ELECTRONIC & RADIO ENGINEER

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In this issue

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JUNE 1958 Vol 35 *new series* No 6

GENERAL RADIO COMPANY

Trade Mark

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U.H.F. Admittance Meter

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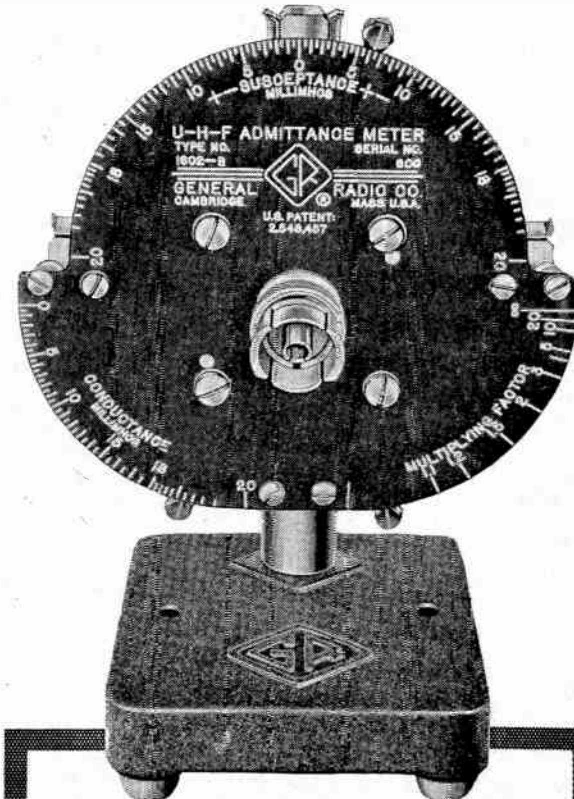

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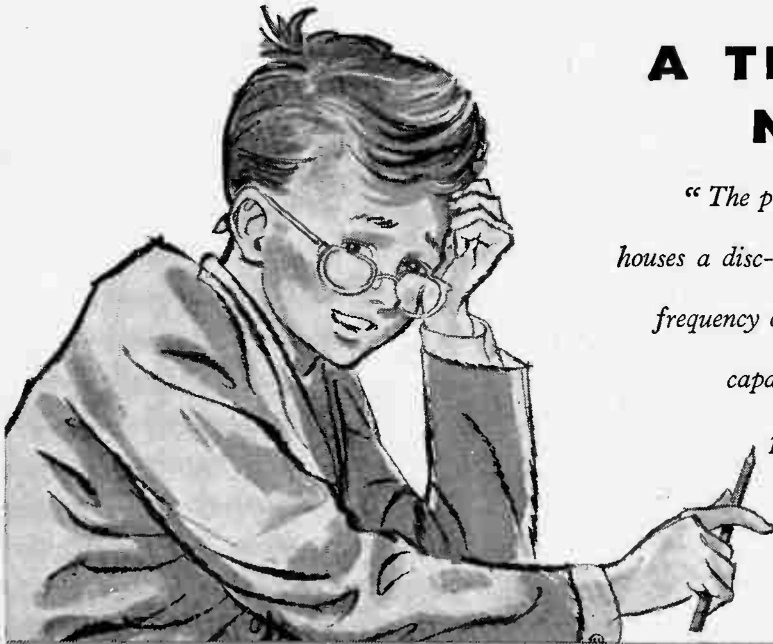

WHERE *does* our young friend get all his information about Marconi instruments? He certainly knows all about our new TF 1041A. He knows that, in addition to its unequalled performance in a.c. measurements, it measures balanced or unbalanced d.c. voltages and a wide range of resistance values. He knows that it has a large mirror-scale meter for fast, precise reading. He knows that there are three optional accessories for extending the a.c. and d.c. ranges and for making measurements on coaxial lines. In short, there's nothing about the TF 1041A that he *doesn't* know.

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A.C. Measurement: Range: 0·05 to 300 volts, or to 2 kV using multiplier. Frequency Response: ±0·2 dB from 50 c/s to 450 Mc/s, —1 dB at 20 c/s, +2 dB at 1000 Mc/s. Input Impedance: 5 MΩ at 1 kc/s with 1·5 μμF in shunt.

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Resistance Measurement: Range: 0·2 ohm to 500 MΩ.

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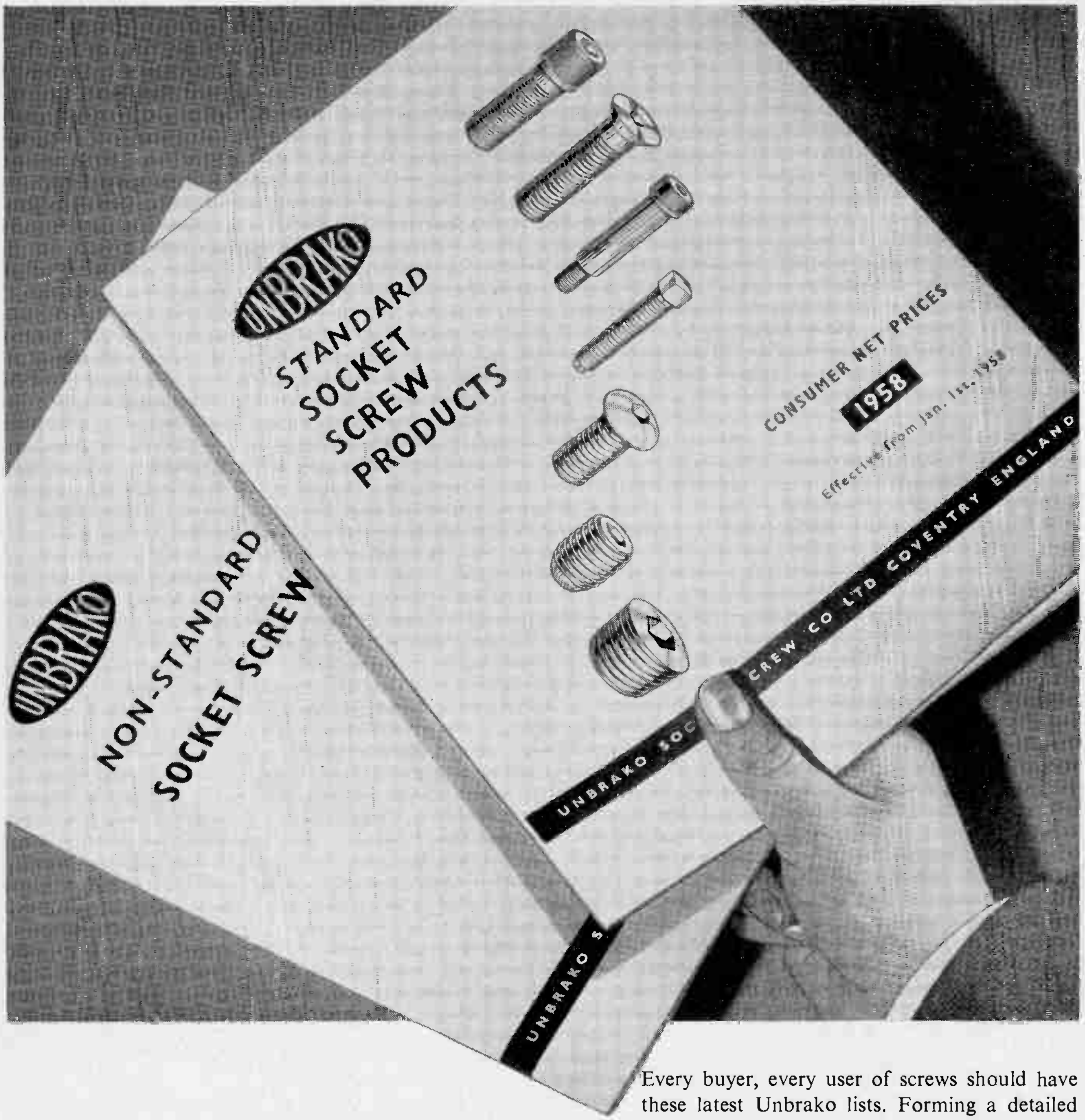

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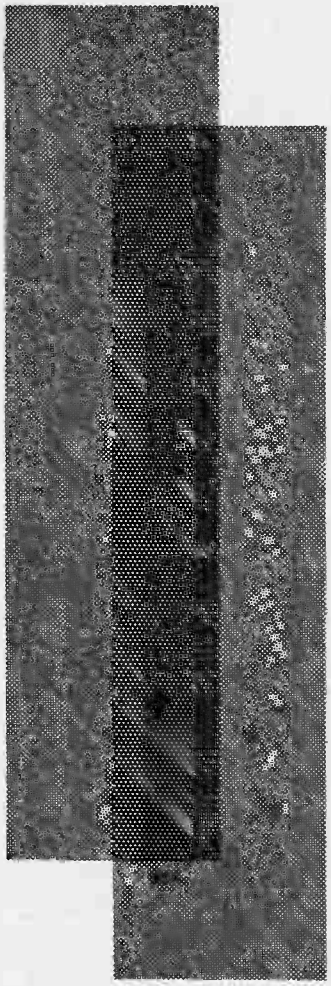


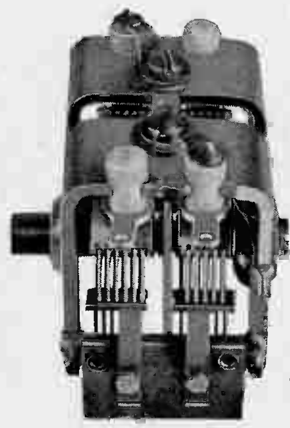
this new **Plessey** device provides

one knob tuning

This new idea brings one knob control for the medium wave and a preset long wave station and eliminates band switches for personal receivers. But the development achieves more than that: it effects considerable saving in component and assembly cost. For the relatively small cost of two capacitors you can now dispense with the long wave coils and switch!

This latest development has been achieved by Plessey at Havant by fitting a switch to their 'W' and 'V' type Variable Capacitors, which connects the extra capacitance required for the long wave station. The capacitor tunes the whole of the medium wave band and automatically switches to the long wave Light Programme at the end of its travel.

In small receivers, using Ferrite rod aerials, the long wave coil, switch and leads can be omitted, resulting in unimpaired efficiency of the medium wave coil. And again, where gramophone pick-up sockets are provided, this new Plessey tuning device could be used to mute the RF stages, a decided improvement on tuning to a quiet part of the wave band.



'W' Type Variable Capacitor

Patent Applied for

The Plessey Range

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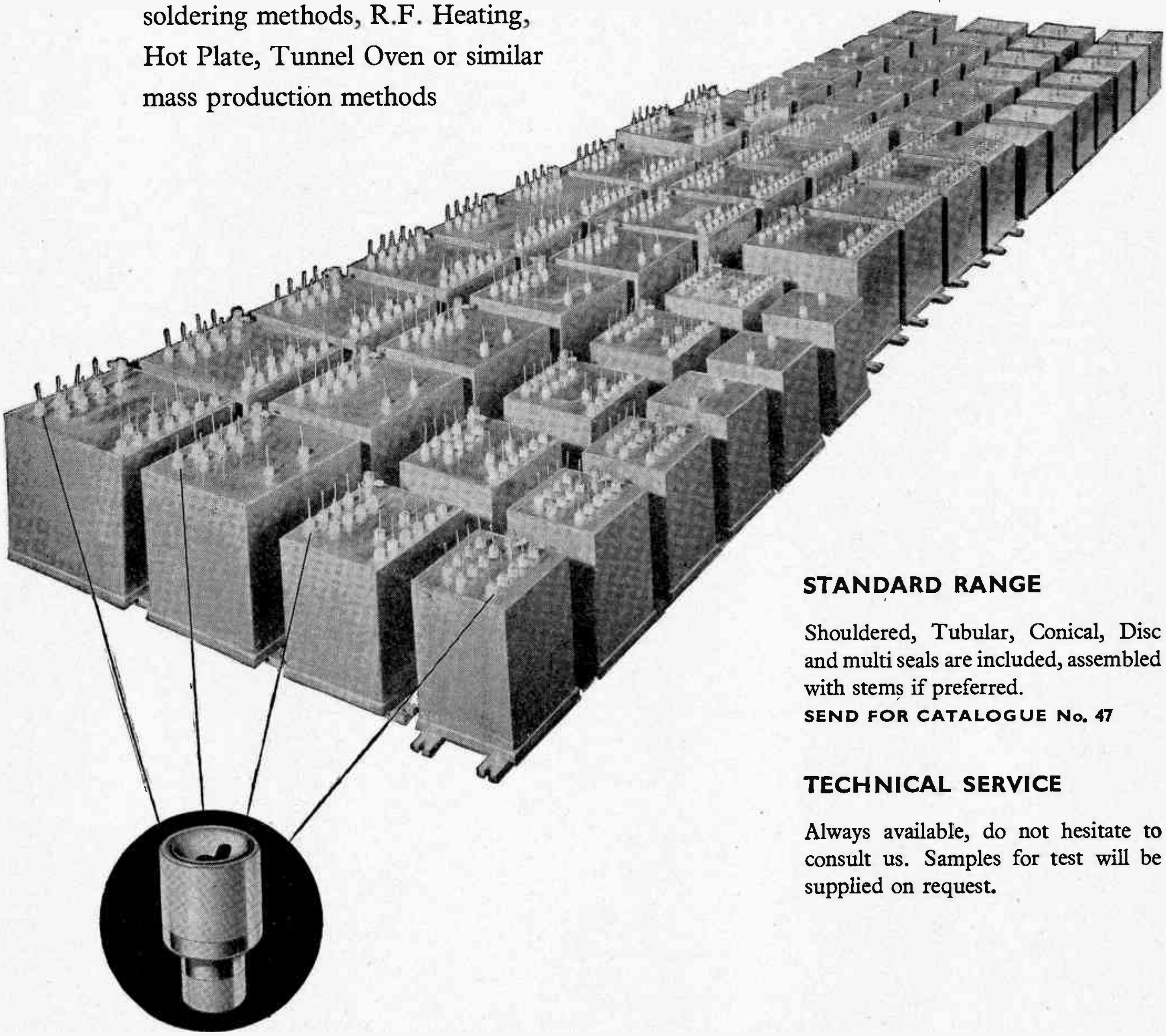

