

LOW-NOISE AMPLIFIER USING SEMICONDUCTOR DIODES

A new device which has interesting possibilities as a low-noise u.h.f. and microwave amplifier is under development at Bell Telephone Laboratories. Although still in the experimental stage, preliminary results indicate that this device, which uses semiconductor diodes as the active elements, can improve the performance of many types of microwave receivers. It is relatively simple to construct and operate, and shows prospects of having a long life.

Commercially available microwave amplifiers and converters add a considerable amount of noise to the incoming signal. A major reduction in this added noise can significantly improve the performance of receivers such as those used in radar, radio astronomy, over-the-horizon radio relays and u.h.f. television systems. The diode amplifier holds promise of providing such a noise reduction.

This amplifier is one of the family of devices known as parametric amplifiers in which a variable reactance or 'varactor' serves as the active component. In the present device, the variable reactance is provided by a semiconductor diode whose capacitance varies with the applied voltage. As with other varactor amplifiers, the applied voltage is derived from a high-frequency pump signal. This signal causes the diode to function as a time-varying capacitance and supplies the energy which is necessary to produce amplification.

Low-noise amplification using varactor diodes was predicted by A. Uhlir, Jr., of Bell Laboratories, and amplication at 6,000 Mc/s was first demonstrated by M. E. Hines and H. E. Elder. At the same frequency, G. F. Herrmann and M. Uenohara later obtained a bandwidth of 8 Mc/s with a noise figure of 5–6 dB. The gain was 18 dB and the pump signal 12,000 Mc/s. Gain can be exchanged for additional bandwidth if desired, and vice versa.

A travelling-wave amplifier configuration using arrays of several diodes shows promise of providing bandwidths of 25% or more in the u.h.f. region. Using four stages with the special diodes in such an array. R. S. Engelbrecht of Bell Laboratories has obtained a bandwidth of 100 Mc/s at a 400 Mc/s signal frequency, with a pump frequency of 900 Mc/s and a pump power of 10 mW. This experimental amplifier has a gain of 10 dB and a noise figure of only 3·5 dB.

A single type of diode can be used to make an amplifier for any desired frequency from the high microwave region down to d.c. The noise performance improves rapidly as the frequency decreases from microwaves down into the u.h.f. region, thus making such an amplifier potentially useful for u.h.f. television receivers.

Interesting aspects of the diode amplifier are its simplicity and potential reliability. The major components are the waveguide structures, the diode itself, and a suitable pump signal source. It appears that these components can be assembled to provide a relatively inexpensive device. No refrigeration is required, and no magnetic fields are necessary. The low-noise characteristics are realizable at room temperatures.

Although the variable capacitance effect is present in commercial diodes, special diffused silicon diodes have been developed which maximize this effect. Series resistance, which could be a source of noise, is minimized in these diodes.

INTERNATIONAL CONVENTION ON TRANSISTORS

The Institution of Electrical Engineers, Savoy Place, London, W.C.2, is organizing an International Convention on Transistors and associated semiconductor devices, to be held in London from 25th to 29th May 1959. Further details will be announced.

A Mathematician's Privilege of Working Backwards

There are occasions when laziness is enlightened. Sometimes useful discoveries have been made simply because the discoverer could not be bothered to do a job by the hard ways previously available, and was therefore obliged to invent a new and better way. Before undertaking a computation, it is worth examining the problem briefly to see if it can be tackled in an indirect way not initially obvious, especially if we wish to consider the effect of variations in one quantity upon those of a related quantity.

In this article we shall consider two particular examples, but the mathematical ideas and techniques involved are very generally applicable. The first example is concerned with the behaviour of a particular feedback amplifier of the kind discussed in general terms in "Mathematical Tools" of August 1957, while the second example is concerned with the problem of group-delay correction.

In the August 1957 "Mathematical Tools", it was shown that the stability of the four-stage feedback amplifier of Fig. 1 was associated with the equation

$$A_0 = \frac{A_L}{(1+pC_1R_1)(1+pC_2R_2)(1+pC_3R_3)(1+pC_4R_4)+A_L\beta} \qquad (1)$$

where p is as usual the Heaviside operator, A_0 is the overall gain, A_L is the gain at low frequencies, and the R and C elements are respectively the values of anode a.c. resistance and coupling resistance in parallel, and of total shunt capacitance, for each stage of the amplifier. β has the value $R_6/(R_5 + R_6)$ in Fig. 1. For the amplifier to be stable, the denominator of the right-hand side of Equ. (1) must be zero for four values of p, none of which has a positive real part, and the July 1957 "Mathematical Tools" was concerned with the conditions for an algebraic equation to be free from such roots. Here we shall for simplicity suppose that

$$C_1R_1=C_2R_2=C_3R_3=\alpha;\ C_4R_4=20\alpha;\ A_L\beta=\beta' \qquad (2)$$

The denominator of (1) then reduces to

$$D = (1+\alpha p)^3\,(1+20\alpha p)+\beta' \qquad (3)$$

It must be possible to express D in the form of two real quadratic factors, so that

$$D = 20(\alpha^2p^2+\lambda\,.\,\alpha p+\mu)\,(\alpha^2p^2+\nu\,.\,\alpha p+\rho) \qquad (4)$$

and we here consider how to determine the variation of λ, μ, ν and ρ for a given variation in β'. For it is the quantities λ, μ, ν, ρ which give the most important information about the behaviour of the amplifier. If λ is much less than $2\sqrt{\mu}$, for instance, the first factor in Equ. (4) will give rise to a lightly-damped oscillation, whereas, if $\lambda = 2\sqrt{\mu}$, the damping will be critical, and if λ exceeds $2\sqrt{\mu}$ there will be no oscillation due to this first factor in Equ. (4). Similar considerations apply independently to the second factor in Equ. (4). If λ or ν is negative, Equ. (4) will be associated with an oscillation which grows exponentially; that is to say, there will be instability.

If D is given by Equ. (3) and has the factors given by Equ. (4), the following relations between λ, μ, ν, ρ and β' can be obtained by equating coefficients of p^3, p^2, p and the constant term respectively:—

$$\lambda+\nu = 3{\cdot}05 \qquad (5)$$

$$\mu+\rho+\lambda\nu = 3{\cdot}15 \qquad (6)$$

$$\lambda\rho+\mu\nu = 1{\cdot}15 \qquad (7)$$

$$\mu\rho = 0{\cdot}05\,(1+\beta') \qquad (8)$$

Now in fact we are given β' and required to determine

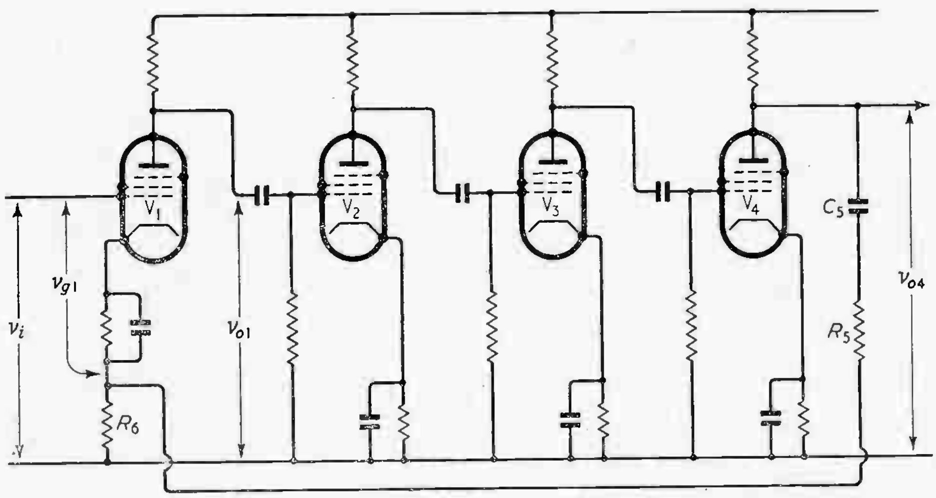


Fig. 1. *Four-stage feedback amplifier*

λ, μ, ν and ρ. It is perfectly possible to attack the problem directly, and factorize the right-hand side of Equ. (3), as indicated in the first two "Mathematical Tools" (February and March 1957). What we wish to emphasize here, however, is that this is not necessary; it is preferable, for a general exploration of the way in which λ, μ, ν, ρ vary with β', to proceed as if λ was given and not β'. For if λ is given, ν follows immediately from Equ. (5), and Equs. (6) and (7) then become linear simultaneous equations for μ and ρ; finally, substitution into Equ. (8) gives β' immediately. If we choose a number of values of λ and deduce the corresponding values of β', μ, ν, and ρ in this way, it is perfectly possible afterwards to plot λ and μ against β' as in Fig. 2, and ρ against β' as in Fig. 3. The curious behaviour of μ in the neighbourhood of $\beta' = 0$ is not of practical importance; it is associated with the fact that in Equ. (3) there is a factor $(1 + \alpha p)^3$ on the right-hand side. ν is of course deducible from λ by means of Equ. (5), so that ν has not been included in Fig. 3.

Once Figs. 2 and 3 have been obtained, we can in effect factorize (to graphical accuracy) the right-hand side of Equ. (3) at sight; it therefore seems worth while to extend Figs. 2 and 3 to cover all values of β', and not merely those between -1 and $19{\cdot}91$ for which the amplifier is stable. If β' is less than -1, stability breaks down because μ becomes negative; if β' exceeds $19{\cdot}91$, stability breaks down because λ becomes negative. For large λ, we have approximately from Equs. (5)–(8)

$$\nu \approx -\lambda;\ \mu \approx \rho \approx \tfrac{1}{2}\lambda^2\ \beta' \approx 5\lambda^4 \qquad (9)$$

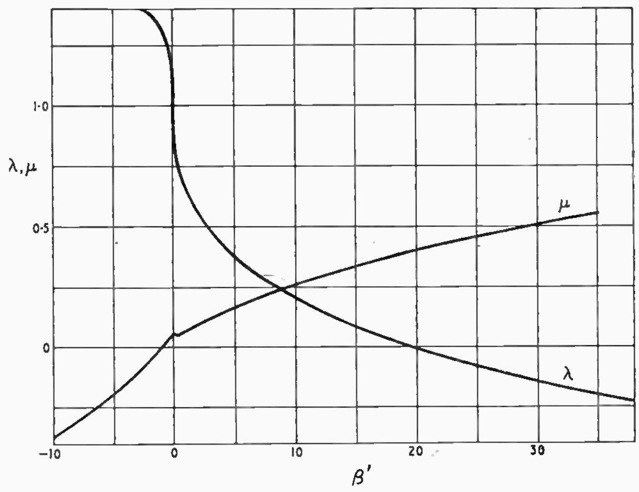


Fig. 2. *Plot of λ and μ against the feedback parameter β so that Equ.* (3) *has the factors* (4)

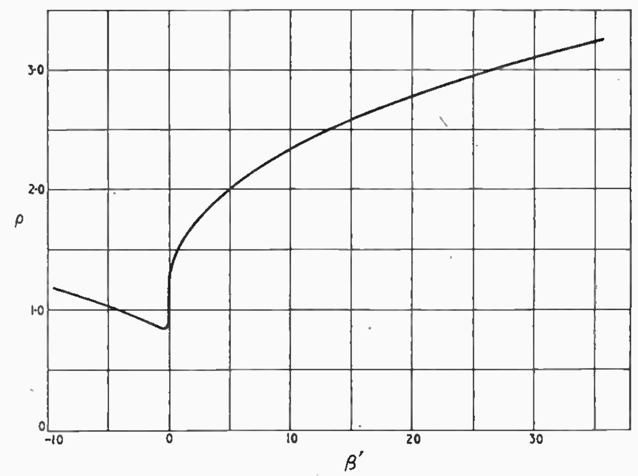


Fig. 3. *Plot of ρ against the feedback parameter β' so that Equ.* (3) *has the factors* (4)

Our second example is concerned with the practical realization of a network whose group-delay shall be 'quartically flat'. Suppose that we have a network whose transfer function is $1/f(p)$ where

$$f(p) = 1+x_0(p/\omega_0)+Kx_0^2(p/\omega_0)^2 + (K-\tfrac{1}{3})x_0^3(p/\omega_0)^3+\tfrac{1}{3}(K-0{\cdot}4)x_0^4(p/\omega_0)^4 \qquad (10)$$

Then the group delay G, which is the real part of $f'(p)/f(p)$, with p^2 replaced by $-\omega^2$ at frequency $\omega/2\pi$, is given by

$$G = \frac{x_0}{\omega_0}\cdot\frac{1+(1-2K)\xi^2+\frac{1}{15}(15K^2-20K+6)\xi^4+\frac{1}{3}(K-\frac{1}{3})(K-0{\cdot}4)\xi^6}{1+(1-2K)\xi^2+\frac{1}{15}(15K^2-20K+6)\xi^4+(\frac{1}{3}K^2-0{\cdot}4K+\frac{1}{9})\xi^6+\frac{1}{9}(K-0{\cdot}4)^2\xi^8} \qquad (11)$$

where $\xi = x_0\omega/\omega_0$. Gouriet (reference 1) discovered that if $f(p)$ was given by Equ. 10, the group delay G was such that if G is expanded in ascending powers of ξ^2 (or ω^2), there is no term in ω^2 or ω^4 whatever the values of x_0 and K may be. If a lattice network has transfer function $f(-p)/f(p)$, the gain will be unity at all frequencies, and the group delay $2G$, so that such a lattice can sometimes be used to correct the group delay of a network which is reasonably satisfactory at low frequencies, but not at frequencies near $\omega_0/2\pi$. The way in which G varies with K is discussed by Gouriet in reference 1. x_0 is merely a scale factor.

Equation (10) is associated with a realizable network if $K>0{\cdot}4$ but, in order to realize it, we require the factors of the right-hand side of Equ. (10) ; reference 1 gives the technique for realization when a lattice network with transfer function $\phi(-p)/\phi(p)$ is used, $\phi(p)$ being quadratic. If K is between $0{\cdot}4$ and $3/7$, G always decreases as ω increases; for $K>3/7$, G at first increases, passes through a maximum, and thereafter steadily decreases. The rise is very gradual for K up to about $0{\cdot}45$, but steep for K exceeding say $0{\cdot}5$.

From the general behaviour of the network whose group-delay requires correction, we can suppose that K and x_0 in Equ. (10) are approximately known (or the values of each K and x_0 if several correcting lattices are needed). We then have to determine a, b and b' so that Equ. (10) can also be written

$$f(p) = \left(1+a\frac{px_0}{\omega_0}+b\frac{p^2x_0^2}{\omega_0^2}\right)\left(1+[1-a]\frac{px_0}{\omega_0}+b'\frac{p^2x_0^2}{\omega_0^2}\right) \qquad (12)$$

As in the amplifier example above, we are in fact given K and require a, b and b' but, as we wish to explore the variation of a, b and b' for a wide variation of K, it is better to proceed as if a was given, and the other quantities required. By equating coefficients of p^2, p^3 and p^4 in Equs. (10) and (11), we obtain

$$b + b' + a(1 - a) = K \quad \ldots \quad (13)$$

$$ab' + (1 - a)b = K + \tfrac{1}{3} \quad \ldots \quad (14)$$

$$bb' = \tfrac{1}{3}(K - 0{\cdot}4) \quad \ldots \quad (15)$$

and if a is regarded as given, Equs. (13) and (14) become linear simultaneous equations for b and b' in terms of a, K. Their solution is easily found to be

$$(1 - 2a)b = K(1 - a) + a^2(1 - a) - \tfrac{1}{3} \quad \ldots \quad (16)$$

$$(1 - 2a)b' = \tfrac{1}{3} - aK - a(1 - a)^2 \quad \ldots \quad (17)$$

Multiplying Equs. (16) and (17), and substituting for bb' from Equ. (15), we obtain a relation between K and a which is only quadratic in K, and reduces to

$$CK^2 - \tfrac{1}{3}CK(1+6C) + (C^3 + 0{\cdot}2C - [1/45]) = 0 \quad (18)$$

where $C = a(1 - a)$; the solution of (18) is

$$K = C + \tfrac{1}{6}\left\{1 \pm \left[(1 - 4C)\left(\frac{0{\cdot}8}{C} - 3\right)\right]^{1/2}\right\} \ldots \quad (19)$$

Thus, given a, we immediately deduce C; K is then derived from Equ. (19) and finally b and b' from Equs. (16) and (17) and, in order to obtain a set of corresponding values of a, b, b' and K, the most difficult operation required has been the solution of the quadratic equation (19). The minus sign in this equation is only relevant for values of K between $0{\cdot}4$ and $5/12$ since, if $K < 0{\cdot}4$, the network given by Equ. (10) is unrealizable. For $K = 0{\cdot}4$, a is $0{\cdot}43055$ and C is $0{\cdot}24518$. a and b are plotted against K in Fig. 4, and b' is plotted against K in Fig. 5. In this example, there seems little point in allowing K to vary outside the limits $0{\cdot}4$ to 1; K tends to infinity as a tends to 1, so that if $(1-a)$ is denoted by ϵ and ϵ is small (say less than $0{\cdot}01$)

$$K \approx \tfrac{1}{6}\left\{\left(\frac{0{\cdot}8}{\epsilon}\right)^{1/2} + 1 - 3{\cdot}375(0{\cdot}8\epsilon)^{1/2}\right\} + \epsilon \quad (20)$$

the first term neglected being of order $\epsilon^{3/2}$.

Now Equ. (18) is a very useful result, easily obtained by the present method of attack. Not only does it give the relation between K and a explicitly, but it incorporates the fact that if, in Equ. (11), a is replaced by $(1-a)$ and b is replaced by b' and vice versa, the same factors are present and only the order is reversed. If we had factorized the right-hand side of Equ. (10) directly, not only would we have taken longer to obtain each set of associated values of K, a, b and b', but we would have been most unlikely to derive the true relation between a and K; we would probably have become aware of the fact that replacing a by $(1-a)$ left K unchanged, but no more.

It is always worth bearing in mind that an indirect attack on a problem involving computation, by regarding as independent a variable different from that which is independent in the practical problem, is worth considering. In the case of the amplifier example, this method has enabled us to see clearly how variations in feedback affect damping, and the way in which instability arises (λ becomes negative if β' exceeds $19{\cdot}91$ and μ becomes negative if β' is less than -1). In the case of the group-delay correcting network, the method has given us easily and explicitly a rather complicated relation (Equ. 18) between the quantities K and a.

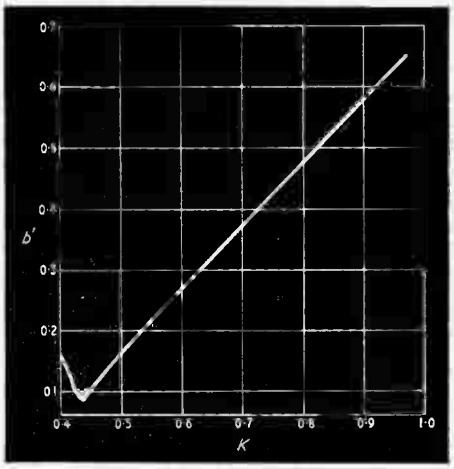


Fig. 5. *Values of b′ in Equ.* (12) *required to factorize Equ.* (10) *for given K*

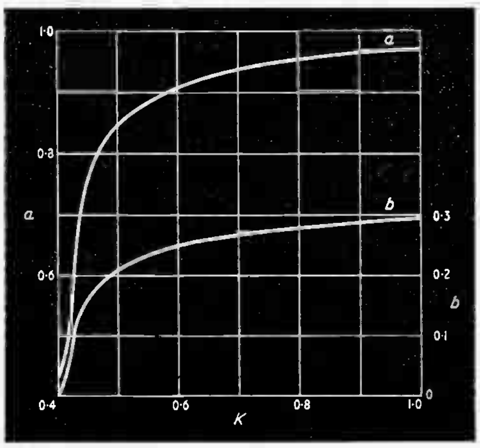


Fig. 4. *Values of a and b in Equ.* (12) *required to factorize Equ.* (10) *for given K*

REFERENCE

[1] G. G. Gouriet, "Two Theorems Concerning Group Delay with Practical Application to Delay Correction", I.E.E. Monograph 275R, December 1957.

BIRTHDAY HONOURS

In the Queen's Birthday Honours List, Dr. Willis Jackson, director of research and education at Metropolitan-Vickers, is created a Knight Bachelor.

P. A. T. Bevan, chief engineer of I.T.A., and E. K. Cole, chairman and managing director of Ekco, are appointed Commanders of the Order of the British Empire.

A new member of the Order of the British Empire is H. V. Griffiths, engineer in charge of B.B.C.'s receiving station at Tatsfield.

THE USE OF SEMICONDUCTOR DEVICES

A 36-page booklet under this title has recently been published by the B.V.A., and is being distributed by contributing manufacturers. It covers diodes, junction transistors and photosensitive devices.

Noise Performance of a Three-Stage Microwave Receiver

By H. V. Shurmer, M.Sc., Ph.D., A.M.I.E.E.*

SUMMARY. *Analysis of the noise performance of microwave receivers has hitherto been confined to cases where the input signal is fed directly into a crystal mixer or detector followed by an intermediate frequency or video amplifier, respectively. This article extends the treatment to include a receiver in which the crystal valve is preceded by a stage of r.f. amplification, such as a travelling-wave tube.*

The present interest in r.f. amplifiers for use in microwave receivers makes it desirable to extend accordingly the established treatments applicable to the noise performance of two-stage receivers.

Microwave receivers are divided into two main classes—heterodyne receivers employing a crystal mixer and video receivers which use a crystal video detector. The noise performance of the former type is described in terms of an 'overall noise figure', and that of the latter by a 'figure of merit'.

Both types of receiver are considered as networks comprised of three stages. The first stage is in each case an r.f. amplifier. With the heterodyne receiver the second and third stages are a crystal mixer and an i.f. amplifier respectively; with the video receiver they are, respectively, a crystal video detector and a video amplifier. In both cases the output meter is considered as part of the third stage.

Heterodyne Receivers

Let F_1, G_1, etc., represent the noise figure and available power gain, respectively, of each stage and let $F_{(1+2+3)}$, $G_{(1+2+3)}$ represent the corresponding values for the complete network. The available output noise power of the network, $\delta N_{0(1+2+3)}$ over an incremental frequency bandwidth, δf, is given by

$$\delta N_{0(1+2+3)} = F_{(1+2+3)}\, G_{(1+2+3)}\, kT\delta f \quad \ldots \quad (1)$$

where

k = Boltzmann's constant.

T = temperature (°K).

($kT = 4\times10^{-21}$ joule for $T = 290$°K)

Now

$$G_{(1+2+3)} = G_1\, G_2\, G_3,$$

so that

$$\delta N_{0(1+2+3)} = F_{(1+2+3)}\, G_1\, G_2\, G_3\, kT\delta f \quad \ldots \quad (2)$$

This available output-noise power may be divided into the components arising from each of the three stages :—

The component from stage 3 originating in the signal source and stage 1 is

$$\delta N_0'{}_{(1+2+3)} = G_2\, G_3\, \delta N_{01} = F_1\, G_1\, G_2\, G_3\, kT\delta f \quad (3)$$

The component from stage 3 originating in stage 2 is

$$\delta N_0''{}_{(1+2+3)} = G_3\, \delta N_{02} - G_2\, G_3\, kT\delta f = F_2\, G_2\, G_3\, kT\delta f - G_2\, G_3\, kT\delta f \quad \ldots \quad (4)$$

The term subtracted represents the noise from the output resistance of stage 1 which is included in Equ. (3).

Similarly, the component from stage 3 originating in stage 3 is

$$\delta N_0'''{}_{(1+2+3)} = F_3\, G_3\, kT\delta f - G_3\, kT\delta f \quad \ldots \quad (5)$$

Substitution in the equation

$$\delta N_{0(1+2+3)} = \delta N_0'{}_{(1+2+3)} + \delta N_0''{}_{(1+2+3)} + \delta N_0'''{}_{(1+2+3)} \quad \ldots \quad \ldots \quad (6)$$

leads to

$$F_{(1+2+3)} = F_1 + \frac{F_2-1}{G_1} + \frac{F_3-1}{G_1 G_2} \quad \ldots \quad \ldots \quad (7)$$

If N_0 is the available noise output power and N_0' is the noise output power which would be available if

* The British Thomson-Houston Co. Ltd., Rugby.

LIST OF SYMBOLS

F	=	noise figure.
G	=	available power gain.
δN	=	available noise power in an incremental frequency bandwidth.
δf	=	incremental frequency bandwidth.
k	=	Boltzmann's constant.
T	=	temperature (°A).
N	=	available noise power over finite bandwidth.
F'	=	effective noise figure over finite bandwidth.
t	=	crystal mixer or detector noise temperature ratio.
L	=	crystal mixer conversion loss.
B_1, etc.	=	stage noise bandwidth.
P_1, etc.	=	contribution to receiver output power from particular causes.
P_s	=	available power of signal at receiver input.
P_L	=	particular value of P_s corresponding to tangential sensitivity.
I	=	instantaneous current appearing at crystal detector output under short-circuit conditions.
V	=	instantaneous applied voltage at crystal detector.
N_D	=	spectral noise density at input to crystal detector.
S	=	signal voltage amplitude at input to crystal detector.
f	=	frequency.
ψ	=	phase angle.
β	=	current sensitivity of crystal detector.
R	=	crystal video resistance.

there were no noise sources within the network, the effective overall noise figure $F'_{(1+2+3)}$ is given by

$$F'_{(1+2+3)} = \frac{N_0}{N_0'}$$

$$= \frac{\int_0^\infty F_{(1+2+3)}\, G_1\, G_2\, G_3\, df}{\int_0^\infty G_1\, G_2\, G_3\, df} \quad .. \quad .. \quad (8)$$

Suppose stage 1 represents an r.f. amplifier, stage 2 a crystal mixer and stage 3 an i.f. amplifier–meter combination. The bandwidth of stage 3 will then be small compared with that of stages 1 and 2, so that G_1, G_2, F_1 and F_2 may be considered constant over the range of integration. Equ. (8) then simplifies to

$$F'_{(1+2+3)} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3' - 1}{G_1\, G_2} \quad .. \quad .. \quad (9)$$

For a crystal mixer, it is customary to write the product $FG=t$, the 'noise temperature ratio'. Instead of referring to the gain of a crystal mixer it is usual to consider its reciprocal, L, the conversion loss. Thus, in Equ. (9)

$$t = F_2\, G_2$$

$$L = \frac{1}{G_2}$$

which gives

$$F'_{(1+2+3)} = \left(F_1 - \frac{1}{G_1}\right) + \frac{L}{G_1}\,(F_3' + t - 1) \quad .. \quad (10)$$

Substitution of $F_1 = G_1 = 1$ into Equ. (10) gives the well known formula for a two-stage heterodyne receiver :—

Effective overall noise figure $= L(F'_{i.f.} + t - 1)$ (11)

where $F'_{i.f.}$ is the effective noise figure of the i.f. amplifier–meter combination and corresponds to F'_3 in Equ. (10).

Equ. (10) shows that if the gain of the r.f. amplifier is large, the overall noise figure of the receiver is dominated by F_1. If G_1 exceeds 20 dB the other terms may usually be neglected.