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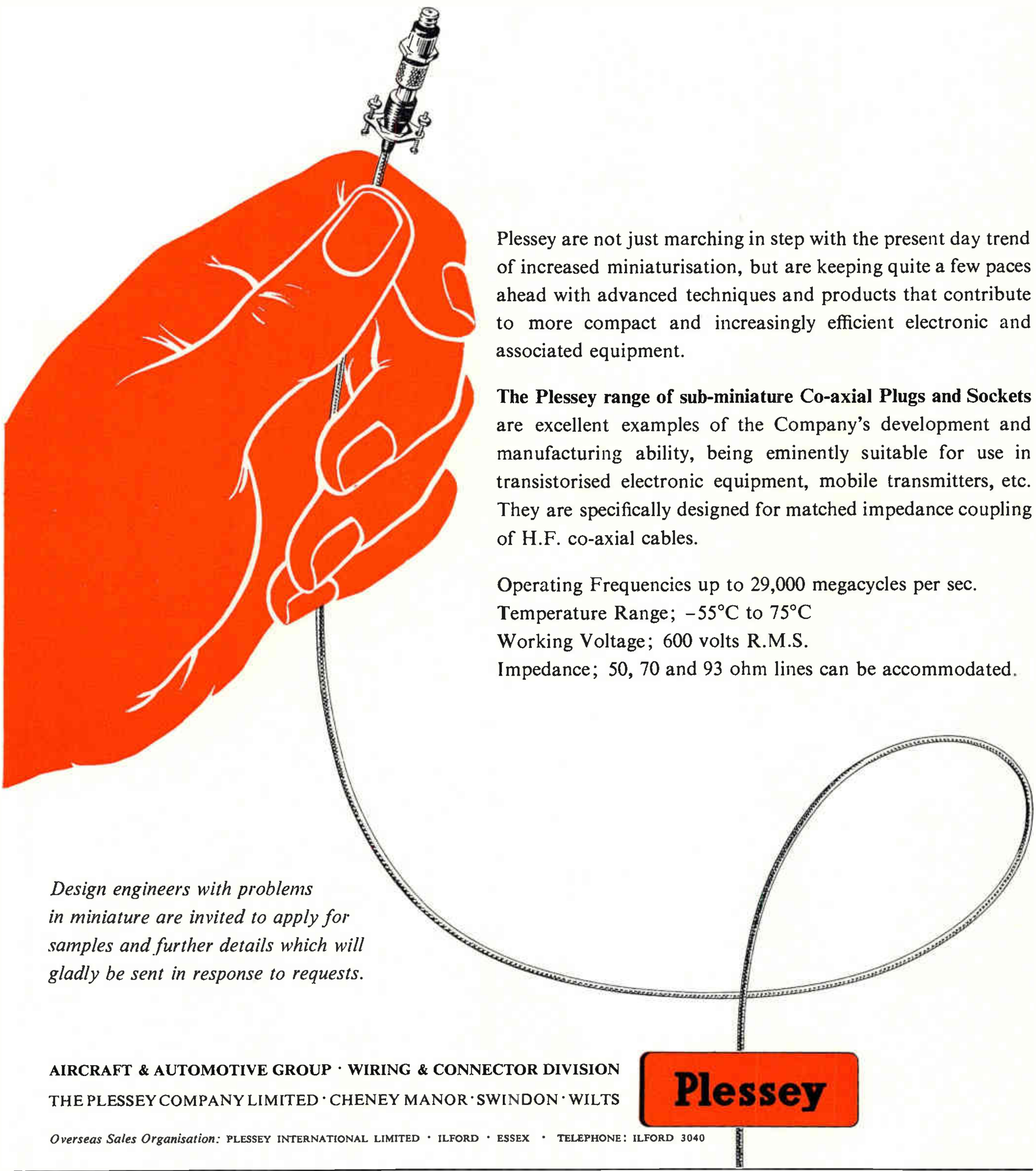

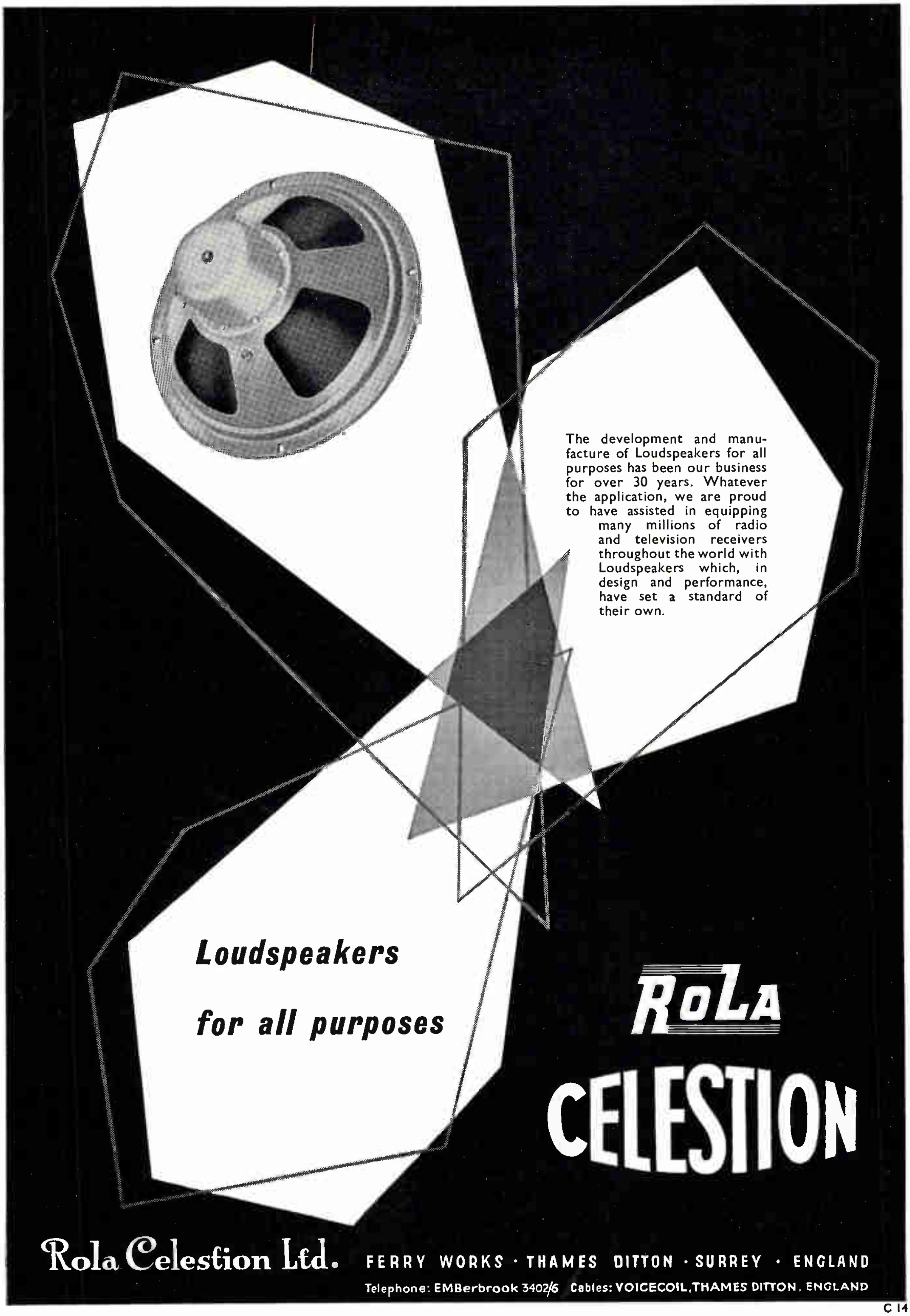
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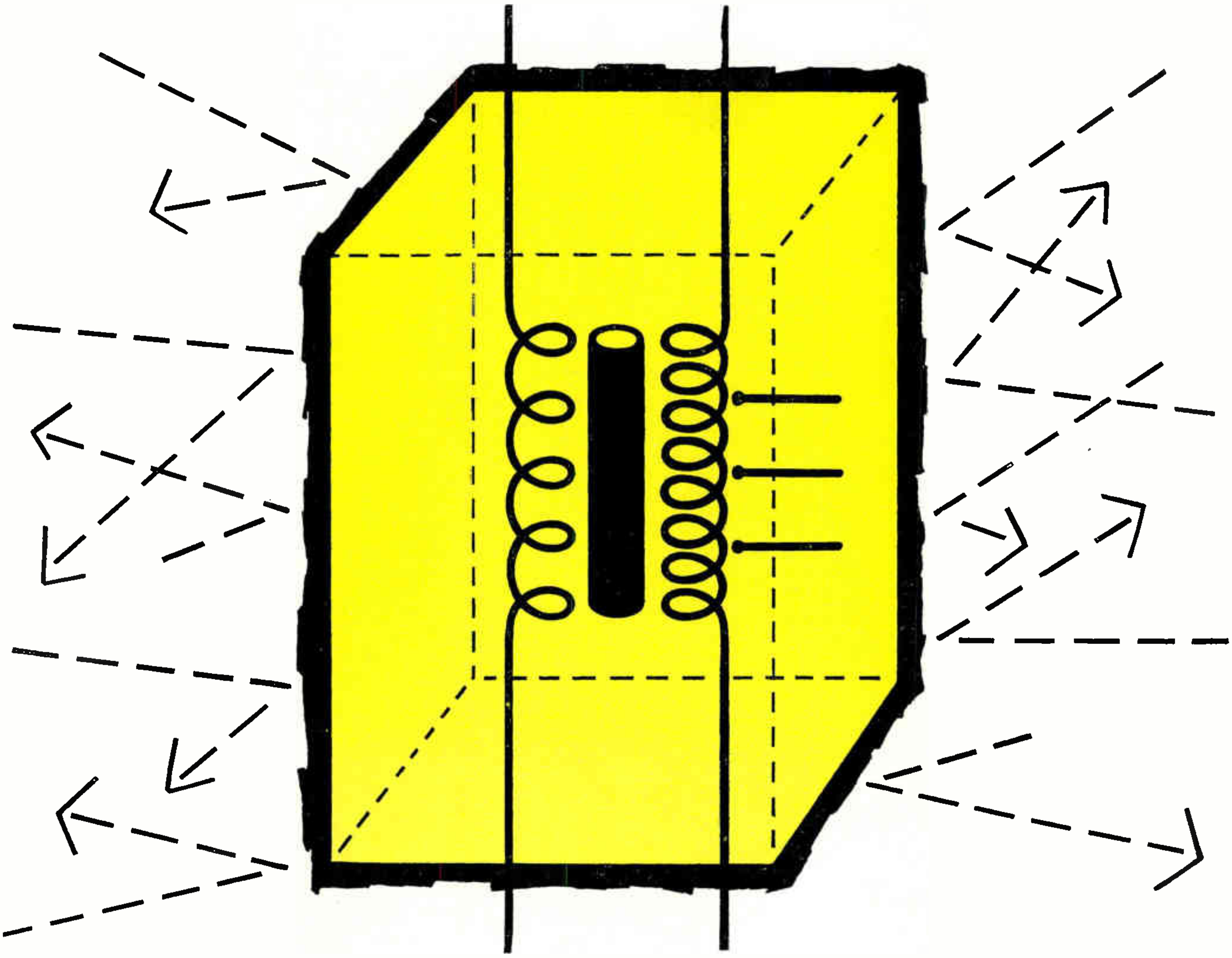