Video Receivers

Consider now a receiver in which a signal of pulsed microwave power is amplified at r.f. before detection by a crystal rectifier, the resulting spectrum of the pulse envelope feeding a video amplifier. The method of evaluating noise performance employed for heterodyne receivers is inapplicable, since detection is a square-law phenomenon and the detector gain is a function of input power level. The noise performance of a video receiver is assessed in terms of the input-signal power level which gives a total power output double that due to noise in the absence of a signal, assuming matched conditions throughout. This input signal is said to correspond to 'tangential sensitivity' and is denoted by P_L.

In the absence of a signal, the receiver output is due to rectification of those noise components transmitted by the video amplifier. These can arise directly as video-frequency noise within the crystal detector and video amplifier or result from mixing by the crystal of noise components transmitted by the r.f. amplifier.

When a signal is applied, the receiver output increases, partly as a result of detection of the signal and also through the mixing of signal and noise by the crystal to produce components within the pass-band of the video amplifier.

The analysis of this type of receiver involves the separation of the input into components which arise in these various ways and expressing each component in terms of the constants of the system. The components of available power present in the absence of the input signal are equated to the additional ones arising from the application of a signal and, hence, an expression is found for the tangential sensitivity.

The following treatment employs elementary mathematical methods, but the results have been verified by more rigorous means such as may be found in standard texts on noise theory[2,3]. It is assumed for the present purpose that the gain and noise figure of the r.f. amplifier may be considered constant over the bandwidth.

Video-Frequency Noise Generated within Crystal and Video Amplifier.

The contribution, P_1, of crystal and video amplifier to the receiver-noise output is conveniently represented by an expression similar to that which applies to the heterodyne receiver, viz.:

$$P_1 = G_2\,(F_2' + t - 1)\,kTB_2, \quad .. \quad .. \quad .. \quad (12)$$

where G_2, F_2' and B_2 represent, respectively, the gain, noise figure and bandwidth of the video amplifier and t is the crystal noise-temperature ratio. It should be noted that at the low power levels at which the crystal operates in this type of receiver t will be very nearly unity unless d.c. bias is applied.

Noise at Video Frequency due to Mixing of R.F. Noise Components with Each Other Within the Crystal Detector.

To evaluate this contribution P_2, consider the process of detection occuring at the crystal rectifier. Assuming a square-law response, the instantaneous short-circuit current is related to the input by a voltage relationship

$$I = kV^2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (13)$$

where k is a constant of proportionality.

Assume that the spectral-noise density is uniform and let the voltage at the detector input associated with unit frequency interval be N_D. If S is the signal voltage amplitude at the detector input the total applied voltage may be written:

$$V = N_D \sum_m \cos 2\pi\,(f_m t + \psi_m) + S \cos 2\pi f_0 t\,, \quad .. \quad (14)$$

where f_m represents the component frequencies of noise and f_0 is the signal frequency.

Combining Equs (13) and (14) gives

$$I = k\, N_D^2\, [\sum_m \cos\, 2\pi\,\, (f_{mt} + \psi_m)]^2$$
$$+\, k\, S^2 \cos^2 2\pi f_0 t$$
$$+\, 2\, k\, N_D\, S \sum_m \cos 2\pi\,(f_m t + \psi_m) \cos 2\pi f_0 t \quad .. \quad (15)$$

Equ. (15) expresses the current flowing in the output circuit of the detector in terms of the input microwave frequency components.

Because of its restricted bandwidth, the video amplifier filters out all high-frequency components, so that only those terms which arise as difference frequencies in the expansion of Equ. (15) may contribute to the receiver output.

There are two such sets of terms, one arising from the mixing of noise components with each other and the second from the mixing of noise components with the signal. Denoting these sets of terms by i_1 and i_2, respectively,

$$i_1 = \tfrac{1}{2} k\, N_D^2 \sum_m \sum_n \cos 2\pi\, [(f_m - f_n)\, t + \psi_m - \psi_n] \quad (16)$$

$$i_2 = k\, N_D\, S \sum_m \cos 2\pi\, [(f_m - f_0)\, t + \psi_m] \quad \ldots \quad (17)$$

Equ. (16) is the one which relates to the mixing of noise components with each other.

Let B_1 represent the bandwidth of the r.f. amplifier and consider a frequency $(f_m \approx f_n)$, within the pass-band B_2 of the video amplifier. In Equ. (16) the range of f_m, f_n will be the r.f. amplifier bandwidth, so that, assuming $B_1 \gg B_2$ there will be $2B_1$ terms of each frequency $(f_m \approx f_n)$. There are B_2 such frequencies transmitted by the video amplifier, so that the instantaneous mean-square current associated with all the transmitted components is

$$i_1{}^2{}_{r.m.s.} = \frac{k^2\, N_D^4\, B_1 B_2}{4} \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

The contribution to the available power at the receiver output is given by

$$P_2 = \frac{G_2\, i_1{}^2{}_{r.m.s.}\, R}{4} = \frac{G_2 k^2\, N_D^4\, B_1 B_2\, R}{16} \quad \ldots \quad (19)$$

where R is the video resistance of the crystal detector.

The constants k and N may be eliminated from Equ. (19) by considering the term in the expansion of Equ. (15) which relates to the short-circuit current due to the noise input; viz.: $k\, N_D^2\, B_1/2$.

This rectified current is also given in terms of the crystal current sensitivity by $\beta\, G_1\, F_1\, kTB_1$, where β is the current sensitivity of the crystal detector and G_1, F_1 are respectively the gain and noise figure of the r.f. amplifier.

Equating the two expressions for rectified current gives

$$kN_D^2 = 2\beta G_1 F_1 kT \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

Substituting Equ. (20) into Equ. (19) gives the required expression for P_2, viz:

$$P = \frac{G_2(\beta G_1 F_1 kT)^2 B_1 B_2 R}{4} \quad \ldots \quad \ldots \quad (21)$$

Noise at Video Frequency due to Mixing of R.F. Noise Components with the Signal.

Referring to Equ. (17), it is noted that there are two terms of each of the difference frequencies $(f_m \approx f_0)$ and that the video amplifier transmits B_2 such frequencies. The mean-square current under short-circuit conditions. is thus given by

$$i_2{}^2{}_{r.m.s.} = k^2\, N_D^2 S^2\, B_2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

Equating the short-circuit current due to the signal, expressed in terms of current sensitivity, to the corresponding term in Equ. (15) gives

$$k\, S^2 = 2\, G_1\, \beta\, P_s\,, \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (23)$$

where P_s is the available signal power at the input to the r.f. amplifier.

Substituting Equs (20) and (23) into Equ. (22) gives

$$i_2{}^2{}_{r.m.s.} = 4\, (\beta\, G_1)^2\, F_1\, kT\, B_2\, P_s \quad \ldots \quad \ldots \quad (24)$$

The contribution, P_3 to the available power output of the receiver, arising from the mixing within the crystal detector of r.f. noise components with the signal, is thus given by:

$$P_3 = \frac{G_2\, i_2{}^2{}_{r.m.s.}\, R}{4} = (\beta\, G_1)^2\, F_1\, kT\, B_2\, G_2\, R\, P_s \quad (25)$$

Direct Detection of Signal Power

The r.f. signal itself appears at the receiver output by virtue of its being modulated or pulsed at a suitable video frequency, so that the bulk of the modulation envelope is transmitted through the video amplifier. Assuming all the available signal power to be transmitted, the resulting contribution to the receiver output is given by

$$P_4 = \frac{G_2\, (\beta\, G_1\, P_s)^2\, R}{4} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (26)$$

For the condition of tangential sensitivity

$$\left.\begin{array}{l} P_s = P_L \\ \text{and} \quad P_1 + P_2 = P_3 + P_4 \end{array}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (27)$$

Substituting for P_1, etc. from Equs (12), (21), (25) and (26) and solving the resulting quadratic in P_L gives

$$P_L = -2F_1 kT B_2 \left\{ 1 \pm \sqrt{1 + \frac{B_1}{4B_2} + \frac{(F_2' + t - 1)}{(\beta F_1 G_1)^2\, kT\, B_2 R}} \right\} \quad \ldots \quad (28)$$

If $F_1 = G_1 = 1$, only the last term in Equ. (28) is significant. This equation then reduces to the well-known formula for a two-stage crystal-video receiver:

$$P_L = \frac{1}{\beta\sqrt{R}} \sqrt{4kT\, B_2\, (F_2' + t - 1)} \quad \ldots \quad \ldots \quad (29)$$

where $\beta\sqrt{R}$ is known as the 'figure of merit'.

Provided that the last term in Equ. (28) is dominant, an increase of F_1 has no significant effect but the effective figure of merit of the receiver is improved in proportion to G_1. This applies only so long as the product $F_1\, G_1$ is sufficiently small and there is little advantage in increasing G_1 beyond the point at which, say, the third term is reduced to the magnitude of the second term. Using this criterion, the maximum value above which it ceases to be worthwhile to increase the product $F_1\, G_1$ is given by

$$\frac{B_1}{4B_2} = \frac{(F_2' + t - 1)}{(\beta\, F_1\, G_1)^2\, kT\, B_2\, R}$$

$$\text{or } F_1\, G_1 = \frac{1}{\beta} \frac{4(F_2' + t - 1)}{kT\, B_1\, R} \quad \ldots \quad \ldots \quad \ldots \quad (30)$$

Typical values for the above quantities are

$\beta = 1{\cdot}0\ \mu\text{A}/\mu\text{W}$, $F_1 = 1{\cdot}6$, $t = 1{\cdot}0$,
$B_1 = 5 \times 10^9$ c/s, $R = 4000\ \Omega$ and
$kT = 4 \times 10^{-21}$ joule.

Substitution of these values into Equ. (30) gives $F_1\, G_1 = 39{\cdot}5$ dB. When this figure is reached, little further improvement in the effective figure of merit can be achieved by further increase of G_1.

Acknowledgments

The author wishes to thank his colleagues at the British Thomson-Houston Co. Research Laboratory for assistance and encouragement in this work and to make acknowledgment to the Admiralty for permission to publish.

REFERENCES

[1] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers", McGraw-Hill Book Co., 1948.
[2] J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals", McGraw-Hill Book Co., 1950.
[3] A. van der Ziel, "Noise", Chapman and Hall, 1955.

Noise-Reducing Speech Transmission System

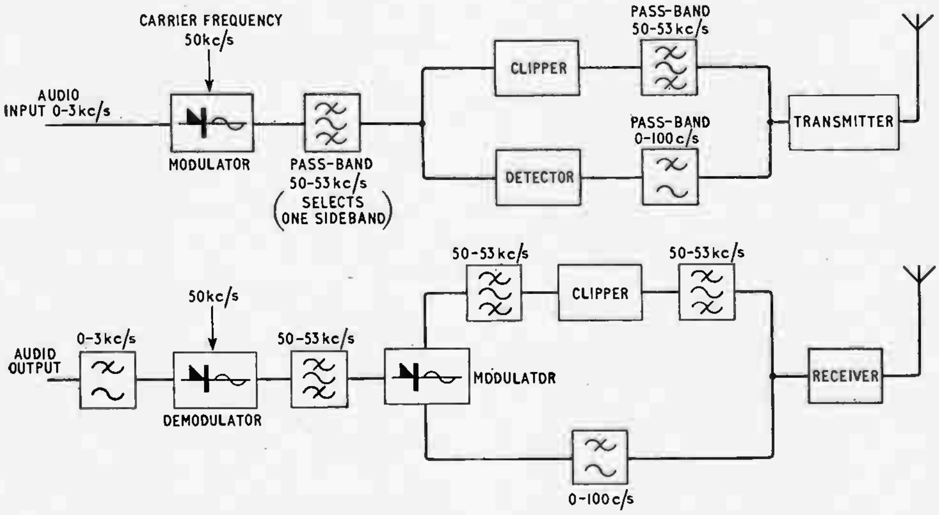
The results of the work of researchers into the characteristics of speech have been applied to a speech transmission system with the object of reducing the effects of noise. An important discovery was that 'infinitely clipped' speech remains intelligible despite the large amount of distortion introduced by the clipping process. Unfortunately, infinitely clipped speech is so unpleasant to listen to that it is not suitable for ordinary communication systems. One reason for its unpleasantness is that most of the harmonics generated by clipping lie within the audio band. They add nothing to intelligibility, because the information in the clipped waveform is contained in the fundamental frequency.

A first step in reducing unpleasantness is therefore to get rid of these harmonics. In the new system, a single-sideband signal at, say, 50 kc/s is first produced, using unclipped speech with 'pre-emphasis' of the higher frequencies. This signal is then clipped, the harmonics now being removed by means of a filter. The output of the filter, which consists of a narrow-band frequency-varying signal during speech, and no signal during the intervals of speech, is transmitted. It is possible to obtain from this signal at the receiver something like an infinitely clipped audio signal (i.e., one which contains information about the original audio frequencies, but no information about amplitude variations). But, in practice, the received signals, in addition to being unpleasant to listen to, are not very intelligible. The reason for this is that, in a noisy channel, the intervals of speech are occupied by noise. The listener has, therefore, constantly to distinguish between noise and speech signals. This process is fatiguing and leads to errors of judgment.

An important feature of the proposed new transmission system is an auxiliary narrow-band (100 c/s) channel, which is used to send information about the amplitude of the original speech sounds, without regard to their frequency. The output from this channel is employed to modulate that of the frequency channel. When there is no speech, the latter is muted.

Below is a block diagram of the system, which has been named *Frena* (from '*fre*quency *a*nd *a*mplitude'). The minimum required signal-to-noise ratio for acceptable intelligibility is found to depend on the nature of the modulation employed in the auxiliary channel. If a.m. is used, the improvement over a normal single-sideband transmission is 2 dB. Frequency modulation yields an improvement of 4 dB. The greatest improvement (6 dB) is achieved by transmitting in the auxiliary channel only the information that a signal is present or not present in the frequency channel. The speech quality is degraded somewhat, but a great simplification of equipment can be effected by transmitting carrier when there is no signal in the frequency channel and no carrier when there is a signal. Thus, the transmitter sends a constant-amplitude signal all the time, the frequency being varied during speech. Under these conditions, the transmitter can operate continuously at maximum efficiency.

The system outlined above is described in *Philips Technical Review*, Vol. 19, No. 3.



Essentials of new transmission system. When on/off amplitude information only is employed, the output of the 0–100-c/s filter at the transmitter is used to control the transmission of carrier frequency and the output of the corresponding filter at the receiver controls a muting device

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Trigger Circuits with Multiple Stable States

SIR,—When a voltage source has a rising load characteristic, like a series d.c. generator, it possesses negative internal resistance, which may be exploited in trigger circuit action. Regenerative changes in voltage and current occur when the resistance of the load is numerically less than the negative internal resistance. It is possible with a stepped load characteristic to arrest the regeneration at intervals, so creating a multiple-state circuit.

The required load characteristic is possessed in a simple form by an assembly of saturated diodes with progressive bias voltages. A rising generator characteristic is provided by a once-stabilized power unit, modified so that the control voltage is developed across a resistor in series with the load. The gradient of the generator curve depends upon this resistance, which is set to 220 ohms in the example given in Figs. 1 and 2. It must vary by less than 5 per cent if all the stable intersections are to be maintained. The filament voltages control the height of the steps in the load characteristic and are set low at 2·3 (± 5 per cent) volts. This permits a 75 per cent saving in heater power and makes the saturation current less critically dependent upon filament voltage. The bias potentials at the diode anodes are adjusted so that roughly equal upward and downward trigger voltages are necessary. Any trigger voltage between 8 and 24 volts is then found to be effective for any transition to the adjacent state.

With the voltmeter or ammeter marked in digits, the circuit behaves as a counter, which adds or subtracts according to the sign of the trigger pulse. Other possible applications are as a capacitor locking circuit[1] or as a voltage quantizing circuit[2].

A simpler, more practical circuit with fewer voltage sources, is shown in Fig. 3. The series generator has been reduced to a form of cathode-coupled multivibrator, in which the transition is arrested by saturated diodes in the pentode cathode lead. Fig. 4 shows the response of this circuit to trigger pulses of alternate sign applied to the triode grid. The complete transition occupies 20 μsec, although the initial rate is comparable with the usual speed of non-gaseous trigger circuits. The later reduction in speed is attributable to the too gradual increase in incremental diode resistance as saturation approaches. When trigger pulses are applied as shown, this limits the counting frequency to 50 kc/s, but higher speeds are possible in conjunction with a step-charging capacitor, when the diode system forms the capacitor locking circuit.

At low speeds and when subtraction is not wanted, it is useful to connect a small neon lamp in series with each diode to indicate the state of the circuit. The sudden striking of a neon effectively sharpens a vertical in the load characteristic. Intersections are easily arranged but the gap between neon striking and extinction voltages makes it difficult to accommodate both falling and rising parts of the load curve, which develops hysteresis.

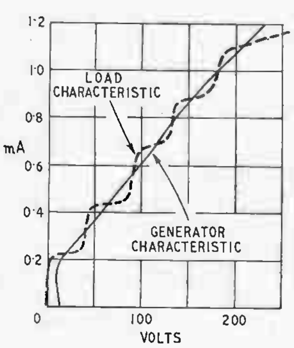


Fig. 1. *Stable states of a five-diode circuit; voltmeter and ammeter readings from the circuit of Fig. 2*

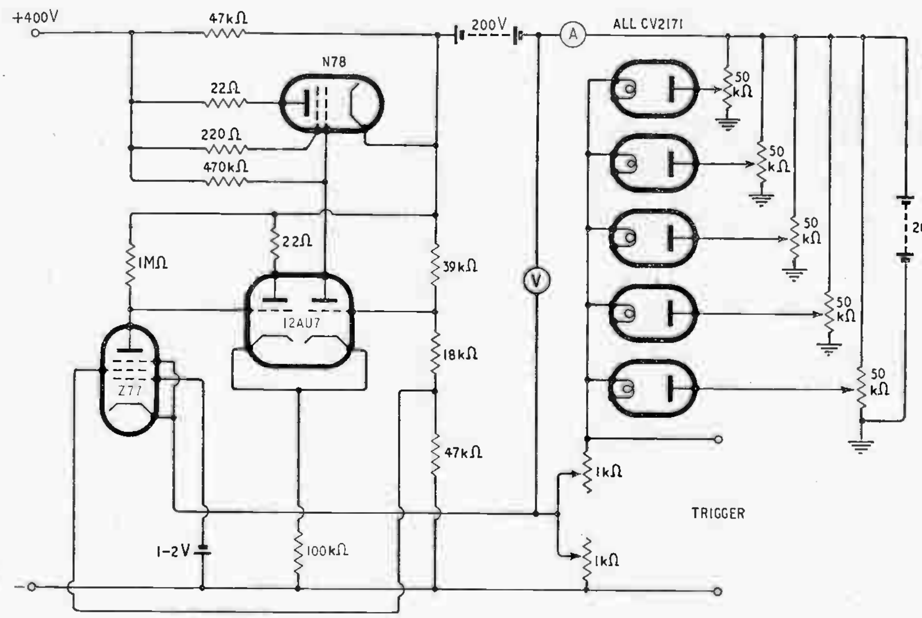


Fig. 2 (*right*). *Experimental circuit using a modified power unit*

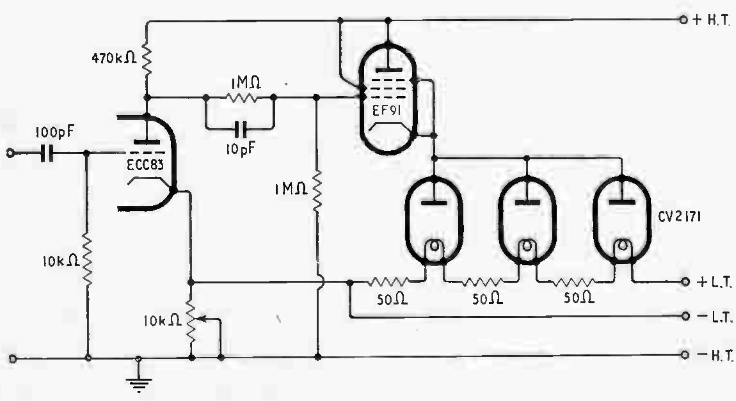


Fig. 3 (*left*). *Simplified circuit with one floating supply; V_{HT} = 300 V, V_{LT} = 90 V*

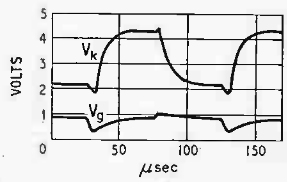


Fig. 4. *Reponse to trigger pulses of alternate sign*

A scale-of-ten circuit built around one multi-anode or multi-cathode diode, with a total anode current of about two milliamperes and low filament power, might possibly compare favourably with other decade systems in simplicity and power consumption. Disadvantages would be close tolerances on some components and voltages, together with the probable necessity for at least one floating supply, but the circuit could be free from gas-discharge tubes and could operate directly in both addition and subtraction.

Dept. of Electrical Engineering, D. Midgley
University of St. Andrews, Dundee.
1st May 1958.

REFERENCES

[1] F. H. Wells, "A Fast Amplitude Discriminator and Scale of Ten Counting Unit for Nuclear Work", *J. Scientific Instrum.*, April 1952, Vol. 29, p. 111.
[2] G. C. Tootill, "Some Storage Circuits Based on Valves", *Proc. Instn elect. Engrs*, Part B, Supplement No. 1, March 1956, Vol. 103, p. 313.

The Transactor

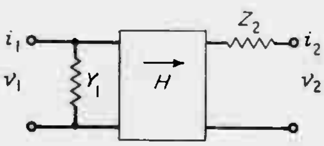
Sir,—Consideration of the newly-proposed circuit element, the transactor[1], appears to throw some light on the use of transistors in place of thermionic valves.

The four possible transactors are shown in Fig. 1 together with their A matrix representation. Such ideal transactors have infinite

TYPE	$\vec{Y}$	$\vec{Z}$	$\vec{H}$	$\vec{G}$
INPUT OUTPUT	VOLTAGE CURRENT	CURRENT VOLTAGE	VOLTAGE VOLTAGE	CURRENT CURRENT
MATRIX	$\begin{matrix} 0 & 1/Y \\ 0 & 0 \end{matrix}$	$\begin{matrix} 0 & 0 \\ 1/Z & 0 \end{matrix}$	$\begin{matrix} 1/A & 0 \\ 0 & 0 \end{matrix}$	$\begin{matrix} 0 & 0 \\ 0 & 1/D \end{matrix}$

$$\begin{vmatrix} v_1 \\ i_1 \end{vmatrix} = \begin{Vmatrix} A \end{Vmatrix} \begin{vmatrix} v_2 \\ i_2 \end{vmatrix}$$

Fig. 1



$$\begin{vmatrix} v_1 \\ i_1 \end{vmatrix} = {}^1/_A \begin{Vmatrix} 1 & Z_2 \\ Y_1 & Y_1 Z_2 \end{Vmatrix} \begin{vmatrix} v_2 \\ i_2 \end{vmatrix}$$

Fig. 2.

or zero input and output immittances and may be rendered non-ideal by adding immittances as shown in Fig. 2 for the H transactor. 'Feedback' admittance has been deliberately omitted. Similar non-ideal $\vec{Y}$, $\vec{H}$ and $\vec{G}$ transactors may be similarly formed.

However, when the well-known Thévenin-Norton 'dual' forms of the generators are written in each of these, we see the essential identity of the four forms which are tabulated in Fig. 3 with a particular symbolism for the parameters present in each. It will be noted that if the parameters are related by the equations

$$r_g\, g_g = r_a\, g_a = 1$$
$$\mu = g_m\, r_a, \quad \beta = r_m\, g_a$$
$$r_m = \mu r_g, \quad g_m = \beta\, g_g$$

the four non-ideal transactors become equivalent and mutually replaceable in circuit.

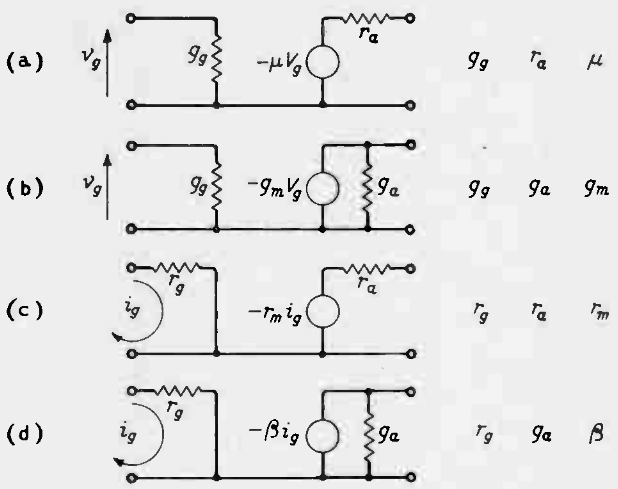


Fig. 3.

3(a) will be recognized as a commonly used equivalent circuit for a triode valve with grid leak. 3(b) for a pentode valve with grid leak, and 3(d) for a common-emitter alloy-junction transistor, valid at low frequencies in each case. The writer does not recognize a suitable equivalent for 3(c) at present.

The above equivalence appears to be the basis for the use of transistors in circuits topologically similar to circuits using thermionic valves, such as the multivibrator shown in Fig. 4. In this circuit the transistor must be taken not as the *dual* of the valve, but as the transactor equivalent.

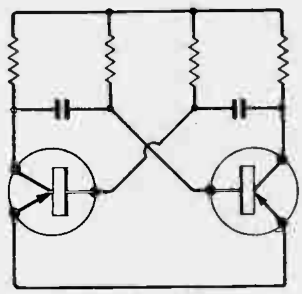
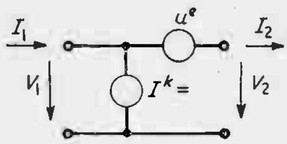


Fig. 4.

Another recent innovation, the 'K-amplifier'[2] can be shown to be equivalent to a transactor with permuted terminals.



$$V_1 = \sqrt{\Delta}.V_2 + 0.\,I_2$$
$$I_1 = 0.\,V_2 + \sqrt{\Delta}.I_2$$

Fig. 5.

Other idealized representations of active four-pole networks, for example that of Kurth[3] (Fig. 5), may lead to similar interrelations.

Decca Radar Ltd., M. R. Nicholls
Chessington, Surrey.
10th May 1958.

REFERENCES

[1] A. W. Keen, "The Transactor", *Electronic and Radio Engineer*, December 1957, Vol. 34, No. 12, p. 459.
[2] J. Zawels, "High Frequency Parameters of Transistors and Valves", *Electronic Engineering*, January 1958, Vol. 30, No. 359, p. 15.
[3] C. Kurth, "Amplifying 4-Poles and their Synthesis by Transistors", *Frequenz*, Vol. 11, No. 4, pp. 107–114.

Oscilloscope Frequency Ranges

Sir,—The excellent and valuable review of oscilloscope practice in the June issue, and the advertisements in the same issue, together provide many samples of an inelegant, anomalous, and indeed quite absurd form of specifying the frequency ranges of these instruments; namely, for example "D.C. to 10 Mc/s". It is surely unnecessary to explain to readers of *E. & R.E.* that current and frequency are not the same thing, but I would be grateful if someone would explain to me why so many otherwise intelligent beings go out of their way to imply that they are, instead of using the completely obvious and (in all other connections) normal form "0 to 10 Mc/s".

One irrationality of this kind leads to another, such as the expression "at d.c.". So before anyone catches the habit and specifies the range of a voltmeter as "X.Y.-250 V" (where "X.Y." in this particular context only means "0") could we not agree *always* to write "0" when we mean "0"?