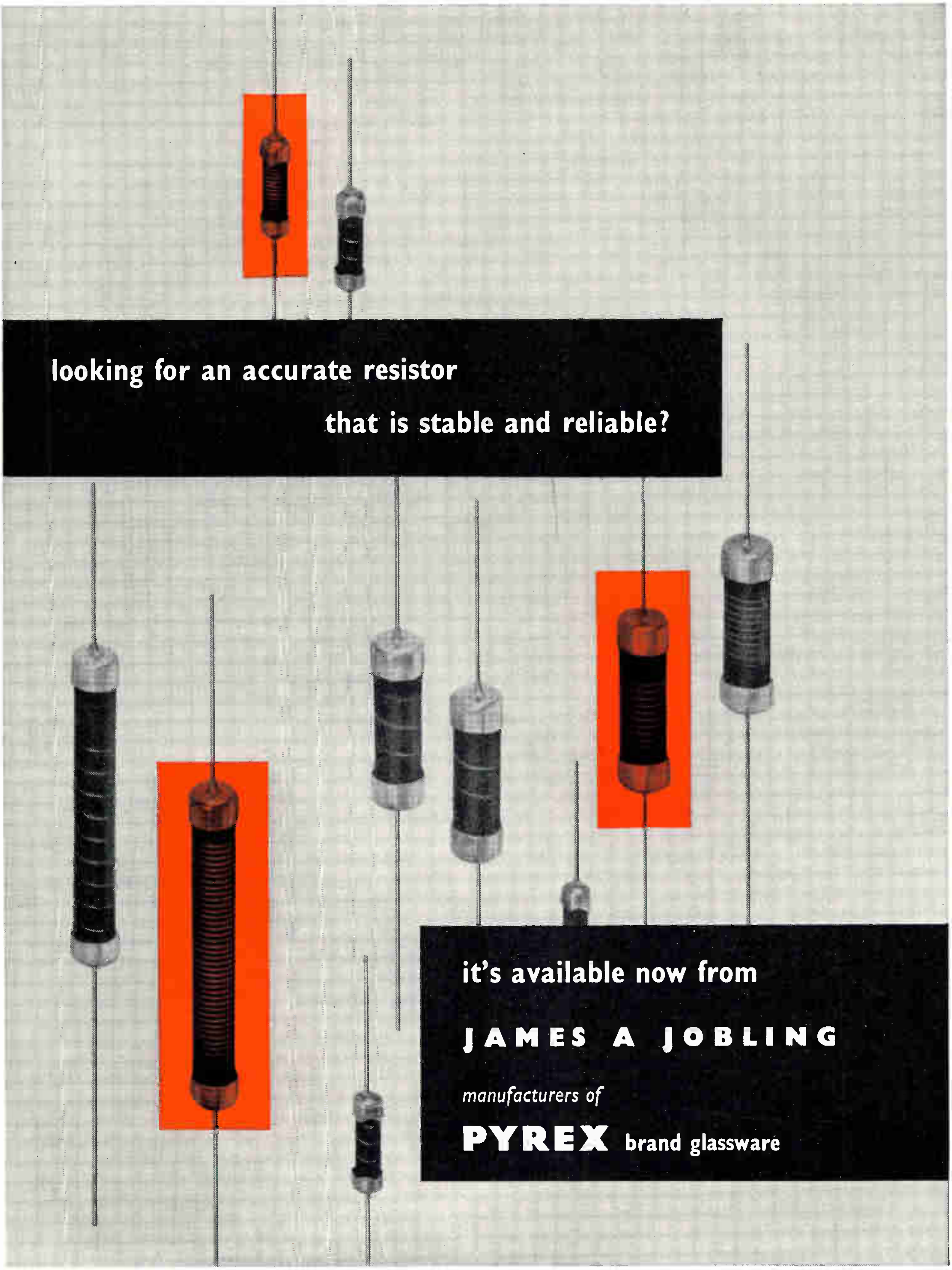
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the metal oxide film resistor style N

Jobling's metal oxide film resistor is manufactured from PYREX brand glass rod, to which is fused a thin metallic oxide film. This is spirally grooved and is then coated externally with moisture resisting varnish for added protection

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Together with these further advantages: the noise level is so low that it is difficult to measure, being largely independent of frequency. It has a low voltage coefficient and low inductive and capacitive reactance

Jobling's metal oxide film resistors style N conform with Joint Service specification RSC (PROV) 114

style	dimensions	
	length	diameter
N20	19/32″ ± 1/16″	11/64″ ± 1/32″
N25	15/16″ ± 1/16″	19/64″ ± 1/32″
N30	2$\frac{1}{16}$″ ± 1/16″	19/64″ ± 1/32″

ref. no.	style	wattage	voltage rating	ohmic value	
				minimum	maximum
2460	N20	½	350	100	25K
2461	N25	1	500	100	100K
2462	N30	2	750	200	250K

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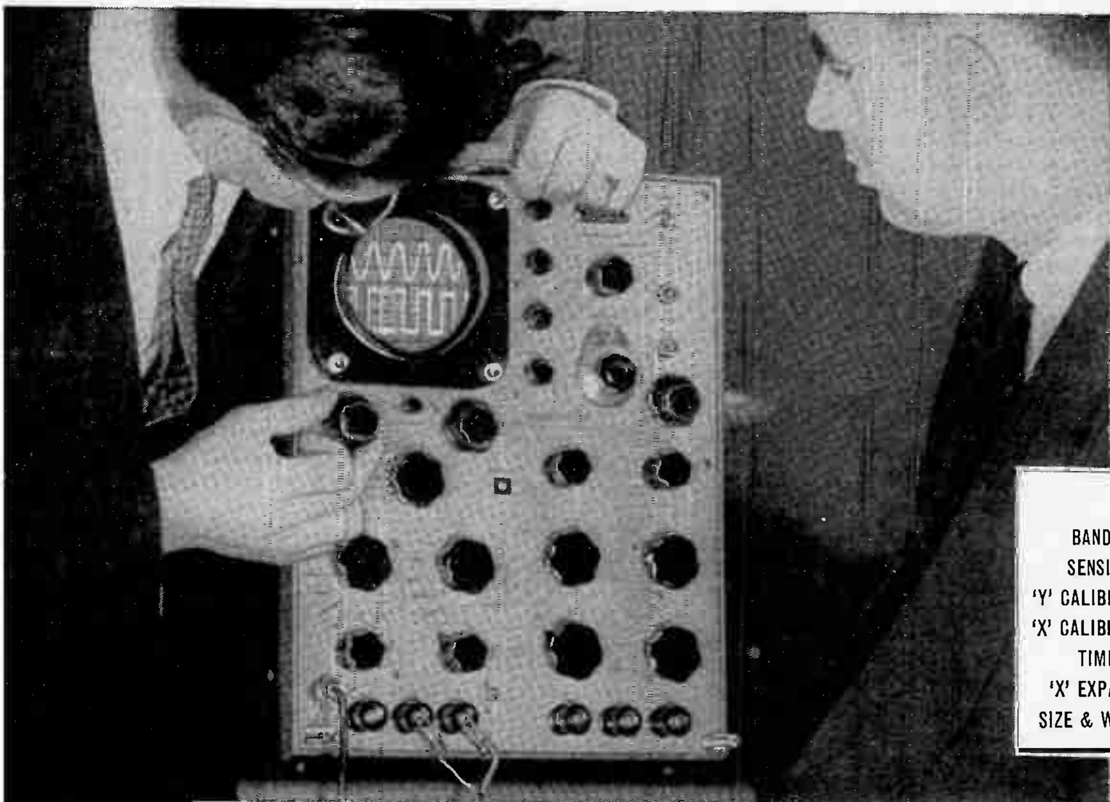
SOLARTRON oscilloscopes



Have you seen the *new* Solartron Double-Beam Solarscope CD 711 being put through its paces? Now, for the first time, you can choose a double-beam 'scope of Solartron quality, embodying all the latest design features and built to the finest standards of electronic and mechanical engineering. Why not write or call us now for a demonstration of the Double-Beam Solarscope, or any of the eight other models listed below? Specialist instrument engineers are immediately available to assist you, whatever your problem or field of application.

	CD 711 DOUBLE BEAM
BANDWIDTH	Max. D.C.—7 Mc/s. at 100mV/cm.
SENSITIVITY	3mV/cm.—100 V/cm.
'Y' CALIBRATION	Cal. shift. Accuracy ±5%
'X' CALIBRATION	Cal. time scale. Accuracy ±5%
TIME BASE	Time scale 0.3μ Sec./cm.—3 Sec./cm.
'X' EXPANSION	Continuously variable ×10
SIZE & WEIGHT	16″ × 13″ × 27½″ deep. 110 lb.

	CD 715	AD 557	CD 518	CD 568	CD 513
BANDWIDTH	Max. D.C.—20 Kc/s.	Max. D.C.—1 Mc/s.	Max. D.C.—5 Mc/s.	Max. D.C.—5 Mc/s.	Max. D.C.—10 Mc/s.
SENSITIVITY	10 mV/cm.—10V/cm.	3 mV/cm.—100 V/cm.	0.25 V/cm.—5 V/cm.	0.25 V/cm.—5 V/cm.	1 mV/cm.—10 V/cm.
'Y' CALIBRATION	Special facilities	Cal. shift ±5%	Shift meter. Accuracy ±3%	Shift meter. Accuracy ±3%	Cal. sensitivity Accuracy ±10%
'X' CALIBRATION	Special facilities	Cal. time scale. Accuracy ±10%	'Cal Pips' and sine wave. Accuracy ±2%	Sine wave. Accuracy ±2%	Cal. time scale Accuracy ±10%
TIME BASE	Sweep time 10 Sec. to 0.1 Sec.	Time scale 1μ Sec./cm.—1 Sec./cm.	Sweep time 100m. Sec.—1μ Sec.	Sweep time 100m. Sec.—1μ Sec.	Time scale 0.1μ Sec./cm.—1 Sec./cm.
'X' EXPANSION	×1, ×0.5, ×0.2, ×0.1, ×0.05	Continuously variable × 10			×0.5, ×1, ×2, ×5
SIZE & WEIGHT	14″ × 10″ × 20″ deep. 47 lb.	16½″ × 10″ × 22″ deep. 70 lb.	12″ × 9″ × 18″ deep. 40 lb.	12″ × 9″ × 18″ deep. 40 lb.	16½″ × 10″ × 22″ deep. 70 lb.

	CD 523S	CD 814	CD 643
BANDWIDTH	Max. D.C.—10 Mc/s.	Constant 1c/s.–9Mc/s.	Constant D.C.—15 Mc/s.
SENSITIVITY	1 mV/cm.—10 V/cm.	30 mV/cm.—30 V/cm.	100 mV/cm.–60 V/cm.
'Y' CALIBRATION	Cal. sensitivity Accuracy ±10%	Comparison method Accuracy ±5%	Cal. shift. Accuracy ±2%
'X' CALIBRATION	Cal. time scale Accuracy ±10%	'Cal Pips' Accuracy ±5%	'Cal Pips' and brightup Accuracy ±2%
TIME BASE	Time scale 0.1μ Sec./cm.—1 Sec./cm.	Repetition Rate 6 c/s—185 Kc/s.	Time scale 0.1μ Sec/cm.—100m. Sec./cm.
'X' EXPANSION	× 0.5, × 1, × 2, × 5	Continuously variable × 10	Continuously variable × 100
SIZE & WEIGHT	16½″ × 10″ × 23″ deep. 70 lb.	14½″ × 10½″ × 19½″ deep. 43 lb.	20″ × 14¾″ × 27¾″ deep. 140 lb.

STOP PRESS

The CD 711 Solarscope has now received Joint Service Approval and has been designated Type No. CT 414. Full details available on request.

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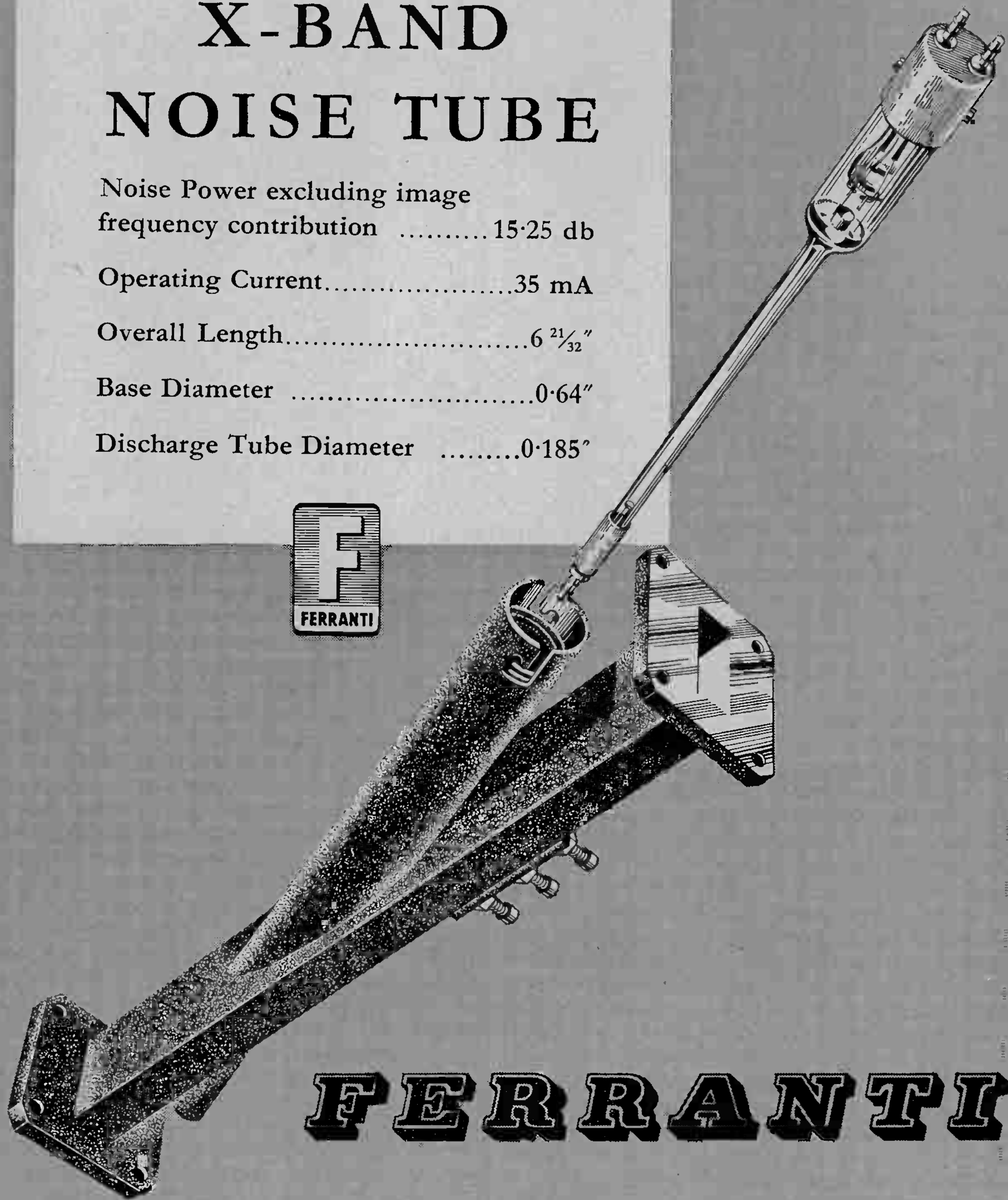
Design Approved

X-BAND NOISE TUBE

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

FERRANTI



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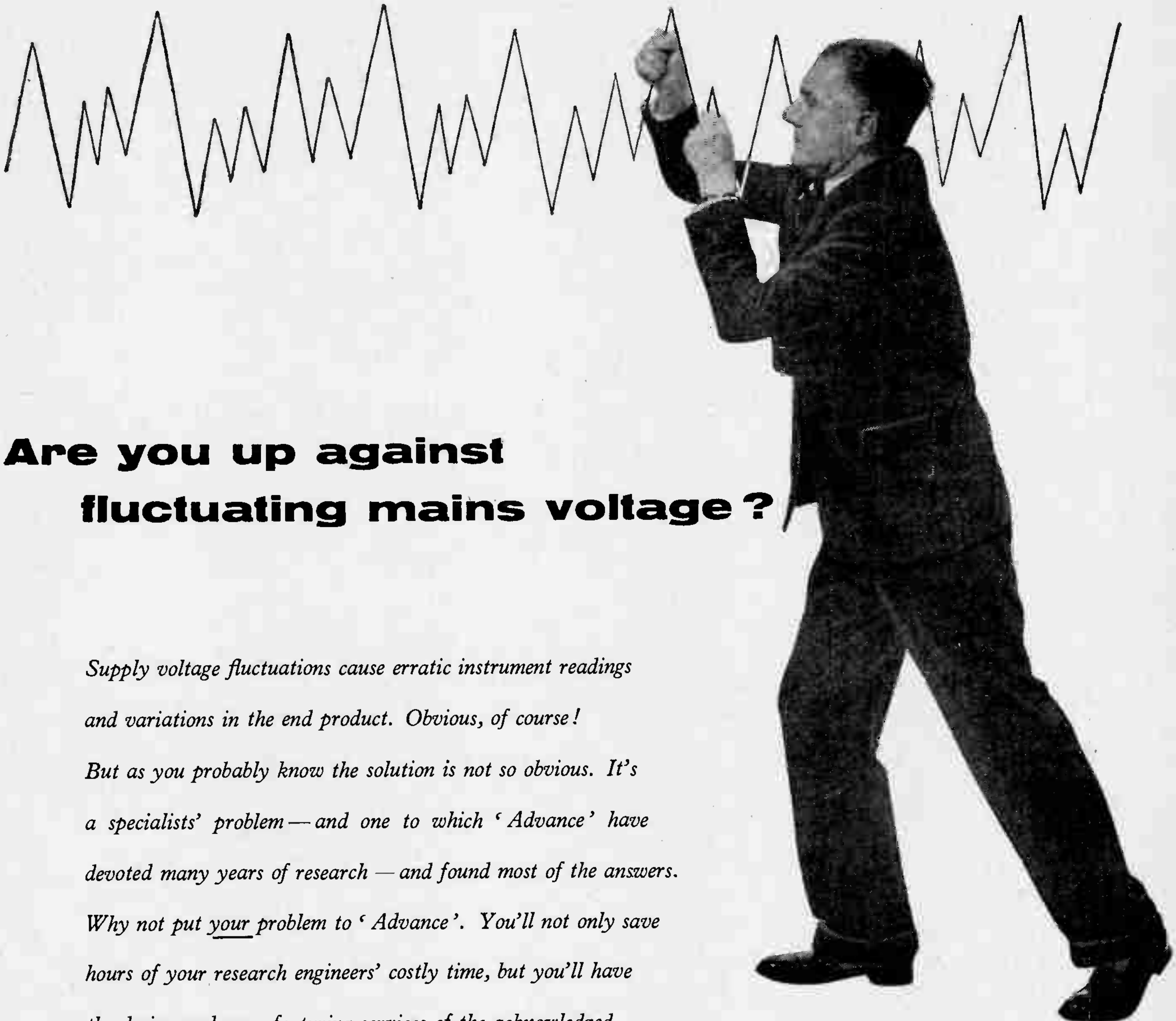

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VOLUME 35 NUMBER 6 JUNE 1958

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VOLUME 35 NUMBER 6 JUNE 1958

ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

Sound Reproduction

THERE is now an enormous interest in high-quality sound reproduction. In spite of the fact that the term 'high-fidelity' is often used to describe it, the aim is no longer to achieve perfection. By this, we mean the production at the listener's ears of precisely those variations of air pressure which would exist if he were in the concert hall itself.

It is now recognized that such perfection is impossible with loudspeaker reproduction because of the acoustics of the listening room. The original sound is affected by the acoustics of one room only, the reproduced by those of two rooms.

Generally speaking, the aim now is to produce the most pleasing effect. Opinions differ about what this is, but it is generally agreed that the first essential is to keep to a minimum all distortion caused by non-linearity of response and that the second is to avoid marked peaks and troughs in the amplitude–frequency characteristic. When this is done, tone and volume controls provide the user with the ability to adjust the performance for what he considers the most pleasing result.

The possibilities have recently been extended, however, by the introduction of stereophonic records. Discs are used in which the necessary two channels are recorded on the two sides of a single groove. The sides are inclined at an angle of about 45° to the face of the disc, and the modulations are at 90° to each other. A single stylus is used in the pickup and it is, in effect, coupled to two transducers so that the complex stylus movement is resolved into two components at right-angles to actuate the transducers independently. Two amplifiers and two spaced loudspeakers are, of course, needed.

Stereophonic reproduction with loudspeakers can hardly give perfection, but it certainly does give an impression of size and position to sounds. It may thus be a distinct advance in the quest for more pleasing reproduction.

It needs more equipment than single-channel reproduction and so it is more expensive if the same grade of apparatus is employed in both cases. Stereophonic reproducers are being made in various grades at various prices and this raises an interesting point.

If it is conceded, which not necessarily everyone would do, that stereophony is an improvement over single-channel reproduction when both employ the same grade of apparatus, which is the better when both cost the same? In that case, the stereophonic effects will be obtained at the expense of more distortion and/or less volume. It is a matter for experiment and, doubtless, opinions will differ.

Whether or not stereophony is liked, it seems probable that it will be confined to recordings. The use of two radio channels for one programme can hardly be contemplated, save for an experiment such as that recently carried out by the B.B.C. At v.h.f. there is the possibility of using both a.m. and f.m. on one carrier, but we doubt if we shall see that for a long time to come.

Manufacture of Silicon Transistors

AN ASSESSMENT OF THE PRESENT STATE OF TECHNOLOGY

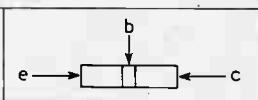
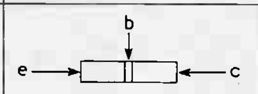
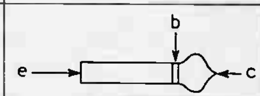
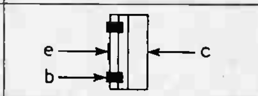
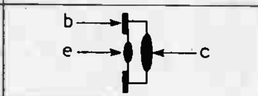
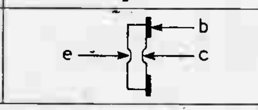
By James T. Kendall, M.A., Ph.D., F.Inst.P.*

The germanium transistor has now become a fairly familiar electronic component to most circuit engineers, and its properties and limitations are widely known. Its chief drawback is the comparatively low ambient temperature in which it will operate. Although 85°C is sometimes quoted as a maximum junction temperature for a germanium transistor, its expectation of life is generally severely limited under such conditions, and a more realistic figure would be in the region of 50°C.

For a silicon transistor maximum junction temperatures of at least 150°C are permissible with good life-expectancy so that, when transistor circuits are considered which must operate in high ambient-temperature conditions, it is essential to use silicon. Even where temperature conditions are only moderate, silicon transistors can offer greater reliability.

For a variety of reasons the commercial production of silicon transistors has been delayed, but they have recently become generally available in this country in production quantities, and the purpose of this article is to assess the present state of their technology.

TABLE I

1		n-p-n GROWN, DOUBLE DOPED	TEXAS BOGUE
2		n-p-n GROWN, DIFFUSED	TEXAS TRANSITRON
3		n-p-n DIFFUSED, MELT-BACK	G.E.
4		n-p-n DIFFUSED	TEXAS G.E. TRANSITRON
5		p-n-p ALLOYED	INTERMETALL RAYTHEON MULLARD
6		SURFACE ALLOYED	PHILCO SEMICONDUCTORS

Materials

The problems associated with the production of a very pure basic material are much greater for silicon than for germanium. The reasons are its much greater chemical reactivity, its higher melting point and the unfortunate circumstance that purification by zone-melting is relatively inefficient, particularly in the case of the impurity boron. As a result, we have the paradoxical situation that in the very pure state the common element silicon is at present much more expensive than the rather rare element germanium. In the long run, this situation must inevitably change, and silicon will become the cheaper basic material.

At the present time, there are a number of sources of silicon of sufficient purity for making transistors, although the only considerable non-dollar source is Pechiney et Cie. of France. However, a number of companies in this country are well advanced on pilot-scale production and by 1959 there may well be an embarrassing surplus of silicon.