Bromley, Kent. M. G. Scroggie.
4th June 1958.

Analysis of Current Pulses

SIR,—In his interesting paper[1] Mr. F. G. Heymann deduces that in class C amplification, where the i/v relationship is

$$i = kv^{\alpha} \text{ for } v>0 \quad \ldots \quad (1)$$
$$(i \text{ is zero for } v<0)$$

the ratio steady-current/peak-current varies appreciably with the exponent α while corresponding variations in the ratio steady-current/fundamental-current are comparatively small.

It may be worth pointing out that these conclusions can also be drawn from standard class C amplifier theory. Denoting the steady component of current, peak value of current, and peak value of the current at fundamental frequency by I_0, I_p, and I_1 respectively, Fourier analysis of (1) for a sinusoidal v leads to[2]

$$I_0/I_p = (\sin\theta - \theta\cos\theta)/\pi(1 - \cos\theta) \quad \text{for } \alpha = 1 \quad \ldots \quad (2a)$$

$$I_0/I_p = \left[\theta\left(1+\frac{1}{2}\cos 2\theta\right) - \frac{3}{4}\sin 2\theta\right]\Big/ \pi(1-\cos\theta)^2 \text{ for } \alpha = 2 \quad \ldots \quad (2b)$$

$$I_0/I_1 = \left(\sin\theta - \theta\cos\theta\right)\Big/\left(\theta - \frac{1}{2}\sin 2\theta\right) \text{ for } \alpha = 1 \quad \ldots \quad (3a)$$

$$I_0/I_1 = \left[\theta\left(1 + \frac{1}{2}\cos 2\theta\right) - \frac{3}{4}\sin 2\theta\right]\Big/ 2\left(\sin\theta - \frac{1}{3}\sin^3\theta - \theta\cos\theta\right) \text{ for } \alpha = 2 \quad \ldots \quad (3b)$$

2θ being the angle of current flow.

In class C operation, θ generally lies between 45° and 90°; Terman[3] quotes typical θ values of 70° and 58° for anode and grid currents respectively. Values of α normally range from 1 to 2, the former figure giving a close approximation for the anode current, while the latter is appropriate for the grid current[4].

Confining consideration to those limits, I_0/I_p, for $\theta = 45°$, equals 0·17 for $\alpha = 1$ and 0·13 for $\alpha = 2$; the corresponding figures for $\theta = 90°$ are 0·32 and 0·25. On the other hand, I_0/I_1 at $\theta = 45°$ varies only from 0·54 ($\alpha = 1$) to 0·53 ($\alpha = 2$), while the variation at $\theta = 90°$ ranges from 0·63 to 0·59.

The above figures show that I_0/I_1 depends very much less on α than does I_0/I_p.

British Telecommunications Research Ltd.,
Taplow, Bucks.
L. J. HERBST
15th May 1958.

REFERENCES

1. F. G. Heymann, "Analysis of Current Pulses", *Electronic and Radio Engineer*, May 1958, Vol. 35, p. 165.
2. J. P. Heyboer and P. Zijlstra, "Transmitting Valves", p. 36 (Philips Technical Library, 1951).
3. F. E. Terman, "Radio Engineers' Handbook", p. 449 (McGraw-Hill, 1945).
4. W. L. Everitt and K. Spangenberg, "Grid-Current Flow as a Factor in the Design of Vacuum Tube Power Amplifiers", *Proc. Inst. Radio Engrs*, 1938, Vol. 26, p. 621.

NEW PASSIVE SEMICONDUCTOR COMPONENT

A two-terminal, passive semiconductor component with novel characteristics has been developed at Bell Telephone Laboratories.

This component, which has been called the field-effect varistor, has a constant-current feature which makes it suitable for use as a current regulator. It can also be used as a current limiter or pulse shaper. Its a.c. impedance is very high, making it useful as a coupling or an a.c. switch.

The device contains a single planar junction which is made by diffusion. Current passes parallel to this junction through a constricted channel. As the voltage across the device is increased, current increases and a depletion layer builds up which eventually reaches through the entire thickness of the channel. At this point, called the 'pinch-off' point, a further increase in voltage does not produce any increase in current. Eventually an avalanche breakdown occurs as the voltage is increased still further. Between the pinch-off and breakdown points the current is essentially constant.

Using silicon, units have been made with a regulated current of 1 mA, pinch-off voltage of 10 V and breakdown of 150 V. Current can be held constant to within 1% over a voltage range of 20–120 V. Germanium units with a rating of 10 mA, pinch-off of 10 V and breakdown of 25 V have been made. It appears feasible at present to produce varistors which regulate current at any level between 10 μA and 10 mA.

New Books

Techniques of Magnetic Recording

By JOEL TALL. Pp. 472. The Macmillan Company, 10 South Audley Street, London, W.1. Price 55s. 6d.

A book for users of magnetic recorders. In addition to historical and general information, there are chapters on recording sound in nature, editing, radio-broadcasting practice, information recording, etc.

B.B.C. Handbook 1958

Pp. 288. B.B.C., Broadcasting House, London, W.1. Price 5s.

L'Automatique des Informations

By F. H. RAYMOND. Pp. 188. Masson et Cie, 120 Boulevard Saint-Germain, Paris VIe, France. Price 1600 fr.

Deals with computing data-handling techniques. (In French.)

Basic Feedback Control System Design

By C. J. SAVANT, Jr. Pp. 418. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 74s.

Contains chapters on the following subjects: Introduction to Feedback Control Systems; Obtaining the System Differential Equation; Steady-State Errors; The Root-Locus Method; Stability, the Frequency-Analysis Method; Servomechanism Equalization; Design of D.C. and A.C. Equalizers; Servomechanism Transducers; Servo Components and Applications; and Non-linearities in Servo Design. Appendixes deal with Laplace Transform Method; Classical Solution of Differential Equations; Use of Determinants; Hurwitz's Stability Criterion; Derivation of the Nyquist Criterion; Derivation of M and N Circles; Relations between Stability Quantities; Design of Bridged- and Parallel-T Networks, and Fourier Series.

The Economic Development of Radio

By S. G. STURMEY. Pp. 284. Gerald Duckworth & Co. Ltd., 3 Henrietta Street, London, W.C.2. Price 30s.

A study of the development of the British radio industry, made with a view to determining the conditions in which progress has occurred.

Einführung in die Microwellen-Elektronik. Teil 2, Lauffeldröhren

By WERNER KLEEN and KLAUS PÖSCHL. Pp. 192. S. Hirzel Verlag, Birkenwaldstrasse 185, Stuttgart N., Germany. Price D.M. 28.

Deals with the theory of travelling-wave and backward-wave valves and allied devices.

An Approach to Audio Frequency Amplifier Design

Pp. 126. The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2. Price 10s. 6d.

Describes amplifiers designed round G.E.C. valves with output powers of 5–1,100 W.

Servomécanismes

By M. BONAMY. Pp. 284. Masson et Cie, 120 Boulevard Saint-Germain, Paris VIe, France. Price 4,200 fr.

Vacuum Tube Rectifiers

Edited by A. SCHURE. Pp. 78. Price $1.50.

Industrial Control Circuits

By SIDNEY PLATT. Pp. 194. Price $3.90.

D.C. Circuit Analysis

By A. SCHURE. Pp. 72. Price $1.35.

Nuclear Energy

By ALEXANDER EFRON. Pp. 63. Price $1.25.

Mechanics

By ALEXANDER EFRON. Pp. 112. Price $1.50.

The above five books are published by John F. Rider Inc. and are obtainable in this country from Mark Paterson & Co., 37 Panton Street, Haymarket, London, S.W.1.

B.B.C. Engineering Division Monographs

No. 17. The Design of a Linear Phase-Shift Low-Pass Filter. By L. WEAVER, B.Sc. Pp. 23. Price 5s.

The filter described was intended to remove signals and noise at frequencies above 3 Mc/s from the output of a television camera control unit, without degrading the picture. The pass-band achieved is flat within 0·1 dB to 3 Mc/s, and the group-delay characteristic varies by only ± 10 mμsec.

No. 18. The B.B.C. Colour Television Tests: An Appraisal of Results

By W. N. SPROSON, M.A., S. N. WATSON, A.M.I.E.E., and M. CAMPBELL, B.Sc.(Econ.). Pp. 39. Price 5s.

Discussion of results of field trials using a modified N.T.S.C. system. It is concluded that the system is capable of providing a satisfactory compatible colour television service in the band 41–68 Mc/s.

The above are obtainable post free from: B.B.C. Publications, 35 Marylebone High Street, London, W.1.

BRITISH STANDARDS

Methods of Testing Vulcanized Rubber

B.S. 903: 1957. Part C.4 deals with the determination of electric strength of insulating soft vulcanized rubber and ebonite. Part D.4 deals with the determination of the cross-breaking strength of ebonite. Prices: Part C.4, 3s.; Part D.4, 2s. 6d.

Universal Decimal Classification (U.D.C.) Trilingual

Abridged Edition. B.S. 1000B: 1958. The international standard U.D.C. abridgement. In German, English and French. Price £6.6s.

Determination of Electric Strength of Solid Insulating Materials at Lower Frequencies for Industrial Purposes

B.S. 2918: 1957. Price 6s.

Glossary of Terms Used in High Vacuum Technology

B.S. 2951: 1958. Covers vacuum systems, pumps and pump components, manometers and gauges, leak detectors, vacuum applications (including radio valve-making) and general terms. Price 7s. 6d.

The Use of Electronic Valves: Magnetrons and Special Quality Valves

B.S. Code of Practice CP. 1005, Part 4: 1958. The section on magnetrons gives guidance on magnet design and on dangerous radiations which may be encountered in use. The section on special-quality valves emphasizes that the term 'special quality' may only apply to some features of a valve. Price 5s.

Safety Colours in the Factory

B.S. 2929: 1957. As these recommendations are applied in factories, mines, shunting yards and other industrial installations, the confusion which now exists through the use of a variety of different colour codes should gradually be removed. Price 4s. 6d.

The above are obtainable from the British Standards Institution, 2 Park Street, London, W.1.

MANUFACTURERS' LITERATURE

S.T.C. Valves Application Report G.10/241E. Pp. 21. Details of the electrical characteristics of the cold-cathode single-pulse decade counter valve type G.10/241E, with recommended circuits.
Standard Telephones & Cables Ltd., Connaught House, 63 Aldwych, London, W.C.2.

Consolidated Net Price List and Buying Guide. Pp. 25. Includes 'Variac' voltage-adjusting transformers, mains transformers, decade-condenser boxes, a.c. automatic voltage stabilizers and motor-operated automatic voltage regulators.
Claude Lyons Ltd., 76 Oldhall Street, Liverpool 3.

Marconi Instruments 1958. Pp. 300. Signal sources, frequency meters, voltmeters, power meters, distortion meters, field-strength meters, transmission measuring sets, station monitors, deviation meters, oscilloscopes, spectrum and response analysers, Q-meters, bridges, systems instrumentation and industrial equipment.
Marconi Instruments Ltd., St. Albans, Herts.

Wiring Cables. Pp. 25. Deals with v.i.r., p.v.c. and polythene insulated cables.
Mersey Cable Works Ltd., Linacre Lane, Bootle, Liverpool 20.

Transistor Information. Pp. 15. General information and tabulated data on Pye transistors.
Pye Industrial Electronics Ltd., Exning Road, Newmarket.

Classified Index to Ferranti Computer Literature. Pp. 11. Lists documents available. Addendum No. 1 (pp. 3) contains additions and amendments.
Ferranti Ltd., London Computer Centre, 21 Portland Place, London, W.1.

Royce Electric Furnaces. Pp. 12. Ovens and furnaces for temperatures of 150°C to 2,500°C.
Royce Electric Furnaces Ltd., Sir Richards Bridge, Walton-on-Thames, Surrey.

METAL SPRAYING CONFERENCE

The second International Metal Spraying Conference is being held from 29th September to 3rd October 1958 at the College of Technology, Gosta Green, Birmingham. The proceedings will be translated simultaneously into French, German and English and distributed by an internal telephone system.

Programmes and application forms for registration are obtainable from The Association of Metal Sprayers, Barclay's Bank Chambers, Dudley, Worcs.

Among the foreign contributions will be papers on Anti-Friction Bearings, Sprayed-Steel Wires and Sprayed Metallic-Zinc Coatings.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for May* 1958

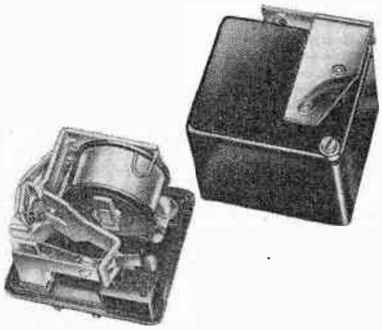
Date 1958 May	MSF 60 kc/s 2030 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	− 1	+ 3
2	− 1	+ 2
3	− 1	N.M.
4	− 1	N.M.
5	− 1	+ 2
6	N.M.	N.M.
7	− 1	N.M.
8	− 1	N.M.
9	− 1	N.M.
10	− 1	N.M.
11	− 1	N.M.
12	0	N.M.
13	− 1	+ 4
14	− 1	+ 4
15	N.M.	+ 4
16	− 1	− 1
17	− 1	N.M.
18	− 1	N.M.
19	0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

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

New Products

Motor-Starting Relay

This design (series 311) is intended as an alternative to a centrifugal switch for starting F.H.P. a.c. motors.



The relay is connected in series with the motor, being adjusted so that it pulls in on the starting current and drops out when the motor is up to speed, open-circuiting the start winding. Single-pole normally-open contacts are operated by a small current differential and coils can be wound to carry currents from 0·5 to 10 amps.

The relay is of the balanced type and may be mounted in any position.

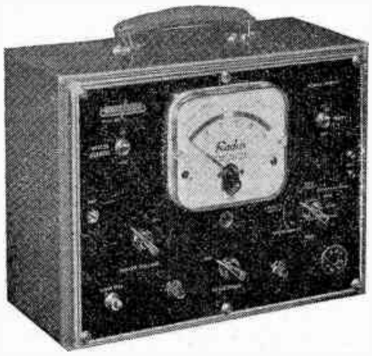
Another type (series 310, not shown) is produced as an alternative to a centrifugal switch for starting single-phase capacitor-start motors. The contact arrangement is single-pole normally-closed. Operation is by a wide coil-voltage differential.

In this case, the relay winding is connected across the start winding of the motor and as the motor speed increases, the p.d. across the start winding and, therefore, across the relay coil, also increases, causing the relay to operate, which disconnects the start winding. The induced voltage in the start winding continues to energize the relay down to the stalling speed of the motor.

Magnetic Devices Ltd.,
Exning Road, Newmarket, Suffolk.

Cathode-Ray Tube Tester/Reactivator

An improved version of the Radar cathode-ray tube tester/reactivator has been introduced.



Primarily intended for testing television picture tubes, it also incorporates facilities for reactivation by a new pulsing method which gives visual indication when the process is complete.

Tests that can be carried out include heater-current measurements, interelectrode leakages and emission. The latter is achieved by measuring the beam current.

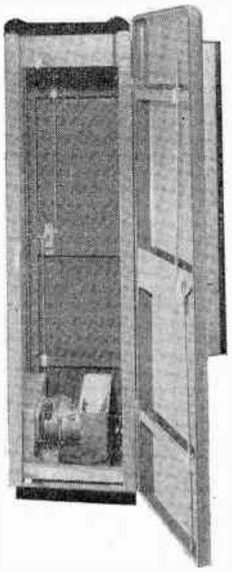
A device for clearing interelectrode leakages is included.

In addition, model 202 contains the facilities required to measure the potentials applied to the tube while operating in a receiver.

Waveforms Ltd.,
Radar Works, Truro Road, London, N.22.

Forced Ventilation Racks

This product is an enclosed rack specially designed to give unusually effective cooling. A powerful blower unit fitted to the base gives (in a 6-ft. rack 19 in. deep with 19 in. panels) one air change every four seconds; semi-sealing of the doors ensures the most efficient cooling commensurate with economy.



Standard forced-ventilation racks are available in heights of 4 ft., 5 ft., 6 ft. and 7 ft. with a choice of panel widths of 19 in. and 22½ in.; the standard depths are 19 in. and 24 in.

The blower unit, which can be bought separately, is fitted with expendable filter pads of glass wool which can easily be changed. If localized cooling is required, trunking may be connected to the flange around the deflecting chamber.

Alfred Imhof Ltd.,
112–116 *New Oxford Street, London, W.C.*1.

Low-Resistance Testing Set

The 'Ducter' ohmmeter is a direct-reading instrument for measuring low resistances from a few ohms down to one microhm.

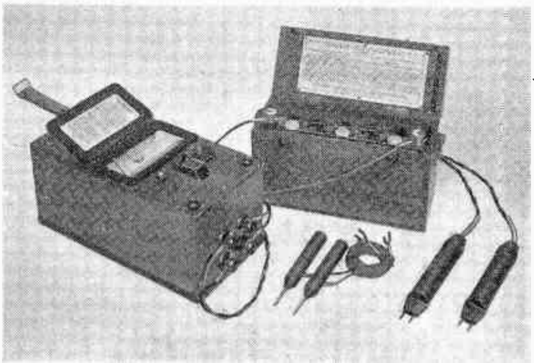
It is said to be especially useful for measuring the resistances of switch contacts, rail bonds, armature windings, transformer windings, copper bars, fuses, soldered joints and earth continuity conductors when a highly accurate reading is necessary.

Various models are available, with ranges of 0·0001–1Ω, 0·001–1Ω and 0·0005–5Ω.

The movement is of the permanent-magnet moving-coils type. The moving element consists of two coils rigidly secured at an angle to one another and pivoted in spring-mounted jewel bearings so that the coils swing in the annular gap between a fixed centre iron and the pole pieces of a permanent magnet. Connections to the moving coils are made through thin ligaments.

The centre coil is connected across a shunt in the main circuit and the current in this coil is therefore proportional to the current in the resistance under test. The deflecting coil is connected across the resistance under test and carries a current proportional to the potential drop.

The polarity of the two coils is so arranged that the torques developed oppose one another. The position of the movement is therefore dependent on the ratio between the potential drop and the current; i.e., on the value of the resistance being



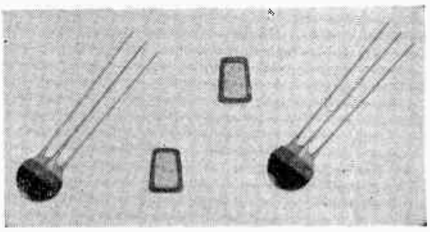
measured. The scale is calibrated directly in ohms or microhms.

As the instrument is a true ohmmeter, its accuracy is unaffected by variation in battery voltage, and no adjustment is required.

Evershed & Vignoles Ltd.,
*Acton Lane Works, Chiswick, London, W.*4.

Ceramic Capacitors

Two new ranges of ceramic capacitors have been added to the standard Hi-K and Hi-Q types. The new types are, first, 'Twin Disc' Hi-K discs, providing a dual capacitance on one disc. (That shown is 1,000 + 1,000 pF, each section 500 V d.c. working.)

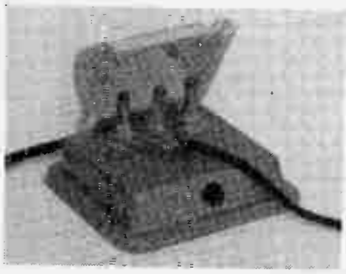


Second, ceramic wedge capacitors for plug-in insertion in printed wiring. (That shown is 1,000 pF at 500 V d.c. working.)

A. H. Hunt (Capacitors) Ltd.,
Bendon Valley, Garratt Lane, Wandsworth, London, S.W.18.

Snap-Action Appliance Connector

The connector illustrated enables equipment to be connected to the mains without using a mains plug. It is provided with spring-loaded quick-release terminals rated



at 15 A a.c. To connect a wire, the domed insulator is tilted, the wire inserted, and the insulator released to spring back and grip the lead.

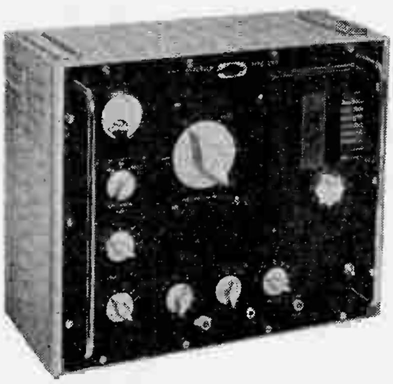
The terminals can be supplied separately.

Mycalex and T.I.M. Ltd.,
Ashcroft Road, Cirencester, Glos.

V.H.F. Wave Analyser

The v.h.f. wave analyser type 248 is described as a selective measuring set for use in the frequency range 5 Mc/s to 300 Mc/s.

The circuit consists of a common-grid r.f. amplifier feeding a diode mixer operating on the fundamental of the injected local oscillator signal. Waveband switching is accomplished by a special v.h.f. coil turret, individual gain adjustment being provided on each band. The resulting i.f. signal is applied to a 2-stage negative feedback amplifier, using a low-noise input valve, followed by switched low-pass filters giving a choice of three cut-off frequencies. The



i.f. signal then passes through precision coarse and fine attenuators before entering the final 3-stage negative-feedback amplifier which operates the meter circuit. Germanium crystals are used for rectification.

A further amplifier is provided for modulation monitoring. It acts as a linear amplifier for amplitude-modulation monitoring and as a limiter-amplifier for frequency-modulation monitoring. After demodulation, either a.m. or f.m., the audio signal is amplified before being passed to the phones jack. A high degree of stability is achieved by feeding the r.f. amplifier and oscillator from a stabilized h.t. supply.

Signals down to 5 μV can be measured. This input gives unity signal-to-noise ratio.

Airmec Ltd.,
High Wycombe, Bucks.

Sub-Miniature 'Magic Eye'

A new sub-miniature cathode-ray indicator tube type DM160 can provide visual indication of the condition of two-state circuits.

The DM160 is a triode valve in which the anode serves as a fluorescent screen, giving a rectangular blue-green glow some 20 mm square in area.

Features of the tube are its high sensitivity (-3 V for minimum light output) and low voltage and current requirements. The filament is rated at 1 V, 30 mA, and typical anode voltage and current are 50 V, 0·5 mA (at maximum light output). At minimum light output, the anode current is less than 5 μA.

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

Stabilizer for Alternating Voltages

The AC7 voltage stabilizer is claimed to maintain an alternating output voltage constant to within 0·25% against mains variations of $-$ 20% to + 10% at load currents of 0–30 A. The response time for

small errors is given as 0·25 sec and for large errors the rate of correction is 12 V per sec. The output waveform is not distorted.

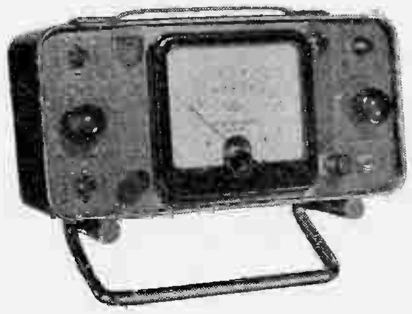
The instrument contains a bolometer bridge, the output of which is fed to a servo amplifier. The servo-motor drives a voltage-correcting device. Velocity feedback is incorporated to reduce the effects of friction and to provide stability. It is claimed that the instrument is unaffected by changes in temperature up to 20° C and accelerations up to 40 g.

Servomex Controls Ltd.,
Crowborough Hill, Jarvis Brook, Sussex.

Transistor Voltmeter

The Amos Model 140 voltmeter incorporates two transistors in a balanced circuit which is claimed to be stable and linear. The range of voltages for full-scale deflections is 0·5–20,000 V d.c. There are six basic ranges, but five of these can be extended by means of a 'divide-by-two' button. Current taken by the instrument is 3 or 6 μA for full-scale deflection on the ranges up to 1,000 V, and more on the higher ranges. A special probe incorporating a 200-MΩ $\pm$ 1% resistor is provided for e.h.t. measurements. For calibration there is an internal reference voltage, the accuracy of which is given as 1%.

The coarse zero-setting control may be used to deflect the meter needle half-way and so make the instrument centre-zero; a suitable scale is provided.



The transistors are positioned in a large heat sink which is lagged and housed in a heat-reflecting case. The current drain on the 4·5-V battery is 300 μA, and that on the 7·5-V reference battery is negligible. The instrument is obtainable from the address below.

Soundrite Ltd.,
82-83 New Bond Street, London, W.1.

Medium-Mu Triode

The 6814 is described as a medium-mu triode of the flexible-lead sub-miniature type for use in electronic computers and other 'on-off' control applications involving long periods of operation under cut-off conditions. The valve is said to be designed to give dependable performance under conditions of shock and vibration and to be especially suited for use in mobile and aircraft equipment. It is rated for service at altitudes up to 80,000 ft without the use of pressurized chambers.

It is claimed that in computer service and other 'on-off' control applications, the 6814 maintains its emission capabilities even after long standby periods and can supply a high value of plate current during its 'on' cycles. Furthermore, consistency of cut-off bias is maintained because of its stable cut-off characteristic and its freedom from grid emission.

The 6814 has a cathode made of special alloy material to minimize cathode interface resistance, a protective shield to prevent deposition of getter material on micas and electrodes and a pure-nickel anode to minimize gas evolution.

General data: heater, 6·3 V $\pm$ 5%, 0·15 A; interelectrode capacitances (with external screen): grid to anode 1·3 pF, grid to cathode and heater 2·4 pF, anode to cathode and heater 2·4 pF; anode voltage 100 V; cathode resistor 150 Ω; anode current 10 mA; mutual conductance 6 mA/V; amplification factor 29; anode resistance 4,800 Ω.

R.C.A. Great Britain Ltd.,
Lincoln Way, Windmill Road, Sunbury-on-Thames, Middx.

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

534.2 **1942**

The Scattering of Sound Waves in Inhomogeneous Waveguides.—A. D. Lapin. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Jan. 1958, Vol. 118, No. 1, pp. 55–58.) Mathematical analysis for a waveguide filled with an inhomogeneous medium which has an average refractive index of about unity. Waveguides with rough interior walls are also examined. See also 3002 of 1957 (Isakovich).

534.232-14-8 : 534.133 **1943**

Amplitude of a Quartz Plate Vibrating in Liquids.—S. Parthasarathy & Harkrishan Singh. (*Nature, Lond.*, 25th Jan. 1958, Vol. 181, No. 4604, p. 260.) The amplitude of vibration of a quartz crystal plate in various liquids calculated according to Vigoureux's formula is compared with the amplitude of the resulting sound wave determined experimentally by a calorimetric method [see 1602 of 1956 (Parthasarathy & Narasimhan)].

534.41 : 621.372.543.2 **1944**

Active, Adjustable Audio Band-Pass Filter.—J. R. Macdonald. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1348–1356.) A filter having seventh-order Butterworth attenuation characteristics (42 dB/octave slopes) is described. Six miniature double triodes and no inductors are used.

534.612 : 061.3 **1945**

Sound Level Meter Symposium.—(*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1330–1341.) A summary is given of 10 papers read at a symposium held by the Acoustical Society of America in New York, 23rd–25th May 1957 in connection with Committee Z24-W-18 of the American Standards Association.

534.75 **1946**

Comparison of the Auditory Threshold as Measured by Individual Pure Tone and by Békésy Audiometry.—W. Burns & R. Hinchcliffe. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1274–1277.)

534.75 **1947**

Threshold of Hearing and Equal-Loudness Relation for Pure Tones, and the Loudness Function.—D. W. Robinson & R. S. Dadson. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1284–1288.) "In connection with standards for audiometry, measurements have been made of the threshold of hearing for pure tones by earphone listening, and these have since been extended to the case of listening in free field. Latterly a redetermination of the equal-loudness relations for pure tones has been completed aimed at resolving discrepancies between former determinations and providing an improved basis for the establishment of a standard set of contours. These results apply to a large team of otologically normal observers, and cover the range from 25–15 000 c/s and up to 130 dB in sound pressure level."

534.75 **1948**

Further Observations on Pitch associated with a Time Difference between Two Pulse Trains.—W. R. Thurlow. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1310–1311.) Changing the time separation between two trains of filtered pulses, each at the same basic repetition rate, makes it evident that there is a pitch which is correlated with the magnitude of the time separation. See also 1855 of 1955 (Thurlow & Small).

534.76 **1949**

An Unexpected Effect in Sound Localization.—D. H. Holding & J. P. Dennis. (*Nature, Lond.*, 28th Dec. 1957, Vol. 180, No. 4600, pp. 1471–1472.) Tests at 1 kc/s show that sound localization by human subjects is improved by the wearing of a light cloth cap.

534.78 **1950**

Further Test of the Constant-Ratio Rule in Speech Communication.—F. R. Clarke & C. D. Anderson. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1318–1320.) "The use of the constant-ratio rule to predict the confusion matrices for each of two five-item subsets given the confusion matrix for a ten-item master set is tested with naive subjects." See also 10 of January (Clarke).

534.78 **1951**

Effects of Ambient Noise on Speaker Intelligibility for Words and Phrases.—J. J. Dreher & J. J. O'Neill. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1320–1323.) Report of experiments based on a 'voice-reflex' effect whereby a speaker with normal hearing unconsciously raises his voice and changes the characteristics of his speech to compensate for the level of noise in which he is speaking.

534.78 **1952**
Effect of Noise and Filtering on Speech Intelligibility at High Levels.—I. Pollack & J. M. Pickett. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1328–1329.) The effects of a high level of background noise on intelligibility are compared with the effects of filtering.

534.78 : 621.39 **1953**
Signal Theory in Speech Transmission.—E. E. David, Jr. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 232–244. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, pp. 716–717.)

534.78 : 621.395.623.54 **1954**
Speech Communication at High Noise Levels: the Roles of a Noise-Operated Automatic Gain Control System and Hearing Protection.—I. Pollack. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1324–1327.) Two aids for hearing conservation, a noise-operated automatic gain control system and an insert ear protection, were evaluated in terms of their effect upon speech intelligibility. At high noise levels, these aids not only do not interfere with the speech intelligibility, they may substantially improve speech intelligibility while affording hearing protection.

534.78 : 621.395.625.3 **1955**
Intelligibility Measurements on Transmission Systems using German Logatoms Recorded on Magnetic Tape.—W. Wisch. (*NachrTech.*, Aug. 1957, Vol. 7, No. 8, pp. 342–345.) A method of forming German logatoms is described. The use of tape recordings for subjective tests appears to be advantageous.

534.78 : 681.188 : 8.03 **1956**
Vowel Recognition in Clipped Speech.—R. Ahmed. (*Nature, Lond.*, 18th Jan. 1958, Vol. 181, No. 4603, p. 210.) In the course of work on speech-input devices for a translation machine a pulse count analysis has been made of vowel sounds in clipped speech. The consistency of the results is illustrated.

534.84 **1957**
Acoustical Field at Normal Modes in Rooms.—T. S. Korn & R. Van de Plas. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1267–1270.) "The physical form of the acoustical field at normal complex modes in rooms can be represented as a set of elementary curvilinear ducts, each carrying a single stationary wave, satisfying individually the boundary conditions at the walls. In spite of the differences in their lengths, all the ducts resonate at the same nominal frequency."

534.844/.845 **1958**
Design and Performance of a New Reverberation Room at Armour Research Foundation, Chicago, Illinois.—D. R. McAuliffe. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1270–1273.) The room has a volume of approximately 3 000 ft^3 and a reverberation time of about 5 s. The four walls and ceiling of the room have been splayed. Measurements indicate that the room has a relatively uniform reverberation time with respect to frequency.

534.88 **1959**
Analysis of a Multiple-Receiver Correlation System.—M. J. Jacobson. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1342–1347.) A theoretical study is made of an acoustic receiving system consisting of two arrays of omnidirectional receivers and one correlator.

621.395.614 : 546.431.824-31 **1960**
Laboratory Standard Ceramic Microphone.—W. Newitt. (*J. acoust. Soc. Amer.*, Dec. 1957, Vol. 29, No. 12, pp. 1356–1365.) Microphones using $BaTiO_3$ disk ceramics in conjunction with a split-tube diaphragm are adaptable for use at high sound pressures, and for sound measurements in liquids as well as in air. Sensitivities from -86 to -88 dB referred to 1 V per dyne/cm^2 were obtained and the frequency response was uniform within ± 1 dB from 10 c/s to 30 kc/s.

621.395.623.74 **1961**
Principles of Loudspeaker Design and Operation.—J. Chernof. (*Trans. Inst. Radio Engrs*, Sept./Oct. 1957, Vol. AU-5, No. 5, pp. 117–127. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 800.)

621.395.625 : 621.317.76 **1962**
'Wow' and 'Flutter'.—Bennett & Currie. (See 2190.)

621.395.625.3 : 621.397.5 **1963**
Television Tape Recorder.—(See 2244.)

621.395.625.3 : 621.397.5 **1964**
Television Recording on Magnetic Tape by the Ampex Method.—Braunmühl & Schmidbauer. (See 2245.)

AERIALS AND TRANSMISSION LINES

621.315.212.029.63/.64 **1965**
Variations of Characteristic Impedance along Short Coaxial Cables.—J. Allison. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 169–176.) The measured deviations from a harmonic series of the resonance frequencies of a short length of open- or short-circuited cable are used to calculate the coefficients of a Fourier series describing the impedance variations along the length of the cable.

621.372.2 **1966**
Outline of a Theory of Nonuniform Transmission Lines.—B. G. Kazansky. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 126–138.) Families of complex differential equations are analysed and the corresponding two-parameter solution curves investigated. The relationship between singular points of the differential equations for the normalized impedance and the nonreflective impedances of the corresponding nonuniform line are established.

621.372.2 : 621.3.012 **1967**
Transformation of the Smith Chart through Lossless Junctions.—H. V. Shurmer. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 177–182.) The relation between the Smith chart representations of the impedances on either side of a lossless junction is analysed using complex variables. A chart is derived for transforming circles of constant voltage s.w.r. relating to any plane on one side of the junction, into corresponding circles relating to the corresponding plane on the other side.

621.372.21 : 621.396.65.029.6 **1968**
Cable and Radio Links for the Microwave Range.—O. Zinke. (*Nachrichtentech. Z.*, Sept. 1957, Vol. 10, No. 9, pp. 425–430.) A comparison of the suitability and cost of coaxial and waveguide transmission lines for the various frequency ranges used in radio links. The performance of single-wire lines in the range 30–3 000 Mc/s is discussed in detail. The alternatives examined are the plain wire (Sommerfeld), the dielectric-coated (Goubau) straight line, and helical plain and coated lines.

621.372.8 **1969**
A Contribution to the Theory of Probes in Waveguides.—L. Lewin. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 109–116.) The theory of a probe spanning a rectangular guide and fed by a coaxial line is analysed in detail in two cases: (*a*) the short-circuit case considering the probe both as a top-short-circuited aerial and also as a waveguide post terminated by the cable impedance; (*b*) a top-loaded probe, terminated by means of a tuning plunger.

621.372.821 : 621.3.095 **1970**
Parallel-Plate Transmission Lines and Equivalent Radiators.—A. B. Hillan. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 170–173.) It is shown that an infinite plane current sheet has no induction field, and that the radiated field is a faithful reproduction of the current density.

621.372.826+621.396.677 **1971**
Excitation of Surface Waves.—B. Friedman & W. E. Williams. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 252–258.) "It is shown how to locate a dipole source above a dielectric surface so as either to produce as pure a surface wave as possible or to maximize the amount of energy carried by the surface wave."

621.372.837.3 : 621.318.134 **1972**
Some Applications of Ferrites to Microwave Switches, Phasers and Isolators.—A. C. Brown, R. S. Cole & W. N. Honeyman. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 722–727.) A switch is described which is matched in both

states, and also one which can be used at a high repetition rate. The scaling of isolator designs is examined.

621.372.85 : 537.226 : 621.315.612.4 **1973**
Vacuum Breakdown in Dielectric-Loaded Waveguides.—G. B. Walker & E. L. Lewis. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, pp. 38–39.) Experiments of the type reported earlier [see 2560 of 1957 (Shersby-Harvie et al.)] were continued with the aim of studying breakdown in TiO_2 dielectric disks. Lead borate enamel is found to be very effective in preventing breakdown.

621.372.85 : 621.318.134 **1974**
Cut-Off Phenomena in Transversely Magnetized Ferrites.—R. F. Soohoo. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 788–789.) The cut-off frequencies are determined for a rectangular waveguide loaded with a lossless ferrite slab, and are checked by experiment.

621.396.67.095.1 **1975**
Transmission between Aerials of Any Type.—K. Baur. (*Frequenz*, Oct. 1957, Vol. 11, No. 10, pp. 308–312.) Summary of practical formulae for evaluating received power together with definitions of various aerial parameters, covering reception in linearly and elliptically polarized fields.

621.396.673 **1976**
The Effect of the Ground Constants, and of an Earth System, on the Performance of a Vertical Medium-Wave Aerial.—G. D. Monteath. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 292–306.) The compensation theory is used to examine the effect of ground conductivity and of an earth system upon the input impedance, the ground-wave field strength for a given aerial current, and the vertical radiation pattern. If the effective height of the aerial exceeds $0 \cdot 1\,\lambda$, an earth system of $0 \cdot 2\,\lambda$ radius is sufficient for ground of high conductivity and of $0 \cdot 3$–$0 \cdot 4\,\lambda$ radius if of low conductivity. The vertical radiation pattern is not materially affected by the earth system.

621.396.677.029.63 **1977**
The Design of Decimetre-Wave Wide-Band Directional Aerials with Extremely Small Reflection Coefficient.—H. P. Wolff. (*Optik, Stuttgart*, Oct. 1957, Vol. 14, No. 10, pp. 458–473.) Two designs of radiating elements for use at 2 kMc/s with a bandwidth amounting to 10% of midband frequency and a reflection coefficient less than $2\frac{1}{2}$% are considered. In one design a paraboloidal reflector 3 m in diameter is fed from a full-wave dipole with the reflector feedback cancelled; in the other a half-wave dipole is fitted but the feedback of the paraboloid is used to match the impedance of the radiator to that of the feeder.

621.396.677.3 **1978**
A Simplified Derivation of the Fourier Coefficients for Tchebycheff Patterns.—J. L. Brown, Jr. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 167–168.) An alternative derivation of Salzer's result (350 of 1957).

621.396.677.6 : 621.396.933.2 **1979**
Study of the Feasability of Airborne H.F. Direction-Finding Antenna Systems.—Carter. (See 2096.)

621.372.2 **1980**
Transmission-Line Theory. [Book Review]—R. W. P. King. Publishers: McGraw Hill, London, 1955, 509 pp., 90s. (*Nature, Lond.*, 28th Dec. 1957, Vol. 180, No. 4600, p. 1443.)

AUTOMATIC COMPUTERS

681.142 **1981**
Digital Codes in Data-Processing Systems.—M. P. Atkinson. (*Trans. Soc. Instrum. Technol.*, Dec. 1957, Vol. 9, No. 4, pp. 124–130.) The design of digital coding systems including error-detecting and error-correcting codes is discussed.

681.142 **1982**
A New Cycle-Counting Instruction for a Three-Address Electronic Digital Computer.—L. Lunelli. (*Ricerca sci.*, Nov. 1957, Vol. 27, No. 11, pp. 3381–3394.) See also 3757 of 1957 (Dadda).

681.142 **1983**
An Electrical Apparatus for Solving Polynomial Equations.—H. Adler. (*NachrTech.*, Aug. 1957, Vol. 7, No. 8, pp. 335–342.) The analogue equipment described can be used for solving equations of up to 8th degree with real coefficients.

681.142 : 514 **1984**
Computation of Arc Tan N for $-\infty < N < +\infty$ using an Electronic Computer.—E. G. Kogbetliantz. (*IBM J. Res. Developm.*, Jan. 1958, Vol. 2, No. 1, pp. 43–53.)

681.142 : 538.221 **1985**
The Application of Square-Hysteresis-Loop Materials in Digital Computer Circuits.—A. D. Holt. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 196–199.) "Theories and the practical design of shifting registers are explained. Core matrix storage systems are reviewed. A description is given of a store in which a shifting register is used to convert the parallel output into serial form; this register is also used during the writing process."

681.142 : 621.374.32 **1986**
A Decimal Product Accumulator.—R. R. Hoge. (*J. Brit. Instn Radio Engrs*, Feb. 1958, Vol. 18, No. 2, pp. 125–132. Discussion, p. 133.) Dekatrons enable the correlation between two series of numbers to be found if all terms are positive. Given pairs of numbers of two decimal digits each, the machine accumulates 100 products per minute.

681.142(410) **1987**
British-Built Computers.—(*Overseas Engr*, Dec. 1957, Vol. 31, No. 360, pp. 167–173.) A review of digital and analogue computers with details of specifications, special features and applications.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.018.7 : [621.314.6+621.375.2 **1988**
Analysis of Current Pulses.—F. G. Heymann. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 165–167.) An approximate method of analysis used earlier (2439 of 1955) is extended to include nonlinear characteristics, and its application to a class-C amplifier circuit is described.

621.3.049.75 : 621.375.1 **1989**
Printed Circuits and High-Quality Amplifiers.—(*Point to Point Telecommun.*, Feb. 1958, Vol. 2, No. 2, pp. 18–24.) Printed circuits are suited to equipment where standardization is essential. Inherent in this type of circuitry is improved performance; special production methods are needed to ensure reliability.

621.3.049.75 : 621.397.62 : 535.623 **1990**
Etched I.F. Amplifier Pares Colour TV Cost.—Ruth. (See 2248.)

621.311.62 : 621.3.072.2 **1991**
Simple Variable-Voltage Power Supply.—J. M. Macintosh. (*Short Wave Mag.*, Jan. 1958, Vol. 15, No. 11, pp. 568–571.) Design details of a circuit giving an output of 0–200 V at currents up to 60 mA.

621.318.42 : 538.221 : 538.566 **1992**
Electromagnetic Fields in a Ferromagnetic Medium, with Particular Reference to Harmonic Distortion due to Hysteresis.—V. G. Welsby. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 218–229.) The frequency dependence of the third-order distortion factor of an inductor at low flux densities is studied theoretically. An analogue method is described which enables the study to be extended experimentally to large flux densities and complex input waveforms.

621.318.5 **1993**
A Numerical-Graphical Method for Synthesizing Switching Circuits.—A. H. Scheinman. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 687–689.) Economical circuit arrangements can be rapidly developed by systematic application of three simple rules to a numerical representation of a Boolean function. Simple examples are given.

621.318.57 **1994**
Minimum-Energy Triggering Signals.—L. A. Beattie. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 751–757.) The optimum signal is defined as that which produces a given current through, or a voltage across, a resistive output element at a given time, while requiring a minimum of energy from the generator. The solution of this problem is discussed and illustrated by a simple case.

621.318.57 : 621.314.7 **1995**
Controlled Saturation in Transistors and its Application in Trigger Circuit Design.—N. F. Moody. (*Electronic Engng*,

March & April 1958, Vol. 30, Nos. 361 & 362, pp. 121–127 & 200–204.) Carrier storage of charge in transistors has a maximum which is defined as controlled saturation; this depends only on the base current. This idea is used to develop a trigger circuit of good resolving time whose output terminal is able to handle heavy currents and presents a low impedance.

621.318.57 : 621.314.7 **1996**

Fast Transistor Relay.—D. L. Anderson. (*Electronics*, 14th March 1958, Vol. 31, No. 11, p. 145.) Describes a push-pull switching circuit incorporating Zener diodes having a rise time of 50 μs and capable of handling currents up to 10 A.

621.372.011.1 **1997**

Generalized Operators for the Approximate Steady-State Analysis of Linear and Nonlinear Circuits.—A. J. O. Cruickshank. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 76–87.) The method is a periodic analogue of the time-series technique by which waveforms are represented by an n-component operator giving values at each of n ordinates. A shift operator translates any waveform by $1/n$th of the appropriate sample period. For linear circuits the method is advantageous where the periodic input is known and a similar description of the output is required without recourse to Fourier analysis. For nonlinear circuits the output is obtained by a series of approximations applied to the steady-state condition.

621.372.012 : 621.391 **1998**

Signal Theory.—W. H. Huggins. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 210–216. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 716.)

621.372.012 : 621.391 **1999**

System Theory as an Extension of Circuit Theory.—W. K. Linvill. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 217–223. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 716.)

621.372.09 **2000**

Derivation of the Phase Characteristic from the Attenuation Curve in Minimum-Phase Systems.—E. A. Graham. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 563–576.) The graphical method described is illustrated by numerical examples.

621.372.412 : 621.372.54 **2001**

Stable Crystal Filter is Parallel Resonant.—J. C. Seddon. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 155–157.) A high-Q, unbalanced crystal circuit is similar to a parallel-tuned circuit with a low L/C ratio over an appreciable frequency range. It is readily adaptable for use in f.m. oscillators, signal generators and variable-bandwidth filters.

621.372.413 **2002**

The Concept of Heterogeneous Surface Impedance and its Application to Cylindrical Cavity Resonators.—A. E. Karbowiak. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 1–12.) Formulae are developed relating the Q-factor and the resonant frequency of a cavity to its dimensions and the Fourier components of the surface impedance function. The cases of circumferential and axial heterogeneity are analysed in detail. In general, a unique value of surface impedance cannot apply to an unbounded periodic sheet unless the period is small compared with λ.

621.372.5 **2003**

Synthesis of Tchebycheff Impedance-Matching Networks, Filters and Interstages.—G. L. Matthaei. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. CT-3, No. 3, pp. 163–172. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 253.)

621.372.5 **2004**

Two Theorems concerning Group Delay with Practical Application to Delay Correction.—G. G. Gouriet. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 240–244.) "Two theorems are stated which enable properties of group delay to be expressed directly in terms of the transfer function of a linear transmission system. Examples are given to show how the results might be usefully applied to certain problems of delay correction."

621.372.5 : 621.374.3 **2005**

Linear Pulse-Forming Circuits.—W. C. Gore & T. Larsen. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. CT-3, No. 3, pp. 182–188. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 253.)

621.372.5 : 621.396.96 **2006**

Synthesis of Delay-Line Networks.—D. A. Linden & B. D. Steinberg. (*Trans. Inst. Radio Engrs*, March 1957, Vol. ANE-4, No. 1, pp. 34–39. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1034.) Description of synthesis procedure for delay systems used in processing continuous-trace radar information.

621.372.54 **2007**

The Network Synthesis on the Insertion-Loss Basis.—J. Zdunek. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 259–291.) Methods of synthesis of most important network functions are derived in terms of the steady-state network performance; realization procedures are explained and expressed in explicit formulae for direct application to the design of conventional ladder structures with an arbitrary number of branches. The symmetrical and the inverse-impedance low-pass networks are solved in detail.

621.372.54 : 621.318.4 **2008**

Exact Ladder Network Design using Low-Q Coils.—L. Weinberg. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 739–750.)

621.372.6 **2009**

Conditions for the Impedance and Admittance Matrices of n-Ports without Ideal Transformers.—I. Cederbaum. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 245–251.)

621.372.6.012.8 : 621.396.822 **2010**

Equivalent Circuits of Noisy Networks.—L. Young. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 205–207.) Equivalent circuits for a combination of amplifiers, attenuators and terminations are presented and a new method for the precision measurement of noise figure is described.

621.372.622 **2011**

Small-Signal Heterodyne Mixers with Excessive Injection Amplitudes.—J. F. Cline. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 739–745.) The theory of small-signal heterodyne mixers operating with very large injection potentials is described. The mixer properties investigated are the conversion transconductance, the relative amplitudes of the difference-frequency and injection-frequency components in the output, and the amplitude of the injection modulation-frequency component in the output in the case where the injection potential is amplitude-modulated to a small degree. A set of universal performance curves is obtained for mixers of a particular class chosen to illustrate the method.

621.373.421.13 **2012**

VXO—a Variable Crystal Oscillator.—H. Shall. (*QST*, Jan. 1958, Vol. 42, No. 1, pp. 11–15.) Description of a stable oscillator which covers the range 3 500–4 000 kc/s and uses six crystals.

621.373.431.1 **2013**

Flip-Flop Stability.—T. G. Clark. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 212–213.) Modifications to the circuit described earlier (716 of March), including the use of a different type of valve, do not affect its stability in respect of pulse duration.

621.373.444 : 621.314.7 **2014**

Transient Processes in a Trigger Circuit with a Transistor.—N. I. Brodovich. (*Avtomatika i Telemekhanika*, March 1957, Vol. 18, No. 3, pp. 273–279.) Transient processes in a common-base circuit using a point-contact transistor are considered. Expressions for the currents are derived and the speed of operation of the circuit is estimated. Requirements are formulated for point-contact transistors to be used in high-speed trigger circuits. See also 2381 of 1957 (Kuz'min).

621.373.444 : 621.314.7 **2015**

High-Sensitivity Transistor Pulse Trigger Circuit.—E. Zaglio. (*Nuovo Cim.*, 1st Sept. 1957, Vol. 6, No. 3, pp. 512–515. In English.) A Ge diode and a differential negative resistance of high stability consisting of a transistor amplifier with positive feedback is used.

621.373.52 **2016**

Transistorized *RC* Phase-Shift Power Oscillator.—L. J. Giacoletto. (*Trans. Inst. Radio Engrs*, May/June 1957, Vol. AU-5, No. 3, pp. 59–62.) Data on the performance of a *CR* phase-shift circuit operating at frequencies around 1 kc/s are given.

621.375.1 : 621.396.822 **2017**

A Theorem concerning Noise Figures.—A. G. Bose & S. D. Pezaris. (*Trans. Inst.*

Radio Engrs, Sept. 1956, Vol. CT-3, No. 3, pp. 190–196. Abstract, *Proc. Inst. Radio Engrs*, Feb. 1957, Vol. 45, No. 2, p. 254.)

621.375.121.1 **2018**

Stagger-Tuned Band-Pass Amplifiers.—Y. Peless. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 175–178.) A procedure is given for designing narrow-band amplifiers with specified gain and overshoot when either bandwidth or rise time are specified. Data are included for single-tuned cascades with up to five stages. See also 3774 of 1957 (Peless & Murakami).

621.375.126.016.35 **2019**

Stability and the Effects of Valve Input Conductance in Wide-Band I.F. Amplifiers.—F. Carassa. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 550–562.) An analysis of the effects of small variations of circuit parameters and of valve input conductance on amplitude and group-delay response of double-tuned interstage amplifiers. See also 2368 of 1957.

621.375.13 : 621.376.23 **2020**

Some Studies on Delayed Feedback Circuits.—H. Seki. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 758–763.) By inserting delayed signals along with the main signal to be observed, an improvement in signal/noise ratio can be achieved. Other applications of this circuit are for producing artificial reverberation [see 3476 of 1955 (Axon et al.)], as a narrow-band filter, and as a short-time storage circuit.

621.375.2 : 621.317.725 **2021**

Bootstrapped Differential Amplifier with Reduced Common-Mode Effects.—R. J. Blume. (*Rev. sci. Instrum.*, Feb. 1958, Vol. 29, No. 2, pp. 122–124.) A common-mode signal of ± 35 V changes the differential gain of the circuit by 0·5% or less. Full-scale deflection can be obtained on a 1-mA 1·5-kΩ recorder with 0·2 V d.c. input.

621.375.2.029.3 **2022**

Push-Pull Audio Amplifier Theory.—M. A. Melehy. (*Trans. Inst. Radio Engrs*, July/Aug. 1957, Vol. AU-5, No. 4, pp. 86–89.) A mathematical analysis is given applicable to all classes of operation of the push-pull amplifier assuming non-linear valve characteristics.

621.375.2.029.3 **2023**

Two-Valve Pre-amplifier.—C. Hardcastle. (*Mullard tech. Commun.*, Jan. 1958, Vol. 3, No. 28, pp. 254–256.) The performance of the amplifier is discussed when used with magnetic and crystal pickups, tape recorder play-back heads, and microphone and radio inputs.

621.375.2.029.63 **2024**

Straight P.A. for 70 Centimetres.—J. A. Plowman. (*Short Wave Mag.*, Jan. 1958, Vol. 15, No. 11, pp. 595–597.) Design and operating details for a r.f. amplifier on 430 Mc/s using a double tetrode.

621.375.23.029.3 **2025**

A Nonlinear Low-Frequency Instability Phenomenon in Audio Amplifiers.—T. Usher, Jr. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 698–701.) Under certain conditions, a.f. amplifiers with feedback may show instability at low frequencies, which is not predictable by conventional linear analysis. The problem is analyzed theoretically with supporting experimental results and a technique given for reducing the nonlinear effect.

621.375.232.3 **2026**

Some Augmented Cathode-Follower Circuits.—J. R. Macdonald. (*Trans. Inst. Radio Engrs*, May/June 1957, Vol. AU-5, No. 3, pp. 63–70.) Direct-coupled cathode-follower circuits with very high input and low output impedance are described, suitable for driver stages (see e.g. 2888 of 1955), buffer stages and frequency-selective amplifiers.

621.375.3 **2027**

Magnetic Amplifiers: Basic Principles and Applications.—L. W. Stammerjohn. (*Bell Lab. Rec.*, Jan. 1958, Vol. 36, No. 1, pp. 16–20.) A basic qualitative description of the principles of operation.

621.375.4 : 621.317.3 **2028**

1-kc/s Transistor High-Gain Tuned Amplifier.—R. A. Hall. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 192–195.) "Particular attention has been paid to stability of gain and bandwidth against transistor variations; over a period of 12 months the maximum gain of about 125 dB has changed by only 1·5 dB with no measurable change in bandwidth. The amplifier has been designed principally for use as a bridge detector and general-purpose amplifier, but provision has also been made for transistor noise-figure measurements by the addition of a stage at the input in which the transistor under test provides sufficient gain to make the effective noise figure of the amplifier negligible for most purposes."

621.375.4.029.3 **2029**

A Stable-Gain Transformer-Coupled Transistor A.F. Amplifier.—H. Kemhadjian. (*Mullard tech. Commun.*, Jan. 1958, Vol. 3, No. 28, pp. 245–251.) The amplifier is designed primarily for use with d.c. choppers giving a square-wave output at 400 c/s. Circuit diagrams are given and the performance is illustrated by graphs. Changes in the transformer design for use at other frequencies are discussed.