Although a solution to the basic raw material supply problem is in sight, there are still some difficulties to be overcome in subsequent processing of the pure silicon. In the case of germanium, the pure material can be melted in contact with graphite at 940°C with very little fear of contamination, purification by zone-melting can be carried out with ease, and perfect crystals almost free from dislocations can be grown. For silicon the situation is much less happy. The pure material can be melted in contact with silica at 1420°C, but continuous contamination from the crucible takes place. The normal technique of final purification by zone-melting in a horizontal boat therefore becomes impossible. This difficulty can be overcome by the floating-zone technique, but the quantity which can be handled in this way at one time is comparatively small, and the removal of boron, because of its unfavourable segregation coefficient, is a very lengthy process.

Much work still remains to be done on improving crystal perfection. At present the best silicon crystals have a dislocation density several orders of magnitude higher than in the best germanium crystals. A further field in which much more development work is required is in the study of the effects of heat treatment on the resistivity and minority-carrier lifetime of silicon crystals. These effects are of importance in the fabrication of silicon transistors, and there is no doubt that improved device characteristics will be obtained as a result of their study.

In addition to the bulk properties of any semiconductor material, the surface properties are obviously

* *Texas Instruments Limited, Bedford.*

of great importance. In this field the state of silicon technology is probably no worse than that of germanium. In both cases the effect of various ambient atmospheres has been studied—especially for water vapour and oxygen—and is partially understood. Preliminary work on the effect of various oxidizing and reducing agents on surface recombination velocity and surface ionization levels has also been carried out, but little practical use can yet be made of the results. This research work which is at present being carried out on semiconductor surface phenomena is likely to be of even greater importance for silicon than it is for germanium devices, both because of their higher operating temperatures and because of their smaller bulk reverse currents.

Fabrication Techniques

The six classes of silicon transistor which are now commercially available are shown in Table 1. Their methods of fabrication will be described briefly, and some of their chief characteristics will be pointed out.

Grown, Double-Doped n-p-n Silicon Transistors

These were the first silicon transistors to become generally available. They were introduced commercially by Texas Instruments Incorporated in America, about the middle of 1954. Their method of manufacture is briefly as follows:

First, one grows a single crystal by pulling it from a melt, and the crystal is shaped as shown in Fig. 1(a). The original silicon is doped to give the desired collector resistivity, approximately 1–2 ohm/cm. When roughly half the crystal has been pulled, additional p-type dope is added to give the desired base resistivity and very shortly afterwards, further n-type dope is added to give the desired emitter resistivity, approximately 0·01 ohm/cm. The width of the central p-region is controlled by the temperature and rate of pulling (which together determine the rate of crystal growth) and by the time interval between the additions of dope. A big advantage of this method of fabricating a grown-junction is the wide and independent control which one has over the resistivity of the collector, base and emitter regions. A typical grown crystal of this type will weigh about 40 grams, and it can be seen from the shape of the crystal that a large portion of it can be used for making transistor bars. This is of course the reason for this rather squat and wide form of crystal.

Rectangular bars are then cut out of the crystal and are shaped as shown in Fig. 1(b). Some 400 bars are obtained from one crystal, and one can grow a crystal in about half an hour. It is therefore apparent that, from the point of view of the material used and the time taken, the method does not compare unfavourably with the alloying process. By etching the bars in the usual CP4 etch the emitter and collector ends can be very easily distinguished, as they etch at different rates, and one end of the bar will thus appear rather dull and the other end rather shiny. This is shown clearly in Fig. 3, which also shows the base connection. The remainder of the process is simply to mount the bar on a base and to attach three connections to the emitter, base and collector regions. The manner of doing this is shown in Fig. 2. The only connection which is at all difficult to make is the connection to the very thin base layer. This

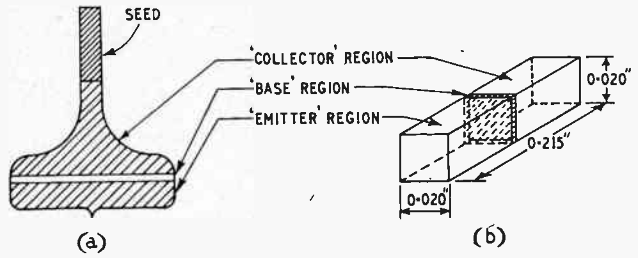


Fig. 1. *Silicon crystal and bar cut from it; (a) silicon crystal (actual size), (b) silicon bar (enlarged)*

layer is generally only 0·0001 to 0·0004 inch wide, and the connection is made with aluminium wire about 0·003 inch diameter. Since aluminium is an acceptor element, this wire does not short out the junction, but merely enlarges it in the immediate vicinity of the connection.

This double-doped type of transistor is not at all difficult to make, but one does require a rather elaborate crystal puller in order to give accurate control over the programming of temperature, pull rate and doping interval. Such a crystal puller is shown in Fig. 4, while a close-up of a growing crystal is shown in Fig. 5.

Grown-Diffused n-p-n Silicon Transistors

A modification of the above process results in the so-called grown-diffused[1] transistor, first developed by Texas Instruments Incorporated. The process of manufacture differs only in the crystal-growing procedure, which results in a much narrower base region with a graded collector junction instead of an abrupt one. The method, in this case, is to grow the collector portion of the crystal as previously and then to add acceptor and donor impurities simultaneously instead of successively.

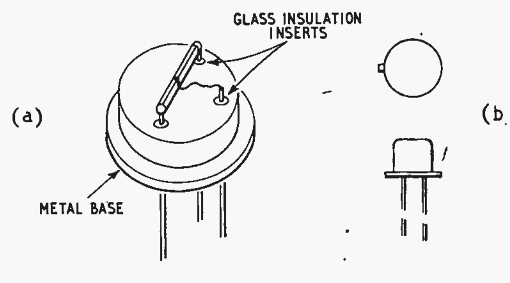
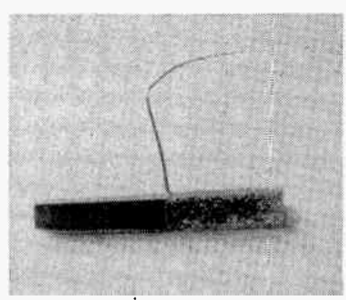


Fig. 2. *Grown, double-doped silicon transistor; (a) internal arrangement (enlarged), (b) actual size*

Fig. 3. *Etched double-doped bar with base connection. The collector and emitter regions are distinguishable after etching*



The crystal is then grown to completion, and during the growth of the lower half or emitter section of the crystal, the acceptor impurity diffuses into the collector or upper half for a short distance. It should be noted that in the case of silicon, acceptors diffuse more quickly than donors—in fact, they diffuse about twenty times faster. The result of the diffusion is thus to produce an n-p-n structure just as before. As previously, the crystal is subsequently cut up into bars and mounted on a header with three connecting wires. Because of the very narrow base regions which can be obtained by this method, these transistors are more suitable for higher-frequency devices. Although classes 1 and 2 have been described separately, the line of demarcation between them has become somewhat blurred recently as some grading of the collector junction is generally deliberately produced in double-doped crystals by temporarily raising the temperature in order to allow diffusion to take place.

Diffused, Melt-Back n-p-n Silicon Transistors

A further modification to the process results in the so-called diffused, melt-back transistor. This process has been developed by the General Electric Co. in America and, although the method of manufacture is quite different, the end result is a transistor which is practically indistinguishable from the second class.

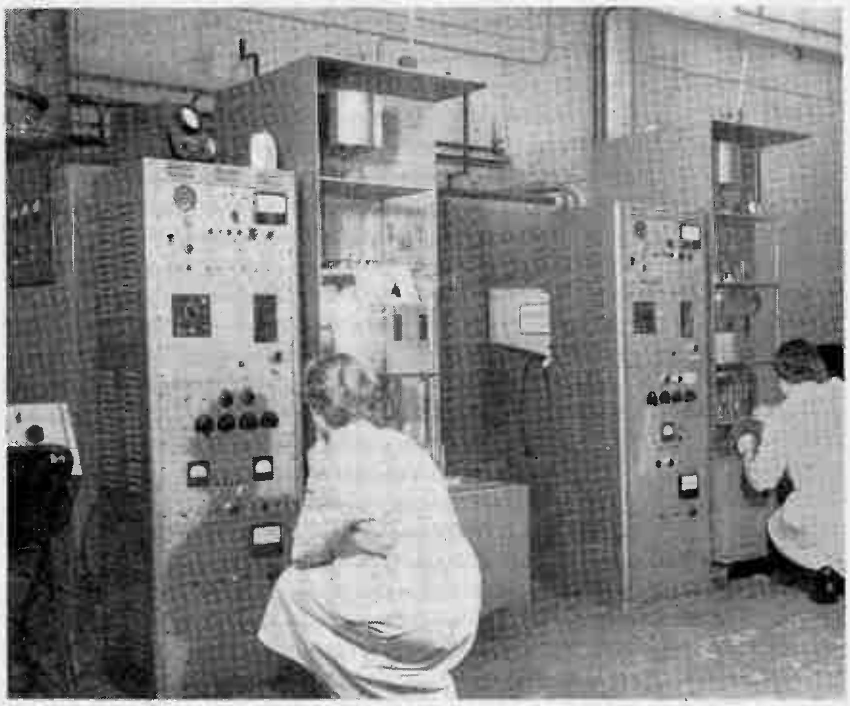
In this case, the diffusion is carried out after the crystal has been cut up into bars, and not during the crystal growth. Thus there is no necessity to grow a short, fat crystal such as is grown for double-doped and grown-diffused types, and the normal sort of carrot-shaped crystal is therefore grown. This crystal is doped right from the start with both p- and n-type impurities at the levels shown in Fig. 6(a). The crystal is then cut up into bars and one end of the bar is melted back about half-way, as shown in Fig. 6(b). By this process the original impurity concentrations are altered by segregation to the distribution shown in Fig. 6(b). It will be seen that we still have an n-type crystal throughout, but there is an abrupt change at the point where the melting-back process ceased. The bars are then placed in a furnace at an elevated temperature, approximately 1250°C, which must obviously be very closely controlled, and diffusion of the impurities in the bar now takes place. A p-type layer is formed owing to the faster diffusion of the acceptor impurities, and the final distribution after diffusion has taken place is shown in Fig. 6(c). As in the second type, there is a graded collector junction and a very thin base layer, thus providing a high-frequency, high-collector-voltage transistor. The final net impurity concentrations are as shown in Fig. 7 and this figure might, in fact, equally well refer to the previous grown-diffused class. Based on the properties of the resulting final transistor there is really nothing to choose between these two classes and both are being successfully made in fairly large quantities.

Diffused n-p-n Silicon Transistors

The fourth class of transistor is quite different from the first three. It is made by diffusion of impurities from the gas phase into a solid wafer of silicon. This class was originally described by the Bell Telephone Laboratories in the *Bell System Technical Journal*, and in that case it referred to a very-high-frequency type, both germanium and silicon; in the case of germanium, reaching something like 500 Mc/s, and in the case of silicon about 200 Mc/s. Since that time, quite a bit of further development work has gone ahead both at Bell Telephone Laboratories and in other places and, as an example of this class of transistor, it is proposed to describe the 37·5-watt power transistor which is at present being made by Texas Instruments Limited in this country.