621.375.4.029.3 : 621.314.7 **2030**

A Transistorized Decade Amplifier for Low-Level Audio-Frequency Applications.—A. B. Bereskin. (*Trans. Inst. Radio Engrs*, Sept./Oct. 1957, Vol. AU-5, No. 5, pp. 138–142. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 800–801.) 1957 National Electronics Conference paper.

621.375.4.072 **2031**

A Generalized Theory of Transistor Bias Circuits.—H. Hellerman. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 694–697.) A general method of analysing bias circuits is presented. Almost all three-terminal bias circuits can be reduced to the same standard form for which the analysis is given. An example from current practice shows the unifying concepts resulting from the general theory.

621.375.4.123 **2032**

Transistor Amplifier Stages Operating from 6, 9 and 12 V.—O. J. Edwards. (*Mullard tech. Commun.*, Jan. 1958, Vol. 3, No. 28, pp. 251–253.) A description of low-level stages using a Type-OC71 transistor which can be used to build *RC*-coupled amplifiers operating in ambient temperatures up to 45°C.

621.375.422.029.3 **2033**

Direct-Coupled Transistor Audio Amplifier.—D. A. G. Tait. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 237–239.) Description of a design procedure for a three-stage amplifier in which the collector of one stage is directly connected to the base of the following stage. See 627 of February (Milnes).

621.375.9 : 538.569.4 : 538.221 **2034**

Parametric Amplification using Low-Frequency Pumping.—S. Bloom & K. K. N. Chang. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, p. 594.) Describes how low-frequency pumping is possible if two pumping sources are used in conjunction with a nonlinear reactance having an odd-order nonlinearity. An analysis of the equivalent circuit is given.

621.375.9 : 538.569.4 : 538.221 **2035**

A Travelling-Wave Ferromagnetic Amplifier.—P. K. Tien & H. Suhl. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 700–706.) Amplification of signal power can be obtained in a propagating structure which is partially or totally embedded in a ferromagnetic medium. It is shown that one form of the structure possesses two propagating modes and can also support a travelling wave supplied by a local oscillator which provides through the magnetic changes a time-varying coupling between the two propagating modes. The principle of operation is illustrated by a simple transmission-line model.

621.375.9 : 538.569.4.029.64 **2036**

Contribution to the Theory of the Molecular Generator.—Yu. L. Klimontovich & R. V. Khokhlov. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1150–1155.) The resonance interaction between an electromagnetic field and a molecular beam, and self-oscillatory processes in a molecular oscillator are investigated for the case of a single-velocity beam. Results obtained taking account of 'nonmonochromatic' molecules in the beam are considered qualitatively.

621.375.9 : 538.569.4.029.64 **2037**

Operation of Three-Level Solid-State Maser at 21 cm.—J. O. Artman, N. Bloembergen & S. Shapiro. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1392–1393.) A three-level paramagnetic solid-state maser has been operated both as an amplifier and as an oscillator at 1 373 Mc/s. A single crystal of about 1 cm^3 of $K_3Co(CN)_6$ containing 0·5% $K_3Cr(CN)_6$ as active element is used.

621.375.9 : 621.385.029.64 : 537.533 **2038**

Parametric Amplification of Space-Charge Waves.—Louisell & Quate. (See 2273.)

621.376.2 : 621.318.134 **2039**

The Magnetic Ring Modulator and Demodulator.—R. Elsner & L. Pungs. (*Nachrichtentech. Z.*, Sept. 1957, Vol. 10, No. 9, pp. 436–438.) Advantages and disadvantages are outlined with reference to the equivalent circuit and to a practical modulator for use with a 10-kc/s carrier frequency.

621.376.32 **2040**

Multiplication of Frequency-Modulated Oscillations over an Extremely Wide Frequency Range.—H. Schönfelder. (*Frequenz*, Sept. 1957, Vol. 11, No. 9, pp. 274–278.) The problem of multiplying in a ratio 1 : 1000 with a fundamental frequency of 100 kc/s is discussed with reference to the performance of a serrasoid modulator (279 of 1954). See also 1378 of May.

621.376.32 : 621.318.134 **2041**

Frequency Modulation by Inductance Variation: a Magnetically Stable Ferrite Modulator.—F. Slater. (*J. Brit. Instn Radio Engrs*, March 1958, Vol. 18, No. 3, pp. 189–204.) The operation of modulators incorporating a reactance valve or a variable inductance is summarized. The losses in the magnetic core of the inductance modulator are considerably reduced by using magnetic ferrites, and details of this type of modulator are given. It can be used in simple variable-frequency circuits up to 500 Mc/s.

621.376.32 : 621.318.134 **2042**

Simple Circuit Stabilizes Ferrite F.M. Modulator.—A. B. Przedpelski. (*Electronic Ind.*, Feb. 1958, Vol. 17, No. 2, pp. 56–57.) Stabilization over a range of frequencies is achieved by the use of an auxiliary fixed-frequency oscillator and associated feedback control circuit.

621.374.32 : 621.387.4 **2043**

Multichannel Pulse Height Analysers. [Book Review]—H. W. Koch & R. W. Johnston (Eds). Publishers: National Academy of Sciences—National Research Council, Washington, D.C., 1957, 205 pp., $2. (*Nature, Lond.*, 21st Dec. 1957, Vol. 180, No. 4599, p. 1381.) Proceedings of an informal conference held at Gatlinburg, Tennessee, 26th–28th September 1956.

621.375.4.029.3 **2044**

Transistor A.F. Amplifiers. [Book Review]—D. D. Jones & R. A. Hilbourne. Publishers: Iliffe & Sons, London, 1957, 152 pp., 21s. (*J. Electronics Control*, Jan. 1958, Vol. 4, No. 1, p. 96.)

GENERAL PHYSICS

530.12 **2045**

A Variational Principle for Classical Field Theories.—J. Pachner. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 353–368.)

530.12 : 537.21 **2046**

The Definition of Current Charge and Electric Force in the Unified Field Theory.—H. Treder. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 369–380.)

535.13 **2047**

On the Deduction of the Lorentz-Einstein Transformation from Maxwell's Electromagnetic Field Equations.—K. Stiegler. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 512–513.)

535.223 : 538.566.029.65 **2048**

Precision Determination of the Velocity of Electromagnetic Waves.—K. D. Froome. (*Nature, Lond.*, 25th Jan. 1958, Vol. 181, No. 4604, p. 258.) Using an interferometer of improved performance operating at 72 kMc/s a value $c_0 = 299\,792 \cdot 50 \pm 0 \cdot 10$ km/s has been obtained.

535.376 : 546.26-1 **2049**

Investigations by Microscope and Oscillograph of the Electroluminescence of Nonconducting Diamonds.—E. Krautz & G. Zollfrank. (*Optik, Stuttgart*, Oct. 1957, Vol. 14, No. 10, pp. 446–457.) The electroluminescence is investigated at field strengths above 10^6 V/cm. With very strong fields and high current densities graphite whiskers form in the diamond at the electrodes and the electroluminescence is thereby enhanced. Oscillograms indicate that recombination processes are a contributory cause of electroluminescence.

537.12 : 538.1 **2050**

Magnetic Interaction of Electrons and Anomalous Diamagnetism—V. T. Geĭlikman. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1206–1211.)

537.311.62 **2051**

On the Theory of Skin Effect in Metals.—M. Ya. Azbel'. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, p. 1259.) The coefficient of e.m. wave transmission through a sufficiently thick film is calculated taking into account a small slowly damped addition derived from the evaluation of the spin magnetic moment.

537.312.62 **2052**

Critical Current for Superconducting Films.—V. L. Ginzburg. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Jan. 1958, Vol. 118, No. 3, pp. 464–467.) Mathematical analysis of critical currents for superconducting films $10^{-5} - 10^{-6}$ cm thick.

537.525.6 **2053**

Emission Mechanism of Cold-Cathode Arcs.—K. G. Hernqvist. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 636–646.) A new theory is proposed in which excited atoms play a predominant role as a source for ion generation in the vicinity of the cathode surface. The processes of resonance ionization and of ionization in the strong electric field at the cathode surface are considered.

537.533 : 621.385.032.269.1 **2054**

A General Theorem for Dense Electron Beams.—A. R. Lucas, B. Meltzer & G. A. Stuart. (*J. Electronics Control*, Feb. 1958, Vol. 4, No. 2, pp. 160–164.) A sufficient condition for the existence of any particular set of trajectories is derived for Pierce-type dense electron beams.

537.533.79 : 537.226.3 : 538.561 **2055**

Energy Loss by a Charged Particle Passing through a Laminar Dielectric: Part 1.—Ya. B. Fainberg & N. A. Khizhnyak. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 883–895.) A general expression has been obtained for losses of a particle moving in an unbounded laminar medium or a waveguide filled with a dielectric. Polarization losses are examined and an expression for the spectral distribution of parametric Cherenkov radiation is derived.

537.56 **2056**

Electron-Electron Interaction and Heat Conduction in Gaseous Plasmas.—L. Goldstein & T. Sekiguchi. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 625–630.) The thermal conductivity in low-gas-pressure Ne and Xe plasmas has been shown experimentally to be determined chiefly by heat flow in the electron gas of the plasma. The experimentally determined values of the thermal conductivity are in agreement, within less than one order of magnitude, with those given by the theory of Spitzer & Härm (*ibid.*, 1st March 1953, Vol. 89, No. 5, pp. 977–981).

537.56 **2057**

Energy Exchange between Electron and Ion Gases through Coulomb Collisions in Plasmas.—A. A. Dougal & L. Goldstein. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 615–624.) The general aspects of energy exchange between the electron, ion and molecular constituent gases and their energy transfer to the boundary have been examined experimentally for a partially ionized gas. Coulomb collisions between electrons and ions are shown to contribute significantly to thermal energy transfer from the electron gas, even in weakly ionized gases. In a plasma produced in Ne gas at 2·25 mm Hg at 300°K, the characteristic time for equipartition of excess mean electron energy with the ion gas varies from 11 to 5 μs as the ion concentration increases from $1 \cdot 6$ to $5 \cdot 8 \times 10^{11}$ cm^{-3}.

537.56 **2058**

Theory of Electric Waves in Inhomogeneous Plasmas.—K. Kischel. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 309–321.) Dispersion functions are derived for longitudinal waves in plasmas of variable electron density. The special case of a quasi-neutral electron beam of spatially variable density is discussed.

537.56 **2059**

Simplification of Equations for the Distribution Function of Electrons in a Plasma.—A. V. Gurevich. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1237–1238.) An analysis based on the Boltzmann kinetic equation, considering initially a spatially homogeneous plasma located in electric and magnetic fields.

537.56 **2060**

Contribution to the Theory of Transport Processes in a Plasma Located in

a Magnetic Field.—E. S. Fradkin. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1176–1187.) The mean statistical characteristics such as velocity, heat flow and stress tensor are determined.

537.56 **2061**

Space-Charge, Field Emission and the Ionic Condenser.—H. Ritow. (*J. Electronics Control*, Feb. 1958, Vol. 4, No. 2, pp. 97–110.) High e.s. fields measured at the electrodes during gas discharge experiments in H indicate that the mechanism initiating a flash of intermittent glow is field emission from the cathode or from the negative space charge. The hypothesis is applied to the experimental results of other workers and is confirmed. It is shown to provide simple explanations for many gas discharge phenomena. It is suggested that the measured high fields are due to concentrated space charges forming ionic condensers with the electrodes or within the gas itself in the form of stationary striations.

537.56 : 535.33 **2062**

Statistical Broadening of Spectral Lines Emitted by Ions in a Plasma.—M. Lewis & H. Margenau. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 842–845.)

537.56 : 538.566 **2063**

Ponderomotive Force in Localized Plasma in the Electromagnetic Field of a Plane Wave.—V. V. Yankov. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 926–927.) A sphere of uniformly ionized gas is considered as a rough model of localized plasma. The radius of this sphere is smaller than the wavelength of the field in vacuum or in plasma. Expressions for the internal and surface intensities of the ponderomotive force are derived.

537.56 : 538.566 **2064**

Excitation of Plasma Oscillations and Growing Plasma Waves.—G. D. Boyd, L. M. Field & R. W. Gould. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1393–1394.) Report of preliminary results of a new experiment in which the beam is modulated by a microwave signal before it interacts with the plasma. A very strong interaction is observed which shows essentially the theoretically predicted form of behaviour.

537.56 : 538.63 **2065**

Movements of Rarefied Plasma in an Alternating Magnetic Field.—Ya. P. Terletskiĭ. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 927–928.) An expression is derived showing that even with constant current the energy of rotation of plasma particles can increase many times in consequence of the decrease of the radius.

538.114 **2066**

Contribution to the Theory of Anisotropy of Ferromagnetic Single Crystals.—N. A. Potapkov. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Jan. 1958, Vol. 118, No. 2, pp. 269–272.) A mathematical analysis based on Dyson's theory of spin-wave interaction (3696 of 1956) is applied to the investigation of anisotropic ferromagnets of hexagonal symmetry.

538.22 : 538.569.4 **2067**

Paramagnetic Resonance and Polarization of Nuclei in Metals.—M. Ya. Azbel', V. I. Gerasimenko & I. M. Lifshits. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1212–1225.) A theory of paramagnetic resonance is proposed based on simultaneous solutions of Maxwell's equations and the kinetic equation for the density operator. It is shown that the polarization varies with depth and also that the paramagnetic resonance gives rise to selective transparency in metallic films.

538.249 **2068**

Basic Features of Néel's Theory of Impurity-Atom After-Effects.—F. Schreiber. (*Z. angew. Phys.*, April 1957, Vol. 9, No. 4, pp. 203–212.) Exposition in simple terms of Néel's theory of magnetic diffusion after-effects (3072 of 1952). The equivalent circuit of a Bloch wall is derived which allows for the influence of impurity atoms. 33 references.

538.3 : 550.38 **2069**

Two-Dimensional Problems of the Decay of Magnetic Fields in Magnetohydrodynamics.—T. G. Cowling & A. Hare. (*Quart. J. Mech. appl. Math.*, Nov. 1957, Vol. 10, Part 4, pp. 385–405.) The normal modes of decay of magnetic fields in conducting fluids are studied for different classes of steady motion. See 1328 of 1955 (Bullard & Gellman).

538.56.029.64 : 531.61 **2070**

On a Macroscopic Measurement of the Spin of Electromagnetic Radiation.—G. Toraldo di Francia. (*Nuovo Cim.*, 1st July 1957, Vol. 6, No. 1, pp. 150–167. In English.) The spin of e.m. radiation may be deduced from the torque exerted at microwave frequencies by a circularly polarized wave upon a screen which can conduct current only parallel to a given direction [see 354 of 1950 (Carrara)]. Calculations are made of the angular-momentum and scattering cross-sections of the screen, and the feasibility of an accurate experiment is discussed.

538.561 : 537.122 **2071**

The Cherenkov Effect in Composite (Isotropic) Media.—A. M. Sayied. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 398–404.) A theoretical investigation based on the invariance of Maxwell's equations. The case of two coaxial dielectrics and permeable cylinders with a common cylindrical interface, and a charged particle moving along their common axis is considered. Results are discussed with particular reference to Cherenkov radiation at radio and microwave frequencies. The effect of coherence on the output of the emitted radiation is outlined.

538.561 : 537.533 **2072**

Radiation from a Point Charge Moving Uniformly along the Surface of an Isotropic Medium.—A. I. Morozov. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1260–1261.) Energy emitted per unit of time by a particle moving along a surface in a nondispersive medium is calculated. Expressions are derived for relativistic and nonrelativistic conditions and for magnetic and ferrite-type media. See also 1628 of 1955 (Danos).

538.566 : 535.42 **2073**

Diffraction by Cylindrical Reflectors.—R. Plonsey. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 312–318.) Measurements of the near-zone diffracted field of a cylindrical reflector with a line source at the centre agree with calculations using geometrical-optics current. The equivalent line currents give slightly better agreement.

538.566 : 537.562 **2074**

The Propagation of Slow Electromagnetic Waves along Inhomogeneous Plasma Layers.—W. Oehrl. (*Z. angew. Phys.*, April 1957, Vol. 9, No. 4, pp. 164–171.) On the basis of Schumann's work (see e.g. 712 of 1953) the special case of surface waves in the plasma and the adjoining layers of air is considered.

538.569.4 **2075**

Molecular-Beam Resonances in Oscillatory Fields of Nonuniform Amplitudes and Phases.—N. F. Ramsey. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 822–825.) The transition probability equations are reduced to forms which are suitable for digital computer calculations. A computer program for the calculation of the shapes of the resonances is described.

538.63 **2076**

Quantum Theory of Galvanomagnetic Effects.—P. N. Argyres. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1115–1128.) A study is made of the effect of the quantization of the electron orbits in a magnetic field on the galvanomagnetic properties of an isotropic semiconductor or semimetal in the phonon-scattering range. The conductivity tensor is calculated using the quantum-mechanical density operator. The Hall coefficient and transverse resistivity are studied for a number of different sets of conditions for electron density, magnetic field, and temperature.

539.2 **2077**

Theoretical Investigation of the Electronic Energy Band Structure of Solids.—F. Herman. (*Rev. mod. Phys.*, Jan. 1958, Vol. 30, No. 1, pp. 102–121.) A review including sections dealing with (*a*) the elements of energy-band theory; (*b*) the uses of group theory and perturbation theory; (*c*) a survey of numerical methods for solving the crystal wave equation; (*d*) examples illustrating complex energy-band structures. A representative list of about 200 references is given. Experimental techniques are discussed by Lax (2078 below).

539.2 **2078**

Experimental Investigations of the Electronic Band Structure of Solids.—B. Lax. (*Rev. mod. Phys.*, Jan. 1958, Vol. 30, No. 1, pp. 122–154.) Experiments are discussed which have been of value in contributing to current knowledge of the band structure of many materials of importance at present. The experiments mentioned include those involving the de Haas-van Alphen effect, cyclotron resonance, the galvanomagnetic effect, and infrared absorption. 172 references.

539.2 **2079**

Electron Interaction in Solids. General Formulation.—P. Nozières & D. Pines. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 741–761.) A general Hamiltonian formalism is developed to treat from first principles the motion of electrons in solids, including their mutual Coulomb interaction. It is shown that, under suitable circumstances, plasmons (the quanta of the plasma oscillations) represent a well-defined elementary excitation of the solid. The 'existence criterion' for plasmons is found to be a high electronic polarizability. After the plasmon modes are separated out, the remaining electron interaction is found to be screened, with a range of the order of the interelectron spacing. The usefulness of this effective Hamiltonian for the calculation of the electronic energy levels and cohesive energy in solids is discussed briefly.

539.2 **2080**

Electron Interaction in Solids. Collective Approach to the Dielectric Constant.—P. Nozières & D. Pines. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 762–777.) A quantum theory of the dielectric constant for solids of both low and high polarizability is developed from first principles. In the latter case, the approach used is collective in that the long-range part of the electron interaction is described by the plasmon field. Both the static and frequency-dependent dielectric constant are derived. It is shown that the interaction between electrons may be described in terms of the dielectric constant of the solid provided the electrons in question form a small minority group which can be isolated from the much larger majority electron group.

539.2 **2081**

Electron Interaction in Solids. The Nature of Elementary Excitations.—P. Nozières & D. Pines. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1062–1074.) Possible elementary excitations in solids are studied with the aid of the general theoretical approach developed in the preceding papers of this series (2079 and 2080 above). Particular attention is paid to the basic theoretical justification for the individual-particle-like elementary excitation ('effective' electrons). It is concluded that good qualitative arguments could now be given for the existence of effective electrons in solids. The presence of an energy gap is shown to be a necessary condition for the existence of strong spatial correlations between minority carriers in solids, and the nature of such correlated minority electron excitations is discussed.

539.2 : 538.221 : 621.318.134 **2082**

Approximate Theory of Ferrimagnetic Spin Waves.—T. A. Kaplan. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, p. 782–787.) "The wave functions and the energy spectrum for the spin-wave problem in a normal spinel are found by means of a straightforward extension of Anderson's approach to antiferromagnetism [*ibid.*, 1st June 1952, Vol. 86, No. 5, pp. 694–701]. Only *A-B* exchange is assumed to exist, and the calculation is carried to second order in the magnitude of the propagation vector, **k**. Five distinct energy surfaces (E *vs* k) are found, two of which are identical, in the classical limit, to the ones previously reported by H. Kaplan [*ibid.*, 1st April 1952, Vol. 86, No. 1, p. 121]."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164.3 : 621.396.11 **2083**

Propagation of Radio Waves from Cosmical Sources.—E. Chvojková. (*Nature, Lond.*, 11th Jan. 1958, Vol. 181, No. 4602, p. 105.) A formula is given for the refraction of radio waves reaching the earth's atmosphere from outside. The formula may be applied to waves from artificial earth satellites.

523.164.32 : 523.75 **2084**

A New Spectral Characteristic in Solar Radio Emission.—A. Maxwell & G. Swarup. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, pp. 36–38.) Equipment at Fort Davis, Texas, covers the band 100–580 Mc/s and comprises a 28-ft-diameter parabolic aerial with three broad-band primary arrays coaxially mounted at its focus [see 1126(*l*) of April (Maxwell et al.)]. A new type of fast burst having a frequency/time graph in the form of an inverted U and of duration 3–10 s is reported, and the possible origin of these bursts is discussed.

550.38 : 523.165 **2085**

Effective Magnetic Meridian for Cosmic Rays.—J. R. Storey, A. G. Fenton & K. G. McCracken. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, p. 34.) Observations of the cosmic-ray meridian made with neutron intensity monitors at Melbourne, Australia, are reported. Doubts are expressed about the equivalent-dipole theory of the earth's magnetic field [see e.g. 3721 of 1956 (Simpson et al.)].

550.389.2 : 629.19 **2086**

Scientific Observations of the Artificial Earth Satellites and their Analysis.—H. S. W. Massey & R. L. F. Boyd. (*Nature, Lond.*, 11th Jan. 1958, Vol. 181, No. 4602, pp. 78–80.) Report of a discussion held by the Royal Society on 29th November 1957.

550.389.2 : 629.19 **2087**

Lifetime of an Artificial Russian Satellite.—J. A. Fejer. (*Nature, Lond.*, 21st Dec. 1957, Vol. 180, No. 4599, p. 1413.) A simple formula for the calculation of lifetime from data on the initial orbit is given. It was applied to observations made at Johannesburg of the Doppler effect on signals received at 40·002 Mc/s from the first Russian satellite.

550.389.2 : 629.19 **2088**

Interception of Radio Signals Transmitted by the Satellite 'Sputnik I'.—P. F. Checcacci, V. Russo & C. Carreri. (*Ricerca sci.*, Nov. 1957, Vol. 27, No. 11, pp. 3252–3260.) Brief outline of apparatus and methods used ; some recordings of the 20- and 40-Mc/s transmissions obtained in Florence are reproduced.

550.389.2 : 629.19 **2089**

Observations in Australia of Radio Transmissions from the First Artificial Earth Satellite.—G. H. Munro & R. B. White. (*Nature, Lond.*, 11th Jan. 1958, Vol. 181, No. 4602, p. 104.) Observations of the strength of 20- and 40-Mc/s signals were made at Bringelly, near Sydney, Australia. The rate of fading appears to be a minimum when the direction of propagation of the wave is nearly perpendicular to the earth's magnetic field.

551.510.535 **2090**

Ionospheric Drift in the F_2 Region near the Magnetic Equator.—B. W. Purslow. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, pp. 35–36.) Graphs of the diurnal variations of drift in the F_2 layer based on observations made at Singapore from September 1953 to August 1956 suggest that the drift is predominantly in an E-W direction and has a maximum value of about 90 m/s.

551.594.5 : 621.396.11 : 551.510.535 **2091**

Artificial Aurora.—(*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 168–170.) The Luxembourg effect is explained, and the possibility of using it to produce a luminous discharge in the E layer is discussed.

551.594.5 : 621.396.96 **2092**

Determination of Auroral Height by Radar.—R. S. Unwin & M. Gadsden. (*Nature, Lond.*, 28th Dec. 1957, Vol. 180, No. 4600, pp. 1469–1470.) Preliminary report of observations made during the I.G.Y. at a radar station erected at 885 ft above sea level at Bluff in the South Island of New Zealand. Echoes have been recorded at 55 Mc/s from heights of about 110 km and ranges up to about 1 250 km to the south.

551.594.6 : 621.317.3 **2093**

Atmospheric Radio Noise.—Harwood & Nicolson. (See 2176.)

LOCATION AND AIDS TO NAVIGATION

621.396.663 **2094**

Cable Symmetry Requirements for Two-Channel Direction Finders.—G. Ziehm. (*Frequenz*, Sept. 1957, Vol. 11, No. 9, pp. 287–294.) The effects of differences in length of the cables connecting d.f. aerial systems to the receiver are investigated. Tolerances for these differences and for matching aerials to the receiver input are evaluated.

621.396.93+621.396.96].061.3 **2095**

Radio Navigational Aids.—(*Wireless World*, May 1958, Vol. 64, No. 5, pp. 210–211.) Digest of papers presented at an I.E.E. Convention held in London, 27th–28th March 1958.

621.396.933.2 : 621.396.677.6 **2096**

Study of the Feasability of Airborne H.F. Direction-Finding Antenna Systems.—P. S. Carter, Jr. (*Trans. Inst. Radio Engrs*, March 1957, Vol. ANE-4, No. 1, pp. 19–23. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1034.)

621.396.96 : 621.372.5 **2097**

Synthesis of Delay-Line Networks.—Linden & Steinberg. (See 2006.)

621.396.96 : 621.396.665 **2098**

The Design of Automatic-Gain-Control Systems for Auto-tracking Radar Receivers.—J. C. G. Field. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 93–108.) The a.g.c. requirements for conical-scanning-type radars and practical methods of control are reviewed. Design methods are illustrated by a description of the development of a system for a particular fire-control radar and the results of operational trials are briefly described.

621.396.969 : 629.13 **2099**

Effect of Precipitation on the Design of Radio Altimeters.—R. K. Moore. (*Trans. Inst. Radio Engrs*, March 1957, Vol. ANE-4, No. 1, pp. 24–29. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1034.) See also 1457 of 1957 (Moore & Williams).

621.396.969.3 **2100**

Atmospheric Angels Mimic Radar Echoes.—V. G. Plank. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 140–144.) The false echoes due to atmospheric conditions, insects, or birds are discussed. Echoes due to layers, clouds, wind-carried and other sources are considered, together with meteorological data. Unexplained traces which move rapidly across the screen and then disappear are also mentioned.

621.396.969.34 : 621.396.933.2 **2101**

The Air Traffic Control Radar Beacon System.—D. S. Crippen. (*Trans. Inst. Radio Engrs*, March 1957, Vol. ANE-4, No. 1, pp. 6-15. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1034.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.561 **2102**

Photoconductivity of Cuprite.—A. N. Krongauz, V. K. Lyapidevskiĭ & Yu. S. Deev. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1012–1017.) Investigations of the temperature dependence of photoconductivity indicate the occurrence of peaks for both positive and negative photoconductivity [see 2448 of 1954 (Krongauz & Lyapidevski)]. The combined effects of X rays and visible light depend on the sequence of the illumination processes.

535.215 : 546.817.221 : 539.23 **2103**

Quantum Efficiency of Photoconductive Lead Sulphide Films.—H. E. Spencer. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1074–1075.) By using photoconductivity measurements it is shown that the quantum efficiency of lead sulphide films is almost unity. Mobilities calculated from photoconductivity data agree with mobilities obtained from noise and Hall measurements. It is concluded that noise at room temperature is not photon noise.

535.215 : 546.863.221 **2104**

Structural Characteristics of Antimony Sulphide Layers.—V. N. Vertsner, B. V. Gorbunov & Ya. A. Oksman. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 957–961.) An electron-diffraction investigation shows that when a layer condenses on a heated backing it consists of an amorphous mass of Sb_2S_3 and a thin surface film of Sb_2O_3. The photosensitivity appears to be determined by the crystalline oxide which acts as an additional trapping centre for the current carriers.

535.37 : [546.472.21 + 546.482.21] : 535.215 **2105**

De-excitation of ZnS and ZnCdS Phosphors by Electric Fields. — H. Kallmann, B. Kramer & P. Mark. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 721–729.) The change of photoconductivity of powder samples brought about by external a.c. and d.c. fields, applied both during and after excitation with ultraviolet light, is described and compared with impedance changes due to irradiation with infrared light. The de-excitation and recovery phenomena observed are explained by a consideration of the distribution of conduction electrons created by the application of the field.

535.37 : 548.52 **2106**

The Synthesis of Single Crystals of the Sulphides of Zinc, Cadmium and Mercury and of Mercuric Selenide by Vapour-Phase Methods.—D. R. Hamilton. (*Brit. J. appl. Phys.*, March 1958, Vol. 9, No. 3, pp. 103–105.) Conditions of growth are examined in the light of recent theory.

535.376 **2107**

Electroluminescence of Zinc Sulphoselenide Phosphors with Copper Activator and Halide Coactivators.—I. J. Hegyi, S. Larach & R. E. Schrader. (*J. electrochem. Soc.*, Dec. 1957, Vol. 104, No. 12, pp. 717–721.)

535.376 **2108**

Electroluminescence and Field Effects in Phosphors.—H. F. Ivey. (*J. electrochem. Soc.*, Dec. 1957, Vol. 104, No. 12, pp. 740–748.) A review of the effects of applied electric fields on phosphors with or without other means of excitation. Possible explanations are given in terms of solid-state and semiconductor theories. 81 references.

535.376 : 546.681.18 **2109**

Grain Boundaries and Electroluminescence in Gallium Phosphide.—D. A. Holt, G. F. Alfrey & C. S. Wiggins. (*Nature, Lond.*, 11th Jan. 1958, Vol. 181, No. 4602, p. 109.) Light sources form a pattern of narrow lines and are identifiable with grain boundaries. See also 1433 of 1956 (Wolff et al.).

535.376 : 621.386.1 **2110**

Investigation of an X-Ray Tube with Control Grid. Application to the Study of Phosphorescence.—A. Pfahnal. (*Rev. gén. Élect.*, March 1957, Vol. 66, No. 3, pp. 159–186.) A comparison of the results of phosphor excitation by X rays and by electron beams is included.

537.226 : 538.566.2.029.63 **2111**

Metal-Flake Artificial Dielectric.—S. Swarup. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 179–182.) The validity of various expressions for the dielectric constant of obstacle-type dielectrics has been checked experimentally at 14·7 cm λ. A new expression for the metal-flake artificial dielectric compares well with experimental results. Some anomalous results are attributed to ferrous impurities in the aluminium powder used.

537.226 : 546.431.824 **2112**

Dielectric Properties of Single-Domain Crystals of $BaTiO_3$ at Microwave Frequencies.—T. S. Benedict & J. L. Durand. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1091–1093.) Describes the measurement of the small-signal dielectric constant at 24 kMc/s from 25°C to 170°C and the measurement of the effect of a d.c. field on the dielectric constant at that frequency.

537.226/.227 : 546.431.824-31 **2113**

The Effect of Small Additions of Magnesia on some High-Permittivity Ceramics Based on Barium Titanate.—K. W. Plessner & R. West. (*J. Electronics Control*, Jan. 1958, Vol. 4, No. 1, pp. 51–57.) "The presence of magnesia in barium titanate and its solid solutions (e.g. with calcium zirconate or calcium stannate) greatly improves the sintering of the ceramic, the pore space being reduced to about 1% in some cases. At the same time the permittivity maximum is flattened and displaced to lower temperatures by about 70°C for 0·5% magnesia. It is shown that the magnesium probably enters the lattice at both barium and titanium sites."

537.227 : 547.476.3 **2114**

Plotting the Thermodynamic Potential of Rochelle Salt from Results of an Optical Investigation of Domains.—V. L. Indenbom & M. A. Chernysheva. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 697–701.)

537.227 : 547.476.3 : 534.213-8 **2115**

Sound Absorption with Phase Transformation in Rochelle Salt.—I. A. Yakovlev, T. S. Velichkina & K. N. Baranskiĭ. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 935–936.) An investigation of plates of Rochelle salt using

1–5-μs pulses at 5 Mc/s. The variation of absorption with temperature is shown graphically.

537.228.1 : 548.0].001.4(083.7) **2116**

I.R.E. Standards on Piezoelectric Crystals: Determination of the Elastic, Piezoelectric, and Dielectric Constants—the Electromechanical Coupling Factor, 1958.—(*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 764–778.) Standard 58 I.R.E. 14.S1.

537.311.33 **2117**

The Theory of Electrical Conductivity of Semiconductors at High Frequencies: Part 1.—H. Stolz. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 394–400.) Boltzmann's equation is solved for a harmonic time-dependent electric field, allowing for the depolarization effects encountered in measurements of conductivity by means of cavity resonators. Results obtained for *n*-type Ge at cm λ are discussed.

537.311.33 **2118**

Recombination of Excess Carriers in Semiconductors.—J. Okada. (*J. phys. Soc. Japan*, Dec. 1957, Vol. 12, No. 12, pp. 1338–1344.) A multiple-level model for recombination is discussed. The results with *n*-type Ge samples are explained in terms of a recombination level at 0·30 eV below the conduction band and a hole trapping level at 0·30 eV above the valence band. For samples with a high dislocation density, the minority-carrier lifetime is found to vary with the dislocation density.

537.311.33 **2119**

Theory of the Enriched Barrier Layer on the Surface of Semiconductors.—G. Heiland. (*Z. Phys.*, 4th March 1957, Vol. 148, No. 1, pp. 28–33.) The model for which the differential equation is solved consists of a surface layer of donors on an *n*-type semiconductor, such as is formed by absorption or reaction (see e.g. 2140 below). Both low and high electron densities are considered.

537.311.33 **2120**

A Possible Mechanism for the Increase of Conductivity of Atomic Semiconductors in a Strong Electric Field.—F. G. Bass. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 863–865.) The effect on conductivity of the slowing down of electron recombination by a strong electric field is examined.

537.311.33 **2121**

Remarks on the Theory of the Electron Plasma in Semiconductors.—V. L. Bonch-Bruevich. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1092–1097.) The scattering of current carriers by charged impurities is examined in the light of the many-electron theory of semiconductors. At low temperatures plasma screening has a considerable effect on scattering. The ionization energy of impurity centres depends on their concentration, and decreases when the latter increases.

537.311.33 **2122**

Special Problems of Depletion-Layer Semiconductors—M. Falter. (*NachrTech.*, Sept. 1957, Vol. 7, No. 9, pp. 405–411.) Survey of theory dealing with the influence of lattice defects and surface recombination on semiconductor characteristics.

537.311.33 **2123**

Transient Recombination of Excess Carriers in Semiconductors.—G. K. Wertheim. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1086–1091.) The recombination equations for a system containing an arbitrary number of Shockley-Read recombination centres (see 420 of 1953) are formulated. Transient solutions are obtained for the decay following the injection of a pulse of carriers into a system containing one or two centres. The specific cases considered include (*a*) the one-centre case, (*b*) the case of a recombination centre with a temporary trap, and (*c*) the case of two recombination centres.

537.311.33 **2124**

Electron-Hole Recombination Statistics in Semiconductors through Flaws with Many Charge Conditions.—Chih-Tang Sah & W. Shockley. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1103–1115.) Treatment of the nonequilibrium but steady-state statistics for holes and electrons in semiconductors with one type of multiple-charge-condition flaw. The problem is simplified by the use of a charge-distribution diagram and a recombination-rate diagram. Through these diagrams the most important charge states of the flaws can be visualized for any injection conditions or carrier densities.

537.311.33 : 538.632 **2125**

The Problem of Saturation of the Hall 'Constant' in Semiconductors in Strong Magnetic Fields.—F. G. Bass & M. I. Kaganov. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1233–1235.) An expression is derived for the Hall coefficient applicable to semiconductors with narrow energy bands.

537.311.33 : 546.23 : 537.226.2 **2126**

Dielectric Constants of Amorphous and Crystalline Selenium at 3·3-cm Wavelength.—F. Eckart & H. Rabenhorst. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 381–393.)

537.311.33 : [546.28 + 546.289 **2127**

Intrinsic Optical Absorption in Germanium-Silicon Alloys.—R. Braunstein, A. R. Moore & F. Herman. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 695–710.) The absorption spectrum has been measured as a function of temperature and composition over the entire composition range. The absorption near the threshold ($K < 100$ cm^{-1}) exhibits a temperature dependence which is characteristic of phonon-assisted indirect electronic transitions. There is no temperature-independent component attributable to disorder-assisted transitions. When the phonon contribution to the absorption is explicitly taken into account, as in Macfarlane & Roberts' analysis (e.g. 3646 of 1955), the experimental data yield an equivalent phonon temperature which varies from 270 ± 20°K for pure Ge to 550 ± 50°K for pure Si.

537.311.33 : 546.28 **2128**

Measurement of Ternary Distribution Coefficients in Silicon.—D. Navon. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 579–582.) "A method is described for determining the distribution coefficient of an impurity introduced into a semiconductor by a fused metal contact. Coefficients for the segregation of antimony, arsenic, gallium and aluminium from a silver contact to silicon at 1 250°C are less than at the silicon melting point (1 408°C). Gold-doped contacts to silicon yield similar values for the distribution coefficient of these impurities."

537.311.33 : 546.28 : 537.312.8 **2129**

Weak-Field Magnetoresistance in *p*-Type Silicon.—D. Long & J. Myers. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1098–1102.) Measurements of the three weak-field magnetoresistance coefficients and the Hall mobility at a number of different temperatures between 77°K and 350°K have been made on *p*-type samples ranging in resistivity from 0·15 to 115 Ω.cm. The results indicate a marked temperature dependence of anisotropy in respect of energy band structure and/or scattering.

537.311.33 : 546.28 : 537.533 **2130**

Electron Emission from Silicon *p-n* Junctions.—J. Tauc. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, p. 38.) In addition to the emission of visible light from the breakdown region of a *p-n* junction during reverse current flow [see e.g. 3096 of 1956 (Chynoweth & McKay)], electron emission has been detected using a Geiger counter operating at 2 000 V several mm from the junction.

537.311.33 : 546.28 : 539.185.9 **2131**

Effect of Neutron Irradiation on Infrared Absorption in Silicon.—W. G. Spitzer & H. Y. Fan. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 1011–1012.) It was found that fast-neutron irradiation produced an absorption band at 20·5μ. An absorption band at 3·9μ was also observed.

537.311.33 : 546.28 : 669.046 **2132**

Heat Treatment of Silicon using Zone-Heating Techniques.—H. C. Theuerer, J. M. Whelan, H. E. Bridgers & E. Buehler. (*J. electrochem. Soc.*, Dec. 1957, Vol. 104, No. 12, pp. 721–723.) Donor contamination and changes in lifetime are not necessarily introduced by heat treatment, but may be due to contaminants still left after normal etching and washing processes.

537.311.33 : 546.289 **2133**

Lifetime Measurements of Minority Carriers Across and Along a Dislocation Wall in a Germanium Crystal.—R. R. Hasiguti & E. Matsuura. (*J. phys. Soc. Japan*, Dec. 1957, Vol. 12, No. 12, pp. 1347–1351.) The spacing of dislocations in the wall was about 2×10^{-4} cm; the recombination diameter $2 \cdot 2 \times 10^{-8}$ cm. The lifetime measured along the wall was 210 μs in a sample of mean lifetime 70 μs. It is suggested that the dislocation wall formed a thin *p*-type layer in an *n*-type crystal.

537.311.33 : 546.289 **2134**

Radial Variation of Minority-Carrier Lifetime in Vacuum-Grown Germanium Single Crystals.—C. A. Hogarth & P. J. Hoyland. (*J. Electronics Control*, Jan. 1958, Vol. 4, No. 1, pp. 60–62.) Radial variation is found to exist with vacuum-grown crystals, but not with those grown in an inert gas; this has already been found in Si crystals.

537.311.33 : 546.289 **2135**

Lifetime of the O+ Excited States in Ge[70].—H. W. Kendall. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 861–862.)

537.311.33 : 546.289 **2136**

Lifetime Measurements of Minority Carriers in Deuteron-Irradiated Germanium Crystals.—R. R. Hasiguti, E. Matsuura & S. Ishino. (*J. phys. Soc. Japan*, Dec. 1957, Vol. 12, No. 12, pp. 1351–1354.) Very long lifetimes were measured on the deuteron-irradiated surfaces of *n*-type crystals which, it is suggested, is due to the formation of a thin *p*-type layer.

537.311.33 : 546.289 : 538.63 **2137**

Investigation of the Longitudinal and Transverse Galvanomagnetic Effect in *n*-Type Germanium Single Crystals Cut along the Main Crystallographic Axes.—R. G. Annaev & A. Allanazarov. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Jan. 1958, Vol. 118, No. 1, pp. 47–50.) Experiments carried out at 29°C on pure *n*-type Ge single crystals show how the conductivity and the galvanomagnetic effect depend on the direction of the crystallographic axes. Results for crystals cut along [100], [110] and [111] axes are shown graphically.

537.311.33 : 546.36.87 : 539.23 **2138**

The Optical Absorption, the Photoelectric and Thermal Emission and the Temperature Dependence of the Conductivity of Caesium-Bismuth Layers of Differing Composition.—H. G. Clerc & G. Wallis. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 344–352.)

537.311.33 : 546.41.786 **2139**

Trap Distribution in Calcium Tungstate Single Crystals.—J. R. Cook. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 422–429.) Thermoluminescence and current 'glow' curves were used in the investigation. The resulting trap distribution agrees reasonably well with the results of a theoretical analysis of the photocurrent and phosphorescence decays after irradiation with ultraviolet light or gamma rays from a Co^{60} source.

537.311.33 : 546.47-31 **2140**

The Influence of Hydrogen on the Electrical Conductivity at the Surface of Zinc Oxide Crystals.—G. Heiland. (*Z. Phys.*, 4th March 1957, Vol. 148, No. 1, pp. 15–27.) The formation under the influence of hydrogen of two types of high-conductivity layer on ZnO crystals is discussed. For experimental investigation of the influence of an oxygen atmosphere, see 1105 of 1956.

537.311.33 : 546.472.21 : 548.0 **2141**

Electronic Energy Bands in ZnS: Potential in Zincblende and Wurtzite.—J. L. Birman. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 810–817.) The nature of the chemical bond in ZnS is discussed and evidence indicating mixed covalent and ionic bonding is reviewed. Calculation shows the crystal potential to be the same for both modifications in two corresponding prominent crystallographic directions.

537.311.33 : 546.472.21 : 548.0 **2142**

Electronic Energy Bands in ZnS: Preliminary Results.—C. Shakin & J. Birman. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 818–819.) "Preliminary results of the cellular calculations of electronic energy bands in cubic ZnS have been obtained at three points in the Brillouin zone: Γ, X, Λ. These results indicate that there is a normal order of states in the valence band, and an energy gap of 6–8 V at $\mathbf{k} = (0,0,0)$ corresponding to the $\Gamma_4' \to \Gamma_1$ transition." See also 2141 above.

537.311.33 : [546.623.86+546.682.86 **2143**

Influence of the Surface Area on the Type of Conductivity in AlSb and InSb.—V. F. Synorov. (*Dokl. Ak. Nauk S.S.S.R.*, 21st Jan. 1958, Vol. 118, No. 3, pp. 483–484.) The dependence of the conductivity on grain size is investigated. For grain sizes 2–3 μ the conductivity is very low and with a decrease in grain size conductivity changes from *n* type to *p* type.

537.311.33 : [546.681.19+546.682.19 **2144**

Nuclear Magnetic Resonance in Semiconductors: Part 3—Exchange Broadening in GaAs and InAs.—R. G. Shulman, B. J. Wyluda & H. J. Hrostowski. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 808–809.) "Nuclear magnetic resonance lines have been observed for the more abundant isotopes of the semiconductors GaAs and InAs. The resonances are broader than expected from nuclear dipolar widths alone. The additional broadening is explained by the indirect nuclear exchange mechanism and is consistent with previous measurements on the homologous semiconductors InSb and GaSb." Part 1: 1107 of 1956 (Shulman et al.).

537.311.33 : 546.681.19 : 546.289 **2145**

Some Properties of Gallium Arsenide-Germanium Mixtures.—D. A. Jenny & R. Braunstein. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 596–597.) The solubility of Ge in GaAs is probably less than two atomic percent in spite of the near identity of their crystal parameters. The presence of the maximum soluble amount of Ge reduces the band gap 1·35 eV by about 0·1 eV.

537.311.33 : 546.682.18 **2146**

Some Properties of Semiconducting Indium Phosphide.—W. N. Reynolds, M.T. Lilburne & R. M. Dell. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 416–421.) "Pure polycrystalline samples of InP have been prepared. Point-contact experiments have shown useful rectification on both *n*- and *p*-type samples, as well as transistor effects with power gains up to 20 times. The infrared transmission spectrum has been extended to 20 μ, and the effective mass of electrons has been estimated as 0·02 m. Previous work on general properties has been revised and extended."

537.311.33 : 546.682.23 **2147**

The Indium-Selenium System.—J. C. Brice, P. C. Newman & H. C. Wright. (*Brit. J. appl. Phys.*, March 1958, Vol. 9, No. 3, pp. 110–111.) Polycrystalline specimens of In_2Se_3, InSe and In_2Se were prepared by direct fusion, and their resistivity and spectral absorption characteristics were measured. The first two were found to be semiconductors, but single-phase specimens of the last were not achieved.

537.311.33 : 546.682.86 **2148**

Avalanche Multiplication and Electron Mobility in Indium Antimonide at High Electric Fields.—A. C. Prior. (*J. Electronics Control*, Feb. 1958, Vol. 4, No. 2, pp. 165–169.) The *I/V* characteristic of InSb has been measured up to field strengths of 800 V/cm. Avalanche multiplication was found to occur at values above 150 V/cm. No variation of electron mobility was found.

537.311.33 : 546.682.86 : 537.312.8 **2149**

Magnetoresistance of *n*-Type InSb at 4·2°K.—R. F. Broom. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 470–475.) Investigations of specimens prepared by different techniques from single crystals of differing purity indicate that negative magnetoresistance is not a bulk property of the material but is largely due to the method of preparation. Oscillations in the magnetoresistance as a function of magnetic field have been observed in samples with a balanced donor concentration of less than 10^{16} cm^{-3}. Results are discussed in detail for two samples.

537.311.33 : [546.812.241 + 546.812.231 + 546.682.87 **2150**

Electrical Properties of SnTe, SnSe and InBi at Low Temperatures.—K. Hashimoto. (*J. phys. Soc. Japan*, Dec. 1957, Vol. 12, No. 12, p. 1423.)

537.311.33 : 546.873.241 **2151**

Galvanomagnetic Effects in *n*-Type Bismuth Telluride.—J. R. Drabble, R. D. Groves & R. Wolfe. (*Proc. phys. Soc.*, 1st March 1958, Vol. 71, No. 459, pp. 430–443.) The resistivity, Hall coefficients and low-field magnetoresistance coefficients associated with current flow in the cleavage planes have been measured at 77°K, and over the range 100–300°K in the case of resistivity and Hall coefficients. Experimental results are reasonably consistent with a many-valley model of the band structure in which the energy minima are situated on the reflection planes. Evaluation of parameters of this model leads to relations between the two Hall coefficients and the density of carriers, which are used to obtain the conductivity mobility of electrons for current flow in the cleavage planes.

537.311.33 : 621.314.63 : 537.52 **2152**

Avalanche Breakdown Voltage in Hemispherical (*p-n*) Junctions.—J. Shields. (*J. Electronics Control*, Jan. 1958, Vol. 4, No. 1, pp. 58–60.) This voltage is calculated and a critical radius found at which a difference from a plane junction sets in; curves are given for Si *p-n* junctions. See also 3681 of 1957 (Armstrong et al.) and 471 of 1956 (Miller).

537.32 + 536.2 **2153**

Thermal Conductivity and Thermoelectric Phenomena in Metals in a Magnetic Field.—M. Ya. Azbel', M. I. Kaganov & I. M. Lifshits. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1188–1192.) Asymptotic expressions for the tensors of the thermal conductivity and Thomson coefficients in a strong magnetic field are derived.

538.22 **2154**

Magnetic Properties of Oxide of Manganese at Temperatures from 20° to 300°K.—A. S. Borovik-Romanov & M. P. Orlova. (*Zh. eksp. teor. Fiz.*, May 1957, Vol. 32, No. 5, pp. 1255–1256.) Graphs show the reciprocal of the magnetic susceptibility as a function of temperature for natural samples of Mn_2O_3 and Mn_3O_4.

538.22 : 538.569.4 : 549.517.13 **2155**

Maser Action in Ruby.—G. Makhov, C. Kikuchi, J. Lambe & R. W. Terhune. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1399–1400.) Brief report on the result. of an investigation of the electron-spin resonance properties of ruby; evidence of oscillations and amplification was obtained

538.221+621.318.1 **2156**

Conference on Magnetism and Magnetic Materials.—(*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 237–548.) The texts are given of some 120 papers presented at the conference held in Washington, D.C., 18th–21st November 1957, with abstracts of others and index of authors. Papers are grouped under 13 headings, including magnetization reversal and thin films, small particles and permanent magnets, ferromagnetic resonance, magnetic moments and crystal structure of oxides, applications and testing, anisotropy and magnetostriction, magnetization processes, and magnetic apparatus and techniques. Titles of some of the papers are given below; a few others are abstracted separately.

(*a*) Spontaneous Magnetization of some Garnet Ferrites and the Aluminium-Substituted Garnet Ferrites.—R. Pauthenet (pp. 253–255).

(*b*) Magnetic Resonance of Ferrites with a Compensation Temperature.—J. Paulevé (pp. 259–263).

(*c*) Static and Dynamic Behaviour of Thin Permalloy Films.—D. O. Smith (pp. 264–273).

(*d*) Flux Reversal in Thin Films of 82% Ni, 18% Fe.—C. D. Olson & A.V. Pohm (pp. 274–282).

(*e*) Transverse Flux Change in Soft Ferromagnetics.—F. B. Humphrey (pp. 284–285).

(*f*) Effects of Heat Treatment of Thin Ferromagnetic Films at Intermediate Temperatures.—E. N. Mitchell (pp. 286–287).

(*g*) Reversible Rotation in Magnetic Films.—R. M. Sanders & T. D. Rossing (pp. 288–289).

(*h*) Steady-State and Pulse Measurement Techniques for Thin Magnetic Films in the V.H.F.-U.H.F. Range.—D. O. Smith & G. P. Weiss (pp. 290–291).

(*i*) Ferromagnetic Resonance and Nonlinear Effects in Yttrium Iron Garnet.—R. C. LeCraw, E. G. Spencer & C. S. Porter (pp. 326–327).

(*j*) Substitution for Iron in Ferrimagnetic Yttrium-Iron Garnet.—M. A. Gilleo & S. Geller (pp. 380–381).

(*k*) Measurement of Losses of Magnetic Materials at High Inductions at Frequencies up to 100 Megacycles.—I. Bady (pp. 393–394).

(*l*) Behaviour of the TE Modes in Ferrite-Loaded Rectangular Waveguide in the Region of Ferrimagnetic Resonance.—W. J. Crowe (pp. 397–398).

(*m*) Energy Distribution in Partially Ferrite-Filled Waveguides.—J. E. Tompkins (pp. 399–400).

(*n*) Miniaturized Resonant Antenna using Ferrites.—D. M. Grimes (pp. 401–402).

(*o*) Appraisal of Permanent-Magnet Materials for Magnetic Focusing of Electron Beams.—M. S. Glass (pp. 403–404).

(*p*) Origin and Use of Instabilities in Ferromagnetic Resonance.—H. Suhl (pp. 416–421).

(*q*) Microwave Frequency-Conversion Studies in Magnetized Ferrites.—E. N. Skomal & M. A. Medina (pp. 423–424).

(*r*) Spin-Lattice Relaxation Time in Yttrium Iron Garnet.—R. T. Farrar (p.. 425–426).

(*s*) Ferrimagnetic Resonance in Gadolinium Iron Garnet.—B. A. Calhoun, W. V. Smith & J. Overmeyer (pp. 427–428).

(*t*) Ferromagnetic Resonance in Yttrium Iron Garnet at Low Frequencies.—E. G. Spencer, R. C. LeCraw & C. S. Porter (pp. 429–430).

(*u*) Microwave Properties of Polycrystalline Rare-Earth Garnets.—M. H. Sirvetz & J. E. Zneimer (pp. 431–433).

(*v*) Temperature Dependence of Microwave Permeabilities for Polycrystalline Ferrite and Garnet Materials.—J. Nemarich & J. C. Cacheris (pp. 474–476).

(*w*) Effect of Cobalt on the Relaxation Frequency of Nickel-Zinc Ferrite.—F. J. Schnettler & F. R. Monforte (pp. 477–478).

(*x*) Switching Properties of Permalloy Cores of Varying Coercivity.—T. D. Rossing & V. J. Korkowski (pp. 479–480).

(*y*) Operating Range of a Memory using Two Ferrite Plate Apertures per Bit.—M. M. Kaufman & V. L. Newhouse (pp. 487–488).

(*z*) Further Development of the Vibrating-Coil Magnetometer.—K. Dwight, N. Menyuk & D. Smith (pp. 491–492).

(*aa*) Improved Torque Magnetometer.—W. S. Byrnes & R. G. Crawford (pp. 493–495).

(*bb*) Transparent Ferromagnetic Light Modulator using Yttrium Iron Garnet.—C. S. Porter, E. G. Spencer & R. C. LeCraw (pp. 495–496).

(*cc*) Design of Optimum Inductors using Magnetically Hard Ferrites in Combination with Magnetically Soft Materials.—J. T. Ludwig (pp. 497–499).

(*dd*) Pulse Generator based on High Shock Magnetization of Ferromagnetic Material.—R. W. Kulterman, F. W. Neilson & W. B. Benedick (pp. 500–501).

(*ee*) New Magnetic Core Loss Comparator.—R. E. Tompkins, L. H. Stauffer & A. Kaplan (pp. 502–503).

(*ff*) Skullcap Method for Magnetizing Bowl-Shaped Magnetron Magnets.—F. X. MacDonough (pp. 506–507).

(*gg*) Magnetization Reversal by Rotation.—P. R. Gillette & K. Oshima (pp. 529–531).

(*hh*) Optical Properties of Several Ferrimagnetic Garnets.—J. F. Dillon, Jr (pp. 539–541).