First of all, one grows a single-resistivity silicon crystal, the normal sort of carrot-shaped type of crystal. This is doped to give a resistivity of round about 7 ohms/cm, n-type, and one tries to obtain a uniform resistivity throughout the crystal. One also tries to obtain minimum dislocation density and maximum minority carrier lifetime. This crystal is then cut up into wafers 0·015 inch thick, cut normal to the direction of growth of the crystal. These wafers are placed in a furnace at an accurately controlled temperature of about 1250°C together with sources of donor and acceptor impurities. The usual way of doing this is to seal up the wafers together with sources of the impurities in a quartz tube, and then to put the tube inside a a furnace. The diffusion of the impurities from the gas phase into the silicon wafer takes place, and

Fig. 4. *Crystal puller for grown double-doped transistors*



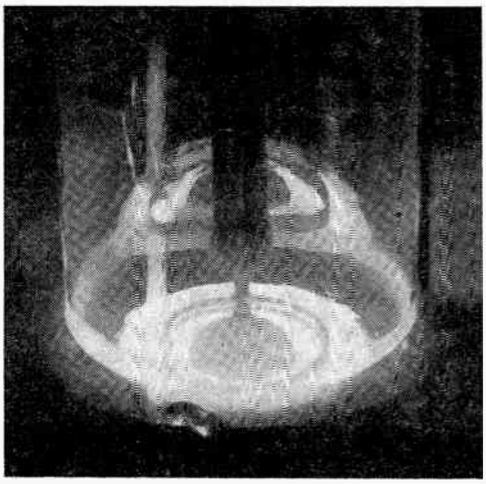
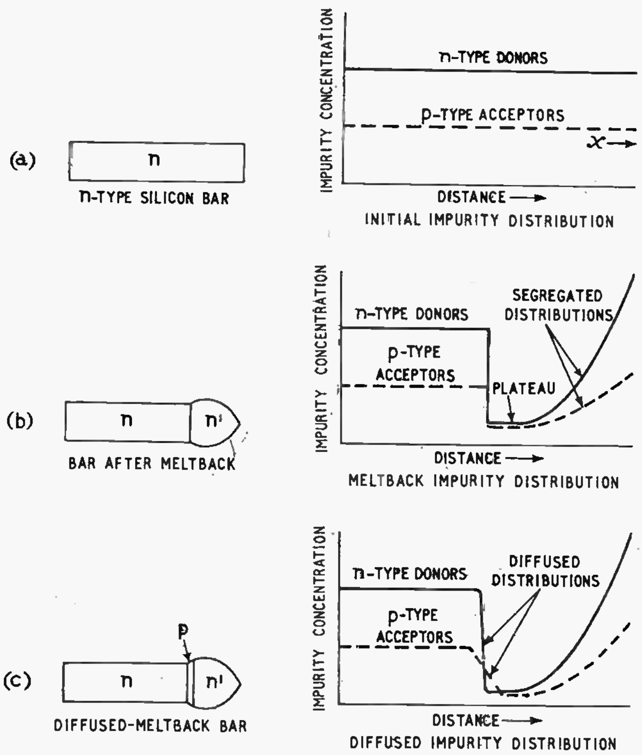


Fig. 5. *Close-up of growing double-doped crystal*

of course during this process, which may take anything up to thirty or forty hours, the temperature of the furnace must be accurately controlled. Now, owing to the faster diffusion of acceptors, the wafers become coated with a skin consisting of an inner p-layer and an outer n-layer. This skin is lapped off one side of the wafer and we thus obtain an n-p-n structure consisting of the main body of the crystal, which is n-type, and two thin p and n skins on top. We have, therefore, made our basic transistor and the only problem is to make contacts to the three regions. The collector region is quite easy; one simply solders the wafer to a heat-sink. One may initially plate the silicon with nickel in order to get a good contact to the main body of the n-type material. Contact to the p-layer is made by alloying an aluminium ring into it from the top. This aluminium ring is shown in section in Fig. 8. The aluminium ring is alloyed right through the top n-layer and, since one obviously cannot control the depth of alloying too precisely, it will go through the p-region as well. So, in fact, this aluminium ring is making physical contact to all three regions, but as aluminium is an acceptor element it makes electrical contact only to the p-type layer, which becomes enlarged in the immediate vicinity of the ring. The emitter contact is finally made by electro-plating a circular contact on to the top surface. The physical configuration of the final transistor is shown in Fig. 8, and it is seen that this kind of structure is very suitable indeed for making a power transistor, since heat can readily be removed from the junctions. Very much lower values of collector saturation resistance are obtained in this structure as compared with the grown-junction structure, which again makes it a more suitable one for handling larger currents.

Fig. 6. *Impurity distributions at each step of the silicon diffused-meltback process*

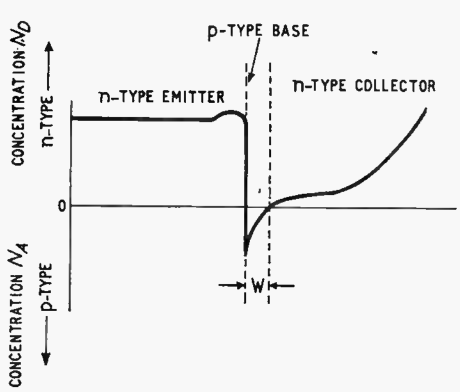


Alloyed p-n-p Silicon Transistors

In the development of silicon transistors, most companies (with one or two notable exceptions) have concentrated their attention on this method of fabrication. Having successfully made alloyed germanium transistors they have hoped to apply the knowledge thus gained to the fabrication of silicon transistors by the same method. Unfortunately, it has turned out that alloyed silicon transistors are exceedingly difficult to make, This is due mainly to the higher alloying temperature which must be used—800°C instead of 500°C—as this makes the control of base-layer thickness even more difficult than it is for germanium. There is also the difficulty that the degree of crystal perfection at present available in silicon is nowhere near as good as it is in germanium; this also tends to make the penetration of alloying dots somewhat variable.

The method of fabrication is very similar to that used for alloyed germanium transistors, and will therefore not be given in detail. Aluminium dots are used for the emitter and collector, which are alloyed on opposite sides of a thin silicon wafer. The base connection

Fig. 7. *Net impurity concentrations for silicon diffused-meltback bar*



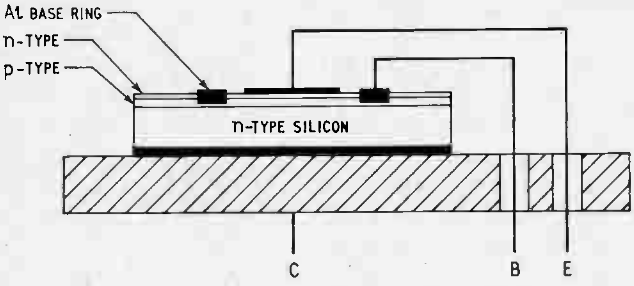


Fig. 8. *Internal arrangement of diffused power transistor*

generally takes the form of a gold ring, lightly alloyed around the emitter dot.

One serious drawback to the alloyed silicon transistors made to date is their mechanical instability when subjected to temperature cycling. This is due to the large difference in thermal expansion coefficients between aluminium/silicon eutectic and silicon. This difficulty becomes more pronounced the larger the area of the alloy dots, and it may possibly preclude the development of an alloyed silicon power transistor, although, as in the case of the diffused structure, quite low values of collector saturation resistance are obtained by this method of fabrication.

Surface-Alloyed p-n-p Silicon Transistors

A modification of the above process has been developed by Philco Corporation in America. The variable penetration which occurs in the normal alloying process is avoided by first obtaining a thin base region by means of electrolytic jet etching on opposite sides of a silicon wafer. Aluminium is then evaporated in vacuo into the etched regions, and a very slight alloying is subsequently carried out. In this way, very thin, uniform base layers can be obtained and, since the amount of aluminium is small, there is little trouble with thermal expansion mismatch. A photograph illustrating the jet-etching technique is shown in Fig. 9.

Types of Commercially Available Silicon Transistors

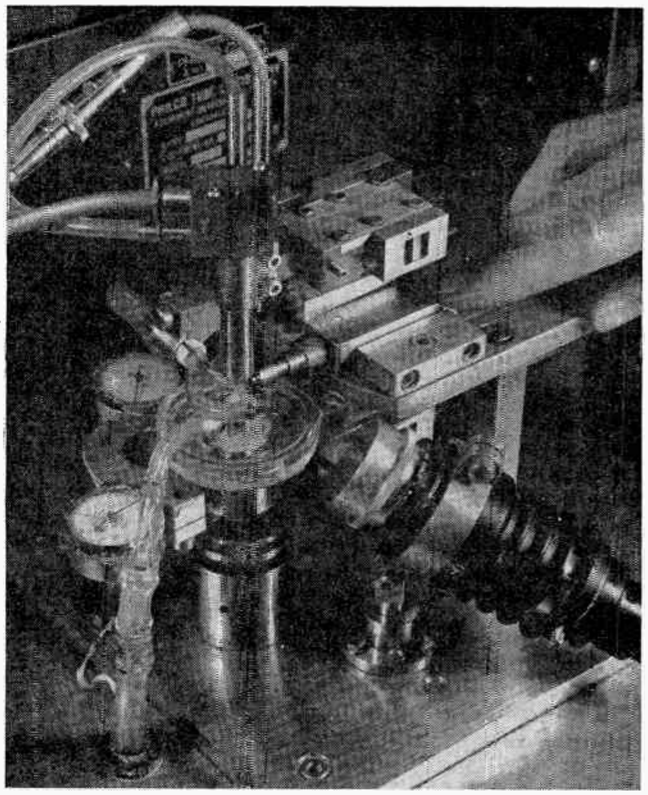
In Table 2 are listed the types of silicon transistor at present available commercially from U.K. sources. It will be seen that among small-signal transistors they include alloyed p-n-p types (Mullard and Semiconductors), and grown-junction n-p-n types (Texas). In passing, it should be noted that Semiconductors' transistors are made by a surface-alloying technique which, as explained above, differs somewhat from the normal alloying process and, in consequence, they do not suffer so badly from the difficulties of mechanical damage on temperature cycling associated with differences in expansion coefficients between aluminium and silicon. It will also be noted that the alloyed types have considerably lower values of collector saturation resistance than the grown types, and this

TABLE 2

Silicon Transistors available from U.K. sources

Type No.	Manufacturer	Maximum Ratings			Typical Characteristics				Construction
		Dissipation at 100°C (watts)	Collector Current (mA)	Collector Voltage (V)	Current Transfer Ratio (β)	Alpha Cut-off Frequency (Mc/s)	Collector Saturation Resistance (ohms)	Collector Capacitance (pF)	
A. Small	**Signal Types**								
OC200	Mullard	0·1	15	25	20	1	Very low	45	Alloyed, p-n-p
OC201	,,	0·1	15	25	30	4	,,	50	,,
2S001	Texas	0·1	25	45	15	8	150	7	Grown, n-p-n
2S002	,,	0·1	25	45	25	10	150	7	,,
2S003	,,	0·1	25	45	40	16	150	7	,,
2S004	,,	0·1	25	45	50	12	150	7	,,
2S015	,,	0·1	25	45	100	12	150	7	,,
2S014	,,	0·05	20	40	60	20	100	1·6	,,
2S005	,,	0·05	20	40	100	30	100	1·6	,,
2N354	Semiconductors	0·05	50	25	>10	10	Very low	<12	Alloyed, p-n-p
2N355	,,	0·05	50	10	>10	35	,,	<12	,,
					Minimum	*Power Gain*			
3S001	Texas	0·05	10	30	18 dB at	12·5 Mc/s	150	1·8	Grown, n-p-n
3S002	,,	0·05	10	30	16 dB at	30 Mc/s	150	1·8	,,
3S003	,,	0·05	10	30	20 dB at	4·3 Mc/s	150	1·8	,,
B. Large	**Signal Types**								
2S006	Texas	0·4	60	55	15	—	125	—	Grown, n-p-n
2S007	,,	0·4	50	85	15	—	150	—	,,
2S008	,,	0·4	40	125	15	—	175	—	,,
2S009	,,	0·4	60	60	15	—	150	—	,,
2S010	,,	0·4	60	60	30	—	150	—	,,
2S017	,,	2·25*	200	60	20	4	20	—	Diffused, n-p-n
2S018	,,	2·25*	200	100	20	4	20	—	,,
2S012	,,	15*	2000	60	30	4	3	—	,,
2S013	,,	15*	2000	60	20	4	6	—	,,

* With Heat Sink.



(Courtesy Semiconductors Ltd.)

Fig. 9. *Precision-etch head for surface-alloyed transistors*

generally makes them more efficient for switching purposes.

The fact that the alloyed types are all of p-n-p configuration, while the grown and diffused junction types are all n-p-n is an inherent result of these methods of fabrication. In the case of the alloyed types, aluminium is the only suitable alloying element so far found, and so must obviously result in a p-n-p structure; while segregation and diffusion coefficients are such that only n-p-n structures can be made by grown-junction or diffusion techniques.

Reliability of Silicon Transistors

Since silicon transistors have only been manufactured in production quantities for less than four years, and moreover since continuous improvements have been made to them during that time, it is inevitable that concrete reliability data over extended periods of operation or storage are conspicuous by their absence. However, there would seem to be no reason whatever why they should be any less reliable than transistors made from germanium, and experience so far seems to show that germanium transistors, provided that they are being operated under conservative ratings, are proving to be exceptionally reliable components—more reliable, indeed, than their associated capacitors and resistors.

Parallel Four-Terminal Networks

TRANSFER VOLTAGE-RATIO

By F. E. Rogers, A.M.I.E.E.*

SUMMARY. *A simple relationship is shown to exist between the output–input voltage-ratio for any number of four-terminal networks in parallel and their parameters of output admittance and short-circuit transfer admittance. The relationship enables the voltage-ratio to be stated without solving the networks. Some examples of its application are included.*

Four-terminal networks, particularly of the T-type, are used in parallel as the equivalents of bridges for electrical measurements, and also as filters. In the first application, the relationships for a null condition only are required ; and these are given readily in terms of the total short-circuit transfer admittance, which is required to be zero at one frequency. For the second application, however, the input–output voltage or current ratio is required over a range of frequencies in order that the discrimination characteristic of the filter may be determined.

The voltage or current output from paralleled networks can be found by reducing each network in turn to an equivalent simple generator by means of Thevenin's theorem. This approach, which has been used by O'Dell[1], is illustrated in Fig. 1 for the case of two networks in parallel.

In Fig. 1(b), Z_{t1} and Z_{t2} denote the output impedances of the networks when E is reduced to zero and terminals

*The Polytechnic, London.

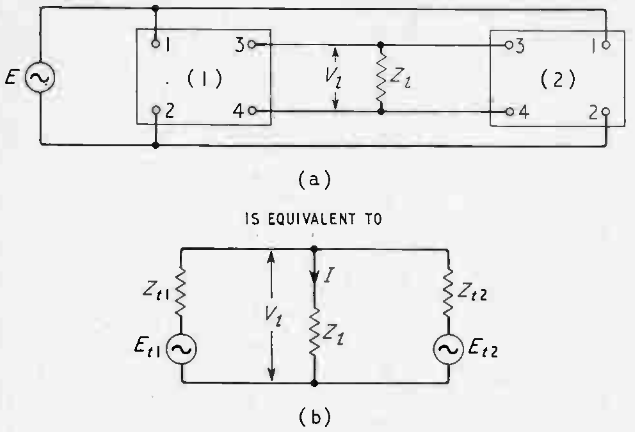


Fig. 1. *(a) Two four-terminal networks in parallel; (b) an equivalent network*

1–2 are short-circuited, whilst E_{t1} and E_{t2} correspond to the output voltages V_{t1} and V_{t2} at terminals 3–4 under open-circuit conditions for each network. The circuit of Fig. 1 (b) is easily solved by means of the superposition theorem, which gives

$$V_l = IZ_l = (I_{(1)}+I_{(2)})\ Z_l$$

$$= \frac{(E_{t2}\ Z_{t1} + E_{t1}\ Z_{t2})\ Z_l}{Z_{t2}\ (Z_{t1} + Z_l) + Z_l\ Z_{t1}} \quad .. \quad (1)$$

where $I_{(1)}$ and $I_{(2)}$ denote the currents in Z_l due to the independent action of E_{t1} and E_{t2} respectively.

A more direct and perhaps more elegant approach is through Millman's theorem, for this permits immediate generalization for the case of n networks in parallel[2]. In this case, the impedance of each equivalent generator is transformed into an admittance as indicated in Fig. 2. According to the theorem,

$$V_l = \frac{E_{t1}\ Y_{t1} + E_{t2}\ Y_{t2} + \ldots .\ E_{tn}\ Y_{tn}}{Y_{t1} + Y_{t2} + \ldots .\ Y_{tn}\ Y_l} \quad .. \quad (2)$$

Neither Equ. (1) nor Equ. (2) gives the transfer voltage-ratio explicitly, however.

Open-Circuit Voltage-Ratio of a Single Four-Terminal Network

It will now be shown that there exists a very simple relationship between the input–output voltage ratio of an unloaded four-terminal network, and its parameters of output admittance Y_t and short-circuit transfer admittance Y_{tr}.

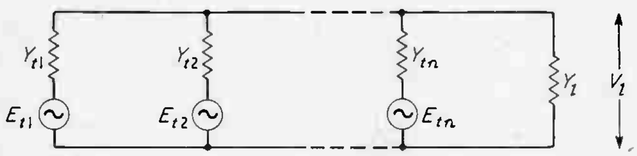
Referring to Fig. 3(a), the short-circuit transfer admittance is defined by

$$Y_{tr} = I/E \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3a)$$

whence

$$I = EY_{tr} \quad .. \quad .. \quad .. \quad .. \quad .. \quad 3(b)$$

Fig. 2. *Equivalent circuit of* n *four-terminal networks in parallel*



In Fig. 3(b) the network is replaced by a simple equivalent generator, in which $E_t = V_t$, where V_t denotes the open-circuit voltage of the network at terminals 3–4, and Y_t denotes the output admittance at these terminals when E is reduced to zero. Thus,

$$I = E_t\ Y_t \quad .. \quad .. \quad .. \quad .. \quad (4a)$$

$$= V_t\ Y_t \quad .. \quad .. \quad .. \quad .. \quad (4b)$$

Comparing Equ. (4b) with Equ. (3b) gives, therefore,

$$\frac{V_t}{E} = \frac{Y_{tr}}{Y_t} \quad .. \quad .. \quad .. \quad .. \quad (5)$$

Since both Y_t and Y_{tr} can generally be stated or easily deduced for a network of given configuration, the open-circuit voltage ratio V_t/E can therefore be stated without solution of the network. While Equ. (5) is of limited importance in relation to a single network, for which the calculation of open-circuit voltage by direct means is likely to be easy, it is of considerable importance when extended to paralleled networks.

Formula for Open-Circuit Voltage Ratio of *n* Parallel Networks

In the case of n four-terminal networks in parallel, driven from a common e.m.f. E, the short-circuit output current is

$$I = E(Y_{tr1} + Y_{tr2} + \ldots .\ Y_{trn}) \quad .. \quad .. \quad (6a)$$

$$= E\,\Sigma\,Y_{tr} \quad .. \quad .. \quad .. \quad .. \quad (6b)$$

The effective output admittance Y_t of a single simple generator equivalent to the n paralleled networks is

$$Y_t = Y_{t1} + Y_{t2} + \ldots .\ Y_{tn} \quad .. \quad .. \quad (7a)$$

$$= \Sigma\,Y_t \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7b)$$

If V_t denotes the open-circuit output voltage of the

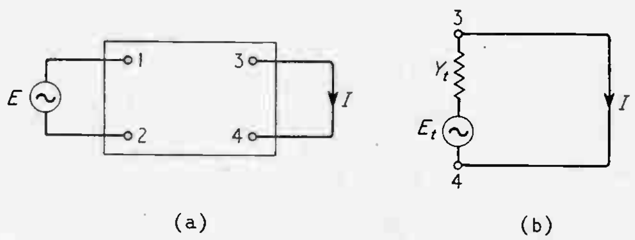


Fig. 3. *(a) General four-terminal network; (b) equivalent used in deriving parameters employed in the article*

paralleled networks, then V_t is also the e.m.f. of the simple equivalent generator, and the short-circuit output current is therefore

$$I = V_t\ \Sigma\ Y_t \quad .. \quad .. \quad .. \quad .. \quad (8)$$

It follows from Equs (6) and (8) that for n four-terminal networks in parallel, the open-circuit voltage-ratio is expressed by

$$\frac{V_t}{E} = \frac{\Sigma\ Y_{tr}}{\Sigma\ Y_t} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Extension of Formula to Include a Common Load

As an artifice, the load may be viewed in the manner of Fig. 4 as a further four-terminal network connected in parallel with the active paralleled-networks.

Obviously, the transfer admittance of Fig. 4 (from the fictitious terminals 1–2 to terminals 3–4) is zero, and the output admittance is $Y_t = Y_l$. Equ. (9) is accordingly extended to give the voltage-ratio V_l/E with reference to a load Y_l, as

$$\frac{V_l}{E} = \frac{\Sigma\, Y_{tr}}{\Sigma\, Y_t + Y_l} \quad \ldots \quad (10)$$

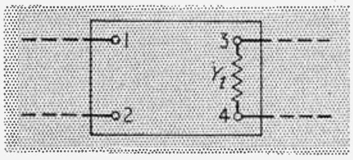


Fig. 4. *Representation of load as a four-terminal network*

The validity of Equ. (10) may be confirmed by applying Norton's theorem. This gives

$$V_l = I/(\Sigma\, Y_t + Y_l) \quad \ldots \quad (11)$$

where I is the output current delivered by the paralleled

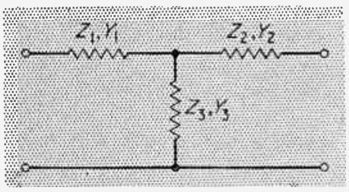


Fig. 5. *T-network*

networks under a short-circuit condition. Substituting for I from Equ. (6b) then gives Equ. (10).

It may now be stated: *the transfer, or output/input, voltage-ratio for any number of four-terminal networks, operating from a common source and into a common load, is equal to the ratio of the total short-circuit transfer admittance to the total output admittance together with that of the load, the output admittance being calculated for each network under short-circuit conditions at the input terminals.*

Illustrations of Application

The short-circuit transfer admittance of any four-terminal network can be calculated according to Fig. 3 and Equ. (3a). In some cases, such as the π-network or a single bridging-arm, it may even be obvious by inspection. In the common case of the T-network, the transfer impedance is well known and expressed, with reference to Fig. 5, by

$$Z_{tr} = Z_1 + Z_2 + Z_1 Z_2/Z_3 \quad \ldots \quad (12)$$

By manipulation of Equ. (12), or otherwise, is obtained the transfer admittance

$$Y_{tr} = Y_1\, Y_2/(Y_1 + Y_2 + Y_3) \quad \ldots \quad (13a)$$

$$= Y_1\, Y_2/\Sigma Y_{123} \quad \ldots \quad (13b)$$

The admittance with terminals 1–2 short-circuited is

$$Y_t = Y_2\,(Y_1 + Y_3)/\Sigma Y_{123} \quad \ldots \quad (14)$$

In practice, the networks are often symmetrical, and Equs. (12)–(14) simplify accordingly.

The theorem will now be used to state explicitly the transfer voltage-ratio of some illustrative configurations, shown in Figs. 6, 7 and 8.

In each case shown in Fig. 6 there are two paralleled networks, and

$$\frac{V_l}{E} = \frac{Y_{tr1} + Y_{tr2}}{Y_{t1} + Y_{t2} + Y_l} \quad \ldots \quad (15)$$

(*a*) *Paralleled T-networks*:—

$$Y_{tr1} = Y_1\, Y_2/\Sigma\, Y_{123}$$
$$Y_{tr2} = Y_4\, Y_5/\Sigma\, Y_{456}$$
$$Y_{t1} = Y_2\,(Y_1 + Y_3)/\Sigma\, Y_{123}$$
$$Y_{t2} = Y_5\,(Y_4 + Y_6)/\Sigma\, Y_{456}$$

Thus,

$$\frac{V_l}{E} = \frac{Y_1 Y_2 \Sigma\, Y_{456} + Y_4 Y_5 \Sigma\, Y_{123}}{Y_2(Y_1 + Y_3)\Sigma Y_{456} + Y_5(Y_4 + Y_6)\Sigma Y_{123} + Y_l \Sigma Y_{123} \cdot \Sigma Y_{456}} \quad \ldots \quad (16)$$

(*b*) *Bridged symmetrical T-network*:—

$$Y_{tr1} = Y_1^2/(2Y_1 + Y_3)$$
$$Y_{tr2} = Y_4$$
$$Y_{t1} = Y_1\,(Y_1 + Y_3)/(2Y_1 + Y_3)$$
$$Y_{t2} = Y_4$$

Thus,

$$\frac{V_l}{E} = \frac{Y_1^2 + Y_4\,(2Y_1 + Y_3)}{Y_1\,(Y_1 + Y_3) + (2Y_1 + Y_3)\,(Y_4 + Y_l)} \quad \ldots \quad (17)$$

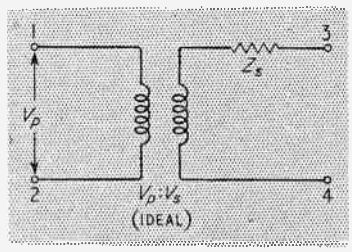


Fig. 7. *Equivalence for transformer with leakage impedances*

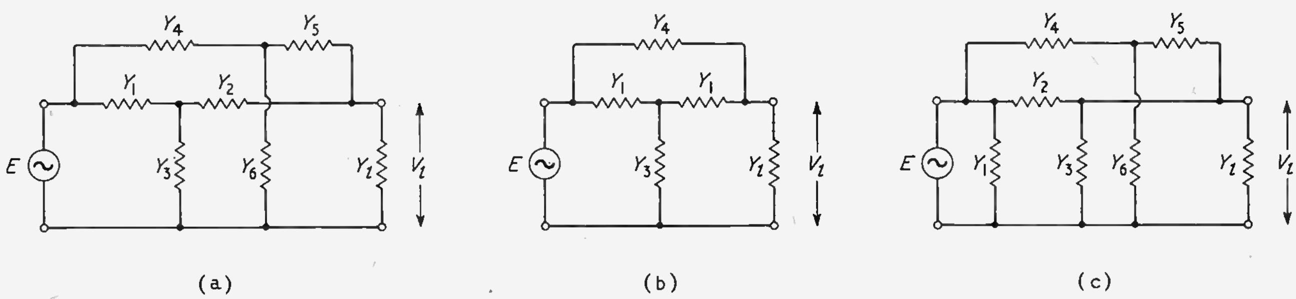


Fig. 6. (*a*) *Twin-T network;* (*b*) *bridged-T network; and* (*c*) *T and π networks in parallel*

(*c*) *Paralleled T and π-networks*

Inspection of the π-network will show that its short-circuit transfer admittance is simply Y_2, and that its output admittance with terminals 1–2 short-circuited is $Y_2 + Y_3$. Thus,

$$Y_{tr1} = Y_2$$

$$Y_{tr2} = Y_4\,Y_5/\Sigma\,Y_{456}$$

$$Y_{t1} = Y_2 + Y_3$$

$$Y_{t2} = Y_5\,(Y_4 + Y_6)/\Sigma\,Y_{456}$$

and therefore

$$\frac{V_l}{E} = \frac{Y_2\,\Sigma\,Y_{456} + Y_4\,Y_5}{Y_5\,(Y_4 + Y_6) + (Y_2 + Y_3 + Y_l)\,\Sigma\,Y_{456}} \qquad (18)$$

It may be noted that Y_1 does not influence the voltage-ratio (as is evident also by inspection of the figure) and it is therefore redundant in this respect.