538.221 **2157**

Domain Wall Motion in Metals.—R. W. DeBlois. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 459–467.) Ferromagnetic domain wall theory is reviewed and a description is given of experimental results of an extension of the Sixtus-Tonks experiment to axial domain wall velocities over 50 km/sec, made possible by use of nearly perfect iron whiskers. Plots of axial velocity *vs* applied field are nonlinear and in some cases show velocity discontinuities.

538.221 **2158**

The Effect of Added Titanium and Aluminium on the Magnetic Behaviour of α Ferric Oxide.—G. Haigh. (*Phil. Mag.*, April 1957, Vol. 2, No. 16, pp. 505–520.) The antiferromagnetic transition at −15°C is still observed in synthesized materials for relatively large impurity contents (≈10%). The magnetic properties of these materials are compared with those of naturally occurring haematites.

538.221 **2159**

Magnetic Domains in Manganese-Antimony.—R. Perthel & W. Andrä. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 265–267.) Domain patterns in Mn_2Sb and MnSb obtained by the Bitter method are reproduced.

538.221 **2160**

Magnetic Scattering of Thermal Neutrons of a Ferromagnetic or Antiferromagnetic Material near the Curie Point.—M. A. Krivoglaz. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Jan. 1958, Vol. 118, No. 1, pp. 51–54.) Mathematical treatment of the thermodynamic theory of magnetic scattering of neutrons.

538.221 **2161**

Electron-Microscope Investigation of Magnetic Powder Patterns.—W. Schwartze. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 322–328.) The relation of magnetic structure to powder pattern is investigated and a method of preparing specimens is described.

538.221 **2162**

A Magnetic Heating Effect in Oxidized Nickel Wires.—C. Schwink. (*Naturwissenschaften*, Sept. 1957, Vol. 44, No. 18, pp. 485–486.) An application of the electron-optical shadow technique (1515 of 1956) showing the variation of the position of the effective magnetic pole in Ni wires under strain, after heating with and without an alternating field at temperatures below the Curie point.

538.221 : 537.312.8 **2163**

Magnetoresistance Coefficients and their Temperature Dependence in Iron and Silicon-Steel.—E. Tatsumoto. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 658–662.)

538.221 : 538.569.4 **2164**

Resonant Modes of Ferromagnetic Spheroids.—L. R. Walker. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 318–323.) The spectrum of modes of oscillation of a ferromagnetic spheroid situated in a uniform d.c. magnetic field is discussed. It is shown that, for samples of normal size, a part of the spectrum consists of magnetostatic modes for which exchange and propagation may be ignored. The analysis presented satisfactorily describes the observed absorption curves.

538.221 : 538.632 **2165**

Hall Effect in Ni_3Mn and Fe-Co as a Function of Order.—S. Foner, F. E. Allison & E. M. Pugh. (*Phys. Rev.*, 15th Feb. 1958, Vol. 109, No. 4, pp. 1129–1133.) In Ni_3Mn the values of R_0* were found to change uniformly from a negative value for a well ordered sample to a positive value for a disordered sample. This dependence is associated with a Curie-point anomaly. In Fe-Co the values of R_0* were found insensitive to the degree of atomic ordering and were related to the band structure.

538.221 : 538.652 **2166**

A Highly Elastic Magnetostrictive Material.—O. Henkel. (*NachrTech.*, Aug. 1957, Vol. 7, No. 8, pp. 346–350.) Report of investigations on cylindrical specimens consisting of Fe particles suspended in rubber-like substances.

538.221 : 538.652 : 539.37 **2167**

Plastic Deformation and Magnetostriction.—G. Rieder. (*Z. angew. Phys.*, April 1957, Vol. 9, No. 4, pp. 187–202.) A combined theory covering internal stresses due to plastic deformation and magnetostriction is developed. Its practical application with reference to investigations of magnetic properties [e.g. 1825 of 1957 (Kersten) and 2215 of 1957 (Dietrich & Kneller)] gives satisfactory results. 32 references.

538.221 : 539.1 **2168**

Polarization of Positrons and Annihilation in Ferromagnetic Materials.—S. S. Hanna & R. S. Preston. (*Phys. Rev.*, 1st Feb. 1958, Vol. 109, No. 3, pp. 716–720.) The angular correlation of two-quantum radiation from annihilation of positrons in ferromagnetic media has been investigated as a function of the direction of magnetization. The observed changes in the correlations are attributed to the polarization of positrons emitted from an unpolarized source.

538.221 : [621.318.124+621.318.134 **2169**

Theory of Magnetostriction and *g*-factor in Ferrites.—N. Tsuya. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, March 1957, Vol. 8, No. 4, pp. 161–255.) The theory is based on the Heitler-London approximation. The conclusions are that four mechanisms are mainly responsible for magnetostriction: (*a*) the interplay of spin-orbit interaction with the strain potential; (*b*) classical dipolar interaction; (*c*) the interplay of anisotropic exchange interaction with the strain potential; (*d*) the interplay of intra-spin-spin interaction with the strain potential. The relative importance of the different mechanisms in various ferrites is discussed. 69 references.

538.221 : 621.318.124 **2170**

The Temperature Dependence of the Magnetic Properties of Barium Ferrite.—E. Schwabe. (*Z. angew. Phys.*, April 1957, Vol. 9, No. 4, pp. 183–187.) The characteristics of magnets composed of anisotropic $BaO.6Fe_2O_3$ are investigated. Tests were made on a loudspeaker magnet assembly to determine the demagnetizing forces and to derive the means of minimizing the effects of low-temperature storage on operation. See also 506 of 1957 (Sixtus et al.).

538.222 : 538.569.4 **2171**

Paramagnetic Absorption at Very High Frequencies of some Mn Salts Located in Parallel Fields.—A. I. Kurushin. (*Zh. eksp. teor. Fiz.* April 1957, Vol. 32, No. 4, pp. 938–939.)

538.222 : 538.569.4 **2172**

Paramagnetic Absorption at High Frequencies in Gadolinium Salts in Parallel Fields.—A. I. Kurushin. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 727–730.) The internal field constant, the isothermal spin relaxation time and the dependence of the absorption coefficient on the stationary field strength are derived.

548.0 : 53 **2173**

Electrical, Optical and Elastic Properties of Diamond-Type Crystals: Part 2.—V. S. Mashkevich. (*Zh. eksp. teor. Fiz.*, April 1957, Vol. 32, No. 4, pp. 866–873.) The vibration spectrum for long waves is investigated. Part 1: 3940 of 1957 (Mashkevich & Tolpygo).

621.315.616 : 539.56 **2174**

Brittleness in Polyethylene.—I. L. Hopkins. (*Bell Lab. Rec.*, Jan. 1958, Vol. 36, No. 1, pp. 5–8.) A qualitative description of the molecular structure responsible for brittleness, and of tests designed to measure material characteristics as the rupture point is approached.

MATHEMATICS

517.5 **2175**

Coefficients for 'Decomposition' of Functions into Laguerre-Function Series.—J. W. Head & G. M. Oulton. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 55–56.) For certain functions a 10-term Laguerre-series approximation may be written down in terms of the value of the function at times proportional to the 10 zeros of the Laguerre function of order 10.

MEASUREMENTS AND TEST GEAR

621.317.3 : 551.594.6 **2176**

Atmospheric Radio Noise.—J. Harwood & C. Nicolson. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 183–190.) Equipment incorporating recent modifications is described for the measurement of the characteristics of atmospheric noise, with main emphasis on very low frequencies. Parameters measured include the average voltage of the noise envelope, the fraction of time for which various voltages are exceeded by the envelope, and the number of excursions of the envelope above various voltage levels.

621.317.3 : 621.314.7 **2177**

A Transistor Tester for the Experimental Lab.—R. A. Hempel. (*Electronic Ind.*, Feb. 1958, Vol. 17, No. 2, pp. 58–61.)

621.317.3 : 621.314.7 **2178**

A Method for Measuring Transistor Current Gain at Radio Frequencies.—F. J. Hyde. (*J. sci. Instrum.*, March 1958, Vol. 35, No. 3, pp. 115.) A null method of measuring complex values of α at frequencies up to tens of megacycles, using simple apparatus, is described.

621.317.3: 621.314.7 **2179**

Transistor Cut-Off Frequency Measurement.—L. G. Cripps. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 781–782.) Considerable advantage may be gained by measuring the alpha cut-off frequency by feeding signals to the base with the emitter grounded, since a value nearer to that for the intrinsic cut-off frequency is obtained.

621.317.3 : 621.396.822 : 621.372.54 **2180**

A Tunable Filter for Use in the Measurement of Excess Noise from Local Oscillators.—W. P. N. Court. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 208–209.) "A method of forming a filter from available microwave components is described. This filter is tunable over a wide band and need not be removed physically from the circuit when filter action is not required."

621.317.3.029.5 : 621.396.822 **2181**

Bridge Method of Measuring Noise in Low-Noise Devices at Radio Frequencies.—K. S. Champlin. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 779.) Noise resistances of several hundred ohms can be measured with an over-all error of less than 5%. Errors due to stray coupling and pre-amplifier noise are minimized.

621.317.31 (083.74) **2182**

Redetermination of the Standard Ampere.—(*Tech. News Bull. nat. Bur. Stand.*, Feb. 1958, Vol. 42, No. 2, pp. 21–23.) The National Bureau of Standards maintains permanent standards of voltage and resistance from which it is possible to calculate a standard ampere. A comparison with the absolute ampere has recently been made (*a*) using a Pellat-type electrodynamometer, (*b*) using a current balance. The observed change since 1942 when the last comparison was made is so slight that it may be due to experimental error only.

621.317.316 : 621.372.413.029.64 **2183**

Measuring Frequency of X-Band Standard Cavities.—W. A. Gerard. (*Electronic Ind.*, Feb. 1958, Vol. 17, No. 2, pp. 66–70.) A flexible method of measuring cavity frequency is described and the determination of stability, temperature compensation, etc., is discussed.

621.317.335.3.029.64 **2184**

A Spectrometer Method for Measuring the Electrical Constants of Lossy Materials.—J. S. Seeley. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 18–26.) The analysis of the propagation of a plane wave at oblique incidence through a strip of lossy material is stated, the behaviour being completely represented by the use of a complex refractive index and a complex reflection coefficient. A technique is described for measuring these constants by means of a 1-cm parallel-plate spectrometer.

621.317.44 **2185**

High-Frequency Magnetic Permeability Measurements using Toroidal Coils.—R. D. Harrington & R. C. Powell. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 784.) The results of a previous investigation [3625 of 1954 (Kostyshyn & Haas)] are discussed in relation to the conclusions reached by Schwartz (44 of January).

621.317.729 : 621.3.013.2 **2186**

An Improved Electromagnetic Analogue.—W. T. J. Atkins. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 151–154.) A quantitative representation of symmetrical e.m. fields is possible enabling dynamic and static effects to be studied.

621.317.75 : 621.395.625.3 **2187**

The Magnetic-Tape Oscillograph.—W. Reinert. (*Elektrotech. Z.*, *Edn B*, 21st Dec. 1957, Vol. 9, No. 12, pp. 493–498.) By means of the process of boundary-displacement recording [1963 of 1952 (Daniels)] oscillograms can be produced on magnetic tape and made visible by the application of iron dust. The design and advantages of such a system are discussed.

621.317.755 : 621.318.57 : 621.387 **2188**

Impulse Voltage Wave Chopping Circuit for Use with a Recurrent Surge Oscilloscope.—J. W. Armitage. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 186–188.) A hydrogen-filled thyratron is used at voltages up to 2·5 kV.

621.317.755 : 621-526 **2189**

An Error-Sampled Sweep-Position Control System.—C. H. Knapp, E. Shapiro & R. A. Thorpe. (*IBM J. Res. Developm.*, Jan. 1958, Vol. 2, No. 1, pp. 14–35.) Discusses the design of equipment developed for controlling the position of an instantaneous portion of a c.r.o. trace with sufficient precision to permit accurate measurements.

621.317.76 : 621.395.625 **2190**

'Wow' and 'Flutter'.—R. G. T. Bennett & R. L. Currie. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 162–164.) A method of measurement requiring only a stable oscillator and an oscilloscope.

621.317.77 **2191**

Some Methods of Phase Measurement used in Transfer Function Analysis.—D. J. Collins & J. E. Smith. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 182–186.) The more conventional systems of measurement are summarized and their limitations are indicated.

621.317.77 : 621.396.11.029.64 **2192**

Single Path Phase Measuring System for Three-Centimetre Radio Waves.—Thompson & Vetter. (See 2205.)

621.317.79.087.9 : 538.632 : 537.311.33 **2193**

A Digital Recording System for Measuring the Electrical Properties of Semiconductors.—R. H. A. Carter, D. J. Howarth & D. H. Putley. (*J. sci. Instrum.*, March 1958, Vol. 35, No. 3, pp. 115–116.) A modification of the self-balancing potentiometer system described earlier [see 3155 of 1956 (Putley)], whereby measurements can be recorded in digital form on teleprinter tape suitable for feeding directly into an electronic computer.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.319.2 : 535.215 **2194**

Photoelectrets and the Formation of a Latent Electrophotographic Image.—V. M. Fridkin. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Jan. 1958, Vol. 118, No. 2, pp. 273–276.) A method similar to xerography using a polarized layer of polycrystalline sulphur 50 μ thick is described. A graph shows that the polarization of the electrets after dropping to 40% of its original value remains constant for several hundred hours.

621.384.612 **2195**

Linear Theory of Synchrotron Oscillations: Part 2—Particle Losses during Acceleration and Tolerance Theory.—L. L. Goldin & D. G. Koškarev. (*Nuovo Cim.*, 1st Aug. 1957, Vol. 6, No. 2, pp. 286–298. In English.) Part 1 of the paper is contained in an article on synchrotron oscillations in strong-focusing accelerators noted in 1839 of 1956. See also 2558 of 1957.

621.384.613 **2196**

Space-Charge Phenomena in a Betatron.—H. W. Schmidt. (*Ann. Phys., Lpz.*, 10th May 1957, Vol. 19, Nos. 6–8, pp. 298–303.)

621.385.833 **2197**

The Use of the Electron-Microscope Shadow Method for Studying the Potential Distribution in *p-n* Junctions.—V. N. Vertsner & L. N. Malakhov. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Jan. 1958, Vol. 118, No. 2, pp. 266–268.) A resolution of the order of 0·1 μ is obtained.

621.398 : 629.19 **2198**

Earth - Satellite Telemetry Coding System.—R. W. Rochelle. (*Elect. Engng, N.Y.*, Dec. 1957, Vol. 76, No. 12, pp. 1062–1065.) Square - hysteresis - loop magnetic materials are used with transistors in magnetically coupled multivibrators. The encoder operates on a time sharing principle and controls a.f. multivibrators. The length of time these are on, the frequency of oscillation and the length of time between oscillation periods give the required channels of information with no dead time in the transmitted signal. See also 1844 of June (Matthews).

PROPAGATION OF WAVES

621.396.11 **2199**

On the Propagation of the Electromagnetic Waves in an Inhomogeneous Atmosphere: Part 2.—Y. Nomura & K. Takaku. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Sept. 1956, Vol. 8, No. 2, pp. 61–96.) The integral representation for e.m. wave propagation (see Part 1: 250 of 1957) is transformed into the sum of residues at poles, and the intensity of diffracted waves in the geometric-optical shadow is calculated. The inhomogeneous atmosphere causes a marked change in the intensity of the diffracted waves as compared with the homogeneous atmosphere. Expressions are derived for diffracted waves at the boundaries where the medium changes continuously, as well as at the ordinary boundaries where the medium changes discontinuously. For a preliminary report of the work, see 545 of 1956.

621.396.11 **2200**

June—Vertical and Oblique Incidences.—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, Sept. 1957, Vol. 11, No. 49, pp. 2–13. In Spanish & English.) A further series of comparisons between I.F.R.B. forecasts and those based on the Gea method. See also 1241 of April.

621.396.11 : 523.164.3 **2201**

Propagation of Radio Waves from Cosmical Sources.—Chvojková. (See 2083.)

621.396.11 : 551.510.535 **2202**

Single-Hop Propagation of Radio Waves to a Distance of 5 300 km.—E. Warren & E. L. Hagg. (*Nature, Lond.*, 4th Jan. 1958, Vol. 181, No. 4601, pp. 34–35.) Single-hop propagation between Ottawa and Slough associated with reflection at apparent heights of 600–900 km, was observed e.g. at 1800 G.M.T. on 11th September 1957 for frequencies of 27–33 Mc/s. Records indicate that the one-hop mode occurs with the same reliability as the two- and three-hop modes at times between 1200 and 2020 G.M.T. for similar distances and frequencies.

621.396.11.029.45 : 551.510.53 **2203**

Observations of Magneto-ionic Duct Propagation using Man-Made Signals of Very Low Frequency.—R. A. Helliwell & E. Gehrels. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 785–787.) Echoes of $\frac{1}{4}$-sec pulse signals from station NSS in Annapolis, Maryland, on 15·5 kc/s have been detected with delays up to nearly 1 sec at Cape Horn, South America, providing a controlled test of the Eckersley-Storey theory of 'whistlers' (see e.g. 142 of 1954). When observed, the echoes were 10–30 dB weaker than the direct wave, and were subject to systematic fading and splitting.

621.396.11.029.62/.63 **2204**

Guglielmo Marconi and Communication beyond the Horizon.—G. A. Isted. (*Point to Point Telecommun.*, Feb. 1958, Vol. 2, No. 2, pp. 5–17.) Marconi's work on propagation at 30 and 600 Mc/s is reviewed. He showed conclusively that communication could be achieved beyond optical path limits using reflection in the lower atmosphere. His apparatus is described which, although primitive, seemed very effective.

621.396.11.029.64 : 621.317.77 **2205**

Single Path Phase Measuring System for Three-Centimetre Radio Waves.—M. C. Thompson, Jr, & M. J. Vetter. (*Rev. sci. Instrum.*, Feb. 1958, Vol. 29, No. 2, pp. 148–150.) A practical system is described for making single-path phase measurements at X-band frequencies over ranges up to about 20 miles with instrumental noise of the order of a fraction of one degree. The techniques for obtaining the necessary transmitter frequency stability of $1:10^9$ are described. The use of such techniques to form a simple microwave repeater system with power gain of the order of 50 dB is discussed.

621.396.8 **2206**

Asymmetry in Long-Distance W/T Circuits.—A. M. Humby. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 204–207.) An analysis has been made of hourly records of reception conditions on the Montreal–Melbourne circuit for the period 1935–1955, for which both short and long paths are mainly over sea. A seasonal asymmetry is noticed, resulting in a significant fall in performance in the direction which entails reception in local summer. No sharply defined hour of onset of this asymmetry is evident, as was reported for transequatorial land routes [see 3208 of 1956 (Humby & Minnis)].

RECEPTION

621.376.2/.3 : 621.372.5 **2207**

The Change of Amplitude Modulation into Frequency Modulation Caused by some Limiter Circuits.—E. De Castro & E. Stanghellini. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 530–544.) Simple formulae for calculating output phase deviation corresponding to a given input modulation depth are derived. See also 2879 of 1956 (De Castro).

621.376.23 : 621.372.54 **2208**

Detection of Pulsed Signals in Noise.—H. S. Heaps & A. T. Isaacs. (*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 190–193.) An analysis of the optimum design of a low-pass filter for the detection of rectangular pulses in a background of white noise. Results are shown graphically.

621.396.62 : 621.376.3 : 621.314.7 **2209**

A Transistorized 150-Mc/s F.M. Receiver.—W. J. Giguere. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 693–699.) A completely transistorized experimental f.m. communication receiver is described. Diffused-base transistors are used and its frequency range is 152–174 Mc/s. Power consumption is 130 mW and adjacent-channel selectivity is better than 70 dB for 60-kc/s channel spacing.

621.396.621.54 **2210**

Variable-Selectivity I.F. Stages in Communication Receivers.—J. B. Dance. (*Short Wave Mag.*, Jan. 1958, Vol. 15, No. 11, pp. 572–578.) Design considerations for the construction of an i.f. unit of variable selectivity for use with an existing receiver.

621.396.662 **2211**

Sensitive Tuning Indicators.—R. Oliver. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 235–236.) A note on circuit arrangements for improving the performance of a 'magic-eye' tuning indicator, particularly by connecting it to an appropriate screen-grid line or, in the case of f.m. receivers, in the limiter anode circuit.

621.396.82 **2212**

Radio Interfereuce: Part 1—Introduction.—D. A. Thorn. (*P.O. elect. Engrs' J.*, Jan. 1958, Vol. 50, Part 4, pp. 226–227.) An introduction to a series of six articles which will present a general survey of the subject with emphasis on the telecommunications aspects.

621.396.82 **2213**

Radio Interference: Part 2—The Post Office Radio Interference System.—G. A. C. R. Britton. (*P.O. elect. Engrs' J.*, Jan. 1958, Vol. 50, Part 4, pp. 227–230.) The organization for dealing with interference complaints, the apparatus and methods used for tracing the interference, are described. The main types of apparatus that cause the interference are listed.

621.396.82 : 621.376.3 **2214**

Theory of Stronger-Signal Capture in F.M. Reception.—E. J. Baghdady. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 728–738.) The amplitude limiter in f.m. receivers spreads out the spectrum of the disturbing signals over a range which is often much greater than that of the message modulation. Hence limiting followed by filtering can suppress disturbances considerably ahead of the discriminator stage.

STATIONS AND COMMUNICATION SYSTEMS

621.391 **2215**

On Nonuniform Sampling of Bandwidth-Limited Signals.—J. L. Yen. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 251–257. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.)

621.391 **2216**

Definition of 'Bandwidth' and 'Time Duration' of Signals which are Connected by an Identity.—D. G. Lampard. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 286–288. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.)

621.391 : 519.2 **2217**

Systems Analysis of Discrete Markov Processes.—R. W. Sittler. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 257–266. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.) See also 1227 of 1957 (Huggins).

621.391 : 519.2 **2218**

Limiting Conditions on the Correlation Properties of Random Signals.—G. Kraus & H. Pötzl. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 282–285. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.)

621.391 : 621.376.2 **2219**

The Theory of Analytic Band-Limited Signals Applied to Carrier Systems.—J. R. V. Oswald. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 244–257. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.)

621.391 : 621.376.5 **2220**

A Study of Rough Amplitude Quantization by means of Nyquist Sampling Theory.—B. Widrow. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. CT-3, No. 4, pp. 266–276. Abstract, *Proc. Inst. Radio Engrs*, May 1957, Vol. 45, No. 5, Part 1, p. 717.)

621.391 : 621.396.822 **2221**

The Penetration of Noise into Communication Channels in the Light of Information Theory.—P. Neidhardt. (*NachrTech.*, Sept. 1957, Vol. 7, No. 9, pp. 419–422.) The reduction of information content by noise is discussed in terms of communication theory particularly as derived by Shannon.

621.396.1 **2222**

On Channel Spacing.—(*Point to Point Telecommun.*, Oct. 1957, Vol. 2, No. 1, pp. 19–23.) Channel separations of 50 kc/s are called for in the 80- and 160-Mc/s bands in the United Kingdom. Some of the problems of this and possible smaller separations are discussed briefly.

621.396.2.029.6 **2223**

Standardizing Microwave Communication Systems.—T. Clark. (*Electronic Ind.*, Feb. 1958, Vol. 17, No. 2, pp.

50–54.) Suggestions by a committee of the Electronic Industries Association for standardization and definition of terms for microwave communication systems.

621.396.3 **2224**

Automatic Error Correction on H.F. Telegraph Circuits.—P. R. Keller. (*Point to Point Telecommun.*, Feb. 1958, Vol. 2, No. 2, pp. 25–48.) A seven-character three-mark code based on the Van Duuren system recommended by the C.C.I.T. is described with respect to error correction. The techniques involved and the basic block diagrams for circuitry are given. Emphasis is placed on the need for fully electronic equipment and an appendix is given on cold-cathode valve circuits.

621.396.3 : 621.376.4 **2225**

Kineplex, a Bandwidth-Efficient Binary Transmission System.—R. R. Mosier & R. G. Clabaugh. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 723–727. Discussion, pp. 727–728.) Description of a high-capacity phase-shift data transmission system which will operate in a single telephone voice band of standard quality or equivalent, and will accept at its input, with suitable conversion, binary data from any source.

621.396.41 : 621.376.32 **2226**

F.M. Exciter for Sight or Scatter Systems.—A. E. Anderson & H. D. Hern. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 148–151.) The exciter accepts multichannel output of the telephone terminal equipment as a modulating signal and produces an output power of 15 W from 700 to 1 200 Mc/s, and 8 W from 1 700 to 2 400 Mc/s. The unit handles 132 voice channels in addition to an order-wire system.

621.396.41 : 621.396.65 **2227**

6 000-Mc/s Radio Relay System for Broad-Band Long-Haul Service in the Bell System.—M. B. McDavitt. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 715–722.) The system is designed to transmit telephone messages or colour television signals over distances up to 4 000 miles. Objectives and requisites of the system and its principal features are described. Details are given of some of the more interesting new components.

621.396.5 : 621.396.65 **2228**

An Introduction to 'Rural Radio.'—C. B. Wooster. (*Point to Point Telecommun.*, Oct. 1957, Vol. 2, No. 1, pp. 24–48.) Techniques are discussed which permit the full integration of single-channel v.h.f. links with line telephone systems in areas where it is impracticable to erect landlines. The equipment is fairly simple and not very difficult to maintain. A wide range of signalling facilities is available. See 2294 of 1957.

621.396.65 **2229**

Convention on Radio Links.—(*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 321–576.) First of a number of issues covering the proceedings of a convention held in Rome 5th–8th June 1957. Abstracts of some of the papers are given individually; titles of others are as follows:

(*a*) The Trends of Telecommunications as Affected by Solid-State Electronic Instrumentation.—M. J. Kelly (pp. 333–345, in English. Discussion, pp. 346–348).

(*b*) Propagation with Horizontal and Vertical Polarization over the Paths Portofino–Monte Beigua and Portofino–Monte S. Nicolao.—P. Quarta (pp. 404–414).

(*c*) Tests of Radio Propagation beyond the Horizon over the Mediterranean.—L. Niccolai (pp. 415–418).

(*d*) Radio Link Italy–Spain for Tropospheric Propagation beyond the Horizon.—L. Pallavicino (pp. 419–427).

(*e*) 6 000-Mc/s Radio Relay System for Broad-Band Long-Haul Service in the Bell System.—M. B. McDavitt (pp. 428–446, in English). See 2227 above.

(*f*) Some Problems Experienced in the Establishment of Radio Link Systems.—A. W. Montgomery (pp. 447–462, in English).

(*g*) The Planning of Tropospheric Scatter Systems.—K. W. Pearson (pp. 463–467, in English).

(*h*) Kineplex.—P. J. Icenbice & S. R. Crawford (pp. 468–473, in English). See 2225 above.

(*i*) Calculation of the Intermodulation Noise in F.M. Radio Links for Telephony.—B. Peroni (pp. 502–529).

(*j*) Possibilities and Limitations in the Use of Radio Links comprising Demodulation-Type Repeaters.—R. B. Chaux (pp. 545–549, in French).

621.396.65 **2230**

Radio Link Milan–Palermo for Sound and Telephony.—B. Peroni. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 371–388.) Detailed description of the relay system for transmitting the television sound broadcasts of the R.A.I. Four two-way high-quality sound channels and 12 speech channels are available, and particular attention has been given to the reduction of cross-modulation and noise. For details of the television link, see 2254 below.

621.396.65 : 621.396.43 : 621.396.8 **2231**

The Calculation of Distortion Factors and Noise due to various Types of Distortion in Multichannel F.M. Radio Links.—H. Marko. (*Nachrichtentech. Z.*, Sept. 1957, Vol. 10, No. 9, pp. 450–457.) The discussion is based on the C.C.I.R recommendations for radio-link systems. See also 2232 below.

621.396.65 : 621.396.43 : 621.396.822 **2232**

The Distribution of Noise in Radio Links for Multichannel Telephony Transmission.—W. Mansfeld. (*Nachr-Tech.*, Aug. & Oct. 1957, Vol. 7, Nos. 8 & 10, pp. 329–335 & 466–472.) The characteristics are outlined of a hypothetical 2 500-km radio link intended as reference system and based on C.C.I.R. specifications. The various types of noise, their origin and elimination are discussed.

621.396.65 : 621.396.822 **2233**

Thermal Noise in Multisection Radio Links.—B. B. Jacobsen. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 139–150.) An investigation of methods for calculating thermal noise in a long radio link when the individual sections are subject to fading. A technique is described for combining the fading statistics from individual sections to obtain the first three moments for the over-all circuit. This is expressed in terms of an 'augmented log-normal distribution', which can readily be translated into the distribution of thermal noise in the complete path. A method is described for synthesizing an over-all circuit distribution from a number of sections with similar fading statistics.

621.396.931 : 621.395.3 **2234**

The Probability of Success of a Call over the Swiss Network for Calling Motor Vehicles.—E. Wey. (*Tech. Mitt. PTT*, 1st Sept. 1957, Vol. 35, No. 9, pp. 387–395.) Description of the system (see 1159 of 1955) and report on the efficiency of the service based on country-wide field-strength measurements.

SUBSIDIARY APPARATUS

621-52 **2235**

A Note on the Evaluation of the Response of a Nonlinear Element to Sinusoidal and Random Signals.—J. L. Douce. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 88–92.) A method is developed based upon a mathematical technique described by Lewis (*Trans. Amer. Inst. elect. Engrs*, 1954, Vol. 73, Part I, pp. 693–700), which is particularly useful for the analysis of nonlinear control systems. Simple cursors may be constructed to determine the effective gain of the nonlinear unit. The application to hysteretic nonlinearities and the response to random signals is illustrated.

621-52 : 681.142 **2236**

Statistical Design Theory for Strictly Digital Sampled-Data Systems.—S. S. L. Chang. (*Commun. & Electronics*, Jan. 1958, No. 34, pp. 702–708. Discussion, pp. 708–709.) Present design theories are based on system response to inputs of known shape and little attention is given to situations in which only the statistical properties of the inputs are known. The basis of a statistical design theory is presented and applied to the optimum design of a strictly digital system. The new theory forms a close parallel to Wiener's theory of optimum filtering and prediction, and its various manifestations in applications to continuous feedback control systems.

621-526 **2237**

A Correlation between the Transient and Frequency Responses in Servomechanisms.—Z. J. Jelonek & G. I. Boomer. (*J. Brit. Instn Radio Engrs*, Feb. 1958, Vol. 18, No. 2, pp. 101–114.) Statistical evidence is used to relate time- and fequency-response parameters. These empirical relations, together with a new theoretical relation for the delay time, enable the step response to be sketched, given the open-loop response. Some non-typical closed-loop responses may be broken down into two subtransient step responses which are separately determinable.

621-526 : 519.2 **2238**

Fundamental Equations for the Application of Statistical Techniques to Feedback-Control Systems.—G. A. Biernson. (*Trans. Inst. Radio Engrs*, Feb. 1957, No. PGAC-2, pp. 56–78. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 572.)

621-526 : 621.375.4 **2239**

Direct-Drive Amplifier for Two-Speed Servos.—B. E. Orr. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 146–147.) A five-transistor servo amplifier drives a standard motor without using an output transformer. For a two-speed system the amplifier contains a switching circuit incorporating Zener diodes, and a three-stage feedback network.

621.311.68 **2240**

Automatic Power Plant for Telecommunications Installations.—K. J. Bladon. (*Point to Point Telecommun.*, Oct. 1957, Vol. 2, No. 1, pp. 5–18.) The development of multichannel radio-telephone services capable of unattended operation for long periods has led to the need for automatic generators of electric power. Some of the factors in the design of Diesel power plants are considered. Both mains-standby and dual alternator systems are discussed.

621.314.22 **2241**

Hot Transformers.—(*Engineering, Lond.*, 15th Nov. 1957, Vol. 184, No. 4784, p. 623.) Brief note on two experimental types of transformer designed for operation at 500°C. One, housed in an evacuated stainless-steel case, is rated 100 VA at 1 600 c/s, the other is of open construction with a nominal rating of 50 VA at 50 c/s.

621.314.58 : 621.373.431.1 **2242**

Magnetic Inverter uses Tubes or Transistors.—C. H. R. Campling. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 158–161.) A means of producing a.c. from d.c. The collector and emitter coil windings of the 'on' transistor of a multivibrator are differentially connected across the input voltage so that the drive winding partially determines the frequency of oscillation.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 : 535.623 **2243**

The Analogous Frequency of the Colour Subcarrier in an Adaptation of the N.T.S.C. System to the 625-Line C.C.I.R. Standard.—E. Schwartz. (*Rundfunktech. Mitt.*, Oct. 1957, Vol. 1, No. 5, pp. 191–195.) The various proposals for a suitable colour subcarrier frequency are discussed. Calculation leads to a value of $4 \cdot 17 \pm 0 \cdot 04$ Mc/s, so that 4·210 937 5 or 4·101 562 5 Mc/s could be chosen. A symmetrical-sideband system can be adopted if a low subcarrier frequency is used. See also 1878 of 1956 (Below & Schwartz).

621.397.5 : 621.395.625.3 **2244**

Television Tape Recorder.—(*Electronic Radio Engr*, May 1958, Vol. 35, No. 5, pp. 193–195.) In the 'vision electronic recording apparatus' of the B.B.C., a 20½-in.-diameter reel of half-inch magnetic tape played at 200 in./sec provides 15 minutes of program. Three tracks are used simultaneously, two for the video signal and one for sound. The modulation system used, and some electrical and mechanical details, are described.

621.397.5 : 621.395.625.3 **2245**

Television Recording on Magnetic Tape by the Ampex Method.—H. J. V. Braunmühl & O. Schmidbauer. (*Rundfunktech. Mitt.*, Oct. 1957, Vol. 1, No. 5, pp. 186–190.) See also 294 of January (Ginsburg).

621.397.6 : 621.314.7 **2246**

The Application of Transistors to Television.—J. N. Barry & A. E. Jackets. (*J. Telev. Soc.*, Oct.–Dec. 1957, Vol. 8, No. 8, pp. 318–334.) After a review of recent progress towards increasing the frequency and power range of transistors, their application to each of the main television circuits is discussed in detail with circuit diagrams. A short note on their suitability in colour television apparatus is included, and possible future applications are discussed.

621.397.61.029.63 **2247**

B.B.C. Band-V Experimental Television Transmission.—(*Engineer, Lond.*, 29th Nov. 1957, Vol. 204, No. 5314, pp. 797–798.) Details of the transmitter equipment installed at Crystal Palace for tests in the range 470–960 Mc/s, but particularly at a vision frequency of 654·25 Mc/s. An initial series of tests on 405 lines will be followed by 625-line transmissions.

621.397.62 : 535.623 : 621.3.049.75 **2248**

Etched I.F. Amplifier Pares Colour TV Cost.—L. Ruth. (*Electronics*, 14th March 1958, Vol. 31, No. 11, pp. 135–137.) Vane-tuned inductances and rejection traps are etched on the same board as the wiring of a 41-Mc/s i.f. strip. Design requirements construction details and performance data are given.

621.397.62 : 621.314.7 **2249**

Transistor Television Circuits: Part 2—Scan Output Stages, Video and Signal-Frequency Amplifiers.—J. N. Barry & G. W. Secker. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 231–235.) Circuit details are given of a complete frame output section for a 17-in. receiver, and the operation of transistors in linear r.f. amplifiers is discussed. Part 1: 1889 of June.

621.397.62.029.63 : 621.372.632 **2250**

Television Reception on Band V.—H. N. Gant. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 244–246.) A converter for use with band-III receivers is described, which incorporates an u.h.f. amplifier stage. A new type of valve with special electrode construction and low anode current is used and a noise factor of about 11 dB for the converter is achieved.

621.397.621.2.002.1 : 535.623 **2251**

Development Problems of Colour-Television Picture Tubes.—I. Bornemann. (*NachrTech.*, Sept. 1957, Vol. 7, No. 9, pp. 399–404.) The differences in afterglow duration of phosphors for tricolour c.r. tubes and their effect on colour fringing are considered. See also 2930 of 1957.

621.397.7 **2252**

The New B.B.C. Television Studios in London.—A. N. Thomas. (*Rundfunktech. Mitt.*, Oct. 1957, Vol. 1, No. 5, pp. 169–180.) Illustrated descriptions are given of the lay-out and the equipment of the Riverside studios [see 609 of February (Nickels & Grubb)], the television theatre at Shepherds Bush in operation since July 1957, and the White City television centre under construction.

621.397.7 **2253**

The New Television Studios of the R.A.I. in Rome.—L. Sponzilli. (*Rundfunktech. Mitt.*, Oct. 1957, Vol. 1, No. 5, pp. 181–185.)

621.397.7 **2254**

Television Radio Link Milan–Palermo.—F. Carassa. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 349–370.) The wide-band relay link described is 1 640 km long and comprises 21 repeater and terminal-stations.

621.397.7 **2255**

The Microwave Television Link Italy–Sardinia.—F. Carassa. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 389–403.) The system described consists of an oversea link of 238 km length. As a result of propagation tests (1890 of 1957) a carrier frequency of 940 Mc/s, that of the main television network, was chosen; this provides a large bandwidth and facilitates connection to the relay system on the mainland (2254 above).

621.397.7 : 621.3.018.78 **2256**

Waveform Distortion in F.M. Television Radio Links.—F. Carassa. (*Alta Frequenza*, Oct. 1957, Vol. 26, No. 5, pp. 474–501.) The waveform distortion due to small sinusoidal deviations from the ideal flat amplitude and group-delay response curves is interpreted in terms of echoes. The effect of various response conditions on the C.C.I.R. test waveform is investigated. Distortion can be reduced considerably, by means of a video-frequency low-pass filter following the demodulator.

621.397.8 **2257**

The Influence of Multipath Propagation on the Carrier-Frequency Spectrum of the Television Signal.—K. Bernath & H. Brand. (*Tech. Mitt. PTT*, 1st Oct. 1957, Vol. 35, No. 10, pp. 401–412.) The case of ideal two-path reception is analysed and the principal factors controlling the shape of the received spectrum are discussed. Experimental results are compared with computed spectra of received signals for two transmitters and various paths, including a case of almost ideal reflection from an intervening lake, and reception via a long mountainous path. The difficulties likely to arise in the reception of colour television signals are examined in the light of statistical investigations of reception in built-up areas.

621.397.8 : 535.623 **2258**

The Definition of Picture Quality in Monochrome and Colour Television.—P. Neidhardt. (*NachrTech.*, Sept. 1957, Vol. 7, No. 9, pp. 389–398.) Detailed discussion of the means of assessing picture quality in clearly definable terms. By analogy with information theory the concept of detail entropy is derived and its application extended to the assessment of colour image quality. 29 references.

TRANSMISSION

621.396.61 : 621.314.7 **2259**

Transistor Transmitter.—L. F. Shaw. (*Wireless World*, May 1958, Vol. 64, No. 5, pp. 241–243.) A portable all-transistor amateur transmitter for the 160-m band is described.

VALVES AND THERMIONICS

621.314.63 : 546.28 **2260**

Determination of the Diffusion Length L and the Inversion Density n_i from the Forward Characteristics of Silicon Alloy-Junction-Type Rectifiers.—A. Herlet. (*Z. angew. Phys.*, April 1957, Vol. 9, No. 4, pp. 155–158.) A method of determining L and n_i is derived on the basis of a comparison of experimentally obtained characteristics with theory. See also 598 of 1956.

621.314.63.029.65 : 546.28 **2261**

Wafer-Type Rectifiers for Millimetre Waves.—W. M. Sharpless. (*Bell Lab. Rec.*, Jan. 1958, Vol. 36, No. 1, pp. 21–24.) See 1607 of 1957.

621.314.632.029.64 : 546.681.19 **2262**

A Gallium Arsenide Microwave Diode.—D. A. Jenny. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 717–722.) The semiconductor properties of GaAs, particularly the high electron mobility and forbidden band gap, besides favourable point-contact rectification characteristics, are of interest for microwave diode applications. Theoretically and experimentally GaAs is potentially superior to Ge and Si in point-contact diodes. Its mixer conversion loss is less, and it should operate at higher temperatures.

621.314.7 **2263**

Developments in Transistor Electronics.—L. B. Valdes. (*J. Electronics Control*, Jan. 1958, Vol. 4, No. 1, pp. 1–16.) The design theory of point-contact and junction transistors is reviewed. Selected examples illustrate the small-signal and large-signal properties of junction transistors. These phenomena are related to the theory of *p-n* junctions.

621.314.7 **2264**

On Carrier Accumulation, and the Properties of Certain Semiconductor Junctions.—J. B. Gunn. (*J. Electronics Control*, Jan 1958, Vol. 4, No. 1, pp. 17–50.) In a semiconductor near an electrode boundary a situation may arise in which minority carriers are drawn by the electric field towards the boundary, across which they can pass only with difficulty. Carrier accumulation will then develop. It is shown theoretically that where a junction exists between relatively pure and relatively impure regions of a semiconductor, carrier accumulation should be observable in the lightly doped section. Such a junction is used in a class of devices known as conductance transistors, which can have a large current gain. This gain is calculated in a particular case, together with emitter and collector impedances.

621.314.7 **2265**

Effects of Low Temperatures on Transistor Characteristics.—A. B. Credle. (*IBM J. Res. Develpm.*, Jan. 1958, Vol. 2, No. 1, pp. 54–71.) Describes measurements on a group of similar *p-n-p* alloy-junction transistors made between 0·5 and 5 Mc/s and from room temperature down to that of liquid N_2, leading to simple equivalent networks. Grounded-emitter pulse tests show that, for a given output pulse current, rise time decreases with temperature, and at the temperature of liquid N_2 rise time decreases as the output current pulse increases. A 30-mW, 3-Mc/s transistor can then deliver a 1-A current pulse with a rise time of a few tenths of a microsecond, with a grounded-emitter gain of about 20. Theoretical expressions for the effect of temperature on transistor parameters are evaluated. See also 4041 of 1957 (Uda).

621.314.7 : 621.317.3 **2266**

A Transistor Tester for the Experimental Lab.—Hempel. (See 2177.)

621.314.7 : 621.317.3 **2267**

A Method of Measuring Transistor Current Gain at Radio Frequencies.—Hyde. (See 2178.)

621.314.7 : 621.317.3 **2268**

Transistor Cut-Off Frequency Measurement.—Cripps. (See 2179.)

621.314.7 : 621.318.57 **2269**

Controlled Saturation in Transistors and its Application in Trigger Circuit Design.—Moody. (See 1995.)

621.314.7 : 621.397.6 **2270**

The Application of Transistors to Television.—Barry & Jackets. (See 2246.)

621.383.032.217.2 **2271**

Interference Photocathodes of Increased Yield with Freely Variable Maximum Spectral Response.—K. Deutscher. (*Naturwissenschaften*, Sept. 1957, Vol. 44, No. 18, pp. 486–487.) A layer of $SbCs_3$ covers a wedge of MgF_2 mounted on an Al mirror. By investigating this interference system under monochromatic light the relation of photocurrent to the reflecting power of the system was determined. The photoelectric yield is a maximum where the reflection is reduced to a minimum.

621.385.029.6 **2272**

Stability of an Electron Beam on a Slalom Orbit.—J. S. Cook, W. H. Louisell & W. H. Yocom. (*J. appl. Phys.*, March 1958, Vol. 29, No. 3, pp. 583–587.) A study of an electrostatic method of focusing an electron beam in a travelling-wave valve in which the beam traverses a slalom-like orbit between the wires of the slow-wave structure. The equations of motion are solved to find initial conditions for stable orbits. Good agreement is found between the computed and measured current transmitted.

621.385.029.64 : 537.533 : 621.375.9 **2273**

Parametric Amplification of Space-Charge Waves.—W. H. Louisell & C. F. Quate. (*Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 707–716.) The principle of using a variable reactance to obtain gain can be applied to an electron beam by modulating it at a frequency of twice the signal frequency. The normal space-charge wave breaks into two parts—one growing and one attenuated; either the 'slow' or 'fast' wave can be amplified. The beam can then be coupled to a slow-wave circuit, in which the 'fast' wave can be amplified again. Noise theorems for previous microwave amplifiers deal with the 'slow' wave and do not apply here. Lower values for the noise figure should in principle be achieved. See also 1606 of May (Bridges).

621.385.032.213.13 **2274**

The Conductivity of Oxide Cathodes.—G. H. Metson. (*Proc. Instn elect. Engrs*, Part C, March 1958, Vol. 105, No. 7, pp. 183–188 & 189–195.)

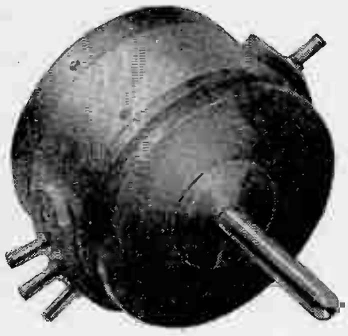
Part 3—Movements of Electrolytic Oxygen in a Conventional Diode System.—Negative oxygen ions from the cathode, if not completely absorbed at the anode, may return to the cathode and produce reactions analogous to the electrolytic actions of the S-type assembly.

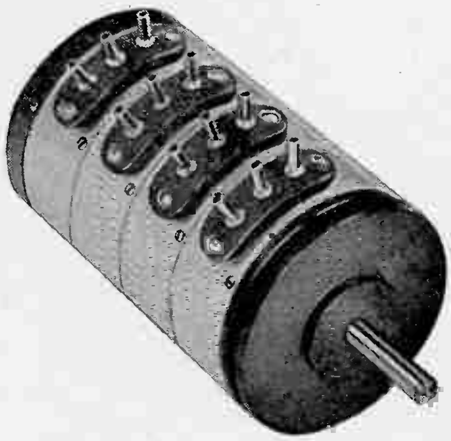
Part 4—Electron Transfer Mechanisms.—At temperatures between 850 and 1 100°K all the electrons are thermionically emitted at the cathode-core surface and pass through the vacuum interstices of the oxide-cathode matrix; their energy is dissipated uniformly by successive non-elastic collisions with solid oxide particles. Between 290 and 600°K, the mechanism is one of solid-state semiconduction and no thermionic emission takes place. These deductions from experimental results agree with the hypothesis of Loosjes and Vink (3208 of 1950).

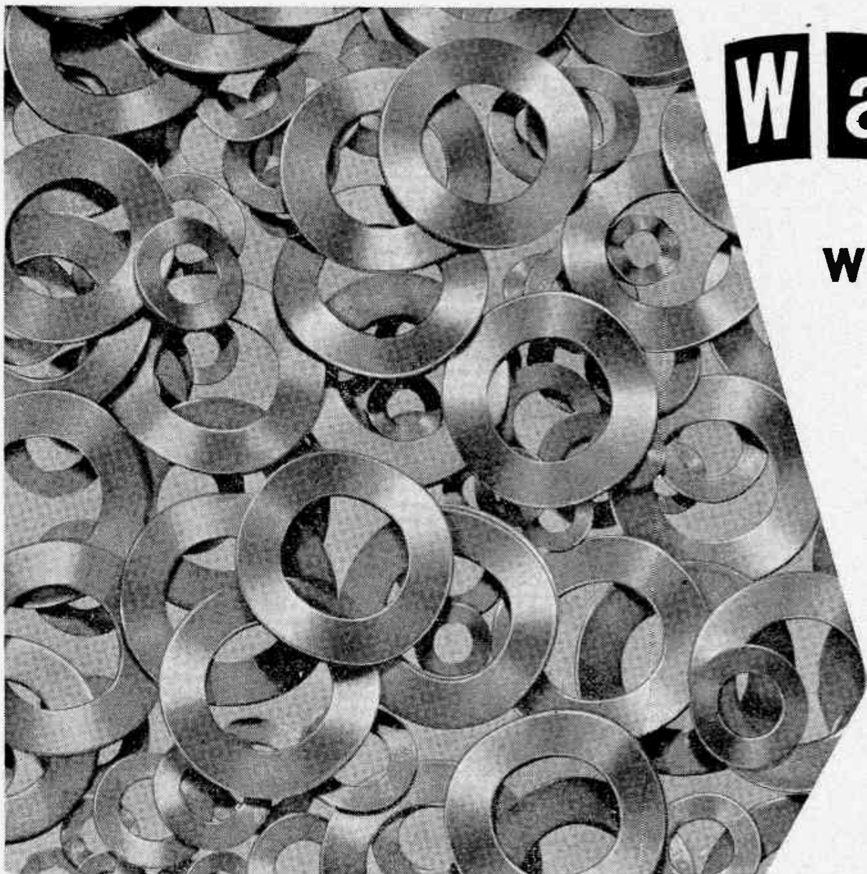
Parts 1 & 2: 4072 & 4073 of 1957. Please note change of U.D.C. number.

621.387.002.2 : 621.318.57 **2275**

A Low-Cost Cold-Cathode Trigger Tube.—A. Turner. (*Electronic Engng*, April 1958, Vol. 30, No. 362, pp. 166–169.) A design for economic production is described.







ELECTRONIC & RADIO ENGINEER

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THE Division of Radiophysics, which is situated within the grounds of Sydney University, has vacancies on its scientific staff for a Research Officer and an Experimental Officer for interesting work on the development of radio aids to aerial navigation.

APPOINTMENT Number 780/184. Applications are invited for appointment as a Research Officer to devise, design and develop special aerials and associated electronic equipment forming part of an overall system to provide D.M.E.-equipped aircraft with an indication of their heading. Applicants should possess a University Honours degree in Engineering or Science, or equivalent qualifications, together with a sound knowledge of modern electronic and u.h.f. techniques, and should be able to show evidence of some years of post-graduate research experience. Dependent upon qualifications and experience, commencing salary will be determined within the salary range of £A1,326–£A1,951 p.a. Research Officer. Salary for a female appointee would be £A170 p.a. less than the corresponding male rates.

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THE appointments will, in the first instance, be for a period of two years, but with the possibility of extension on a more permanent basis at the end of that time and they will be conditional on a satisfactory medical examination.

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APPLICATIONS quoting Appointment No. 780/184 or 780/185 and stating full name, place, date and year of birth, nationality, marital state, present employment, details of qualifications and experience and of war service if any, together with the names of not more than four persons acquainted with the applicant's academic and professional standing, should reach the undersigned, from whom further particulars may be obtained, not later than 18th July, 1958.

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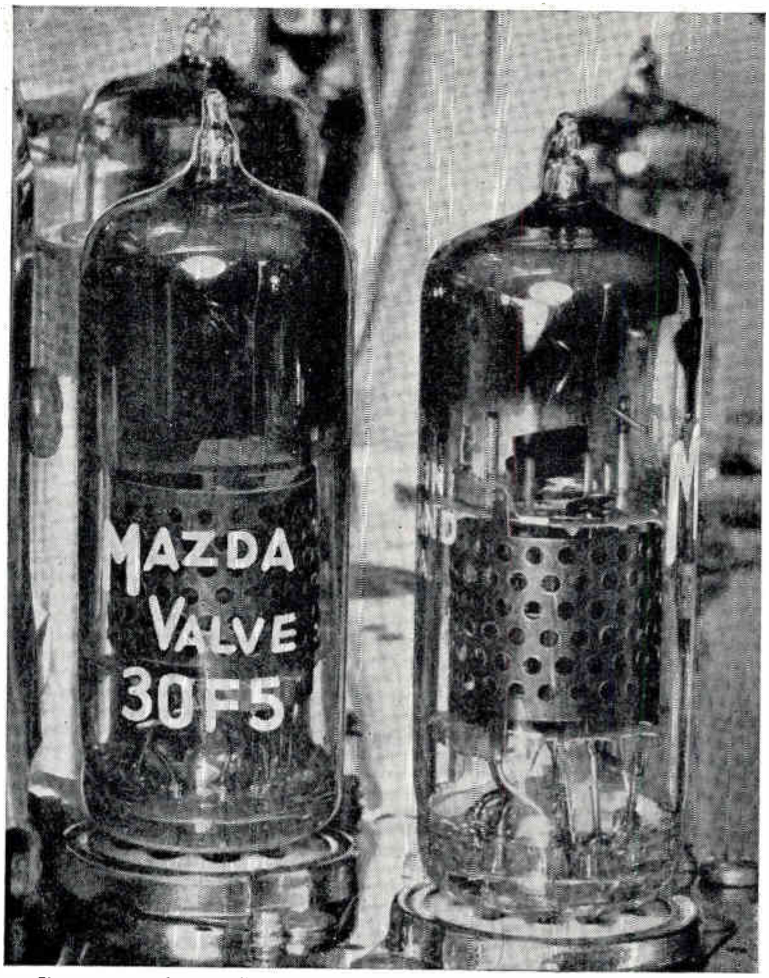

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Heater current (amps)	0.3
Maximum anode watts	3.0
Maximum screen watts	1.0

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Grid bias volts	1.9
Anode current (mA)	10
Screen current (mA)	2.6
Mutal conductance (mA/V)	8.8
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Equivalent grid noise resistance (ohms)	750
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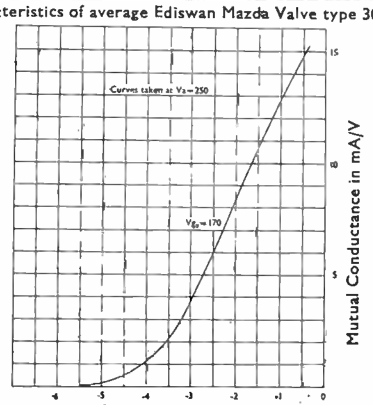
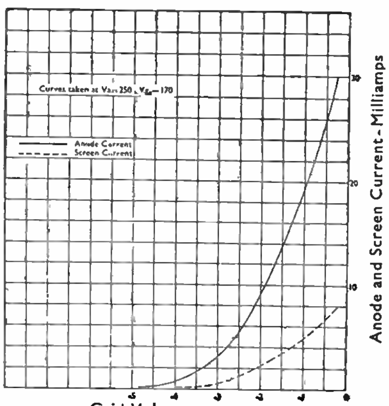
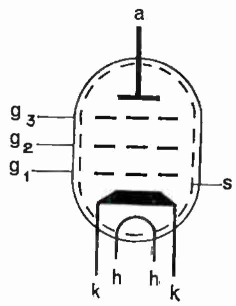
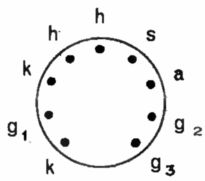
Input Loss at 38 Mc/s (ohms) (with cathodes strapped)	8,800

Characteristics of average Ediswan Mazda Valve type 30F5









B9A Base

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a to E=4.4 pF
g_1 to a=0.0073 pF

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