(*d*) *Transformers in parallel*

In Fig. 7, a transformer is represented by an equivalent circuit, comprising an ideal transformer of voltage-ratio V_s/V_p equal to the open-circuit voltage-ratio of the actual transformer, and an impedance Z_s representing the leakage impedance and winding resistances referred to the secondary.

Referring to Fig. 7, let I_s denote the short-circuit secondary current. Then,

$$Y_{tr} = I_s/V_p$$

$$= \frac{V_s}{Z_s\,V_p} \quad \ldots\ \ldots\ \ldots\ \ldots\ \ldots \quad (19)$$

Under short-circuit conditions at terminals 1–2, the output admittance is

$$Y_t = 1/Z_s \quad \ldots\ \ldots\ \ldots\ \ldots\ \ldots \quad (20)$$

Thus, for *n* transformers operating in parallel from a common supply of voltage V_p and into a common load Y_l,

$$\frac{V_l}{V_p} = \Sigma\,Y_{tr}/[\Sigma\,Y_t + Y_l]$$

$$= \frac{\dfrac{1}{V_p}\left[\dfrac{V_{s1}}{Z_{s1}} + \dfrac{V_{s2}}{Z_{sn}} + \ldots .\dfrac{V_{sn}}{Z_{sn}}\right]}{1/Z_{s1} + 1/Z_{s2} + \ldots .\ 1/Z_{sn} + Y_l} \quad \ldots \quad (21)$$

Equ. (21) corresponds to that which might have been written down almost directly from the mathematical form of Millman's theorem [Equ. (2)] ; but it is interesting to note that it is arrived at independently in this case from a few very simple considerations.

(*e*) *Bridged-Transformer*

In Figs. 8(b) and 8(c) are shown two exact equivalent representations for the circuit of Fig. 8(a). The output-input voltage ratio may be determined from Fig. 8(b) as in the case of the bridged-T network of Fig. 6(b), but Fig. 8(c) provides an alternative and perhaps simpler approach.

In Fig. 8(c) the transformer, assumed non-dissipative, is equivalently represented by an ideal transformer T of impedance transformation ratio $L_1 : k^2L_2$, in association with a shunt inductor k^2L_2 (which accounts for the finite values of L_1 and L_2), and a series inductor $(1-k^2)L_2$ associated with leakage flux.*

For an e.m.f. E applied to terminals 1–2, the secondary voltage of T is $E\sqrt{(k^2L_2/L_1)}$, regardless of the shunt inductor k^2L_2. The short-circuit current traversing terminals 3–4 is therefore

$$I = \frac{Ek\sqrt{L_2/L_1}}{j\omega L_2(1-k^2)} \quad \ldots\ \ldots\ \ldots\ \ldots \quad (22)$$

and the transfer admittance of the transformer is thus

$$Y_{tr1} = \frac{I}{E} = \frac{k\sqrt{L_2/L_1}}{j\omega\,L_2(1-k^2)} \quad \ldots\ \ldots \quad (23)$$

The transformer output admittance at terminals 3–4 under a short-circuit condition at terminals 1–2 is simply

$$Y_{t1} = 1/j\omega\,L_2(1-k^2) \quad \ldots\ \ldots\ \ldots \quad (24)$$

For the bridging capacitor C,

$$Y_{tr2} = Y_{t2} = j\omega C \quad \ldots\ \ldots\ \ldots\ \ldots \quad (25)$$

The transfer voltage-ratio for Fig. 8(a) is therefore

$$\frac{V}{E} = \frac{\Sigma\,Y_{tr}}{\Sigma\,Y_t} = \frac{j\omega C + k\sqrt{\dfrac{L_2}{L_1}}\Big/[j\omega L_2(1-k^2)]}{j\omega C + 1/[j\omega L_2(1-k^2)]}$$

$$= \frac{k\sqrt{\dfrac{L_2}{L_1}} - \omega^2 L_2 C(1-k^2)}{1 - \omega^2 L_2 C(1-k^2)} \quad \ldots \quad (26)$$

The ratio is a maximum (infinity in this assumed non-dissipative case) when

$$1 - \omega^2 L_2 C(1-k^2) = 0$$

or

$$\omega^2 = \omega_\infty{}^2 = \frac{1}{L_2 C(1-k^2)} \quad \ldots\ \ldots\ \ldots \quad (27)$$

and zero when

$$k\sqrt{\frac{L_2}{L_1}} - \omega^2 L_2 C(1-k^2) = 0$$

or

$$\omega^2 = \omega_0{}^2 = \frac{1}{L_2 C(1-k^2)} \cdot k\sqrt{\frac{L_2}{L_1}}$$

$$= \omega_\infty{}^2\,k\sqrt{\frac{L_2}{L_1}} \quad \ldots\ \ldots\ \ldots \quad (28)$$

* See reference 2, pages 413–419

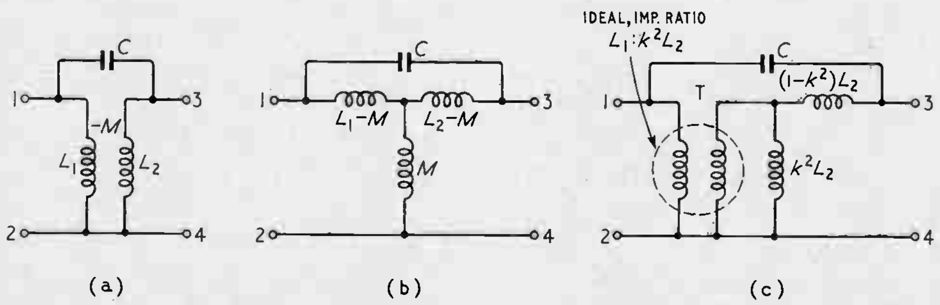


Fig. 8. (*a*) *Bridged transformer;* (*b*) *and* (*c*) *equivalent networks*

From Equ. (28)

$$\frac{\omega_0^2}{\omega_\infty^2} = k\sqrt{\frac{L_2}{L_1}} \quad .. \quad .. \quad .. \quad .. \quad (29a)$$

$$= \frac{M}{\sqrt{L_1 L_2}} \cdot \sqrt{\frac{L_2}{L_1}}$$

$$= \frac{M}{L_1} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (29b)$$

The condition $k\sqrt{(L_2/L_1)} = M/L_1 = 1$ leads to a voltage-ratio of unity at all frequencies. It is evident however from Equ. (29a) that, for a given value of coupling coefficient k, ω_0 can be set above or below ω_∞ by adjustment of the ratio L_2/L_1.

Formulae in Terms of Transmission Parameters

The network may be specified alternatively in terms of its transmission parameters, which comprise an image transfer coefficient Γ and image impedances Z_{i1} and Z_{i2}. For simplicity, attention will be confined to the symmetrical form of network, for which Γ is identifiable with the propagation coefficient, and $Z_{i1} = Z_{i2} = Z_0$, where Z_0 is the characteristic impedance.

By an adaptation of the transmission-line equations, it can be shown that the short-circuit output current from a symmetrical four-terminal network driven from an e.m.f. E is expressed by

$$I = \frac{E}{Z_0 \sinh \Gamma}$$

$$= EY_0 \operatorname{cosech} \Gamma \quad .. \quad .. \quad .. \quad (30)$$

where Y_0 denotes the characteristic admittance.

Thus,

$$Y_{tr} = \frac{I}{E} = Y_0 \operatorname{cosech} \Gamma \quad .. \quad .. \quad .. \quad (31)$$

It can be shown similarly that the output admittance of the network under short-circuit conditions at the input terminals is given by

$$Y_t = Y_0 \coth \Gamma \quad .. \quad .. \quad .. \quad .. \quad (32)$$

Therefore, for n networks in parallel and with a common load Y_l,

$$\frac{V_l}{E} = \frac{Y_{01} \operatorname{cosech} \Gamma_1 + Y_{02} \operatorname{cosech} \Gamma_2 + \ldots Y_{0n} \operatorname{cosech} \Gamma_n}{Y_{01} \coth \Gamma_1 + Y_{02} \coth \Gamma_2 + \ldots Y_{0n} \coth \Gamma_n + Y_l} \quad .. \quad .. \quad (33)$$

Conclusions

It has been shown that a simple general relationship exists between the output–input voltage-ratio for any number of four-terminal networks in parallel, and their parameters of output admittance and short-circuit transfer admittance.

The relationship, which is not restricted to bilateral networks, is thought to be useful, particularly in permitting the voltage-ratio to be stated explicitly and without solution of the networks. It may have notational value in the general analysis and synthesis of networks, and it may also serve as a means for dealing with complicated networks or distribution systems when these can, by inspection, be resolved into parallel arrangements of simpler ones.

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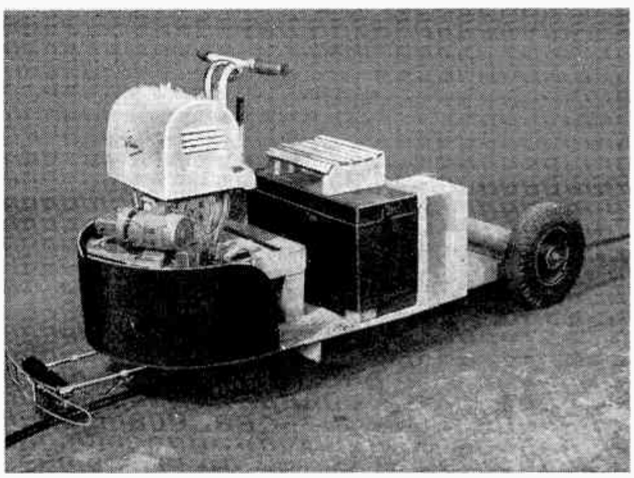
MECHANICAL HANDLING

A method of guiding driverless factory trolleys along predetermined routes was demonstrated at the Mechanical Handling Exhibition. A wire laid along the route is energized by an audio-frequency current, and the resulting induction field is the basis of the guidance system. The trolley carries a pair of pick-up coils and a transistor servo-amplifier which controls the steering so as to keep the nose of the trolley over the guide wire. The system was developed by E.M.I. Electronics.



MANUFACTURERS' LITERATURE

Pullin-Kearfott Servocomponents. Leaflet about size 10 servomotors.
R. B. Pullin & Co. Ltd., Phoenix Works, Great West Road, Brentford, Middx.

Brimar Radio Valve and Teletube Manual No. 7. Pp. 336. Price 6s. Includes information on special valve, rectifiers and transistors, and a selection of circuits for amplifiers, receivers, test instrument, etc.
Standard Telephone & Cables Ltd., Receiver Valve Division, Footscray, Sidcup, Kent.

G.E.C. Valve Manual, Part 1 (Second Edition). Pp. 223. Price 7s. 6d. Covers receiving valves, cathode-ray tubes and semiconductor devices. Tabulated data on obsolete valves.
The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2.

Modern Oscilloscope Practice

PERFORMANCE AND CIRCUITRY

The modern oscilloscope is designed as a measuring instrument. As such, it has the great advantage of making possible the observation of what is being measured, so that the effects of circuit adjustments on waveforms can be studied at first hand. The accuracy with which a good modern oscilloscope measures voltage is not far short of that of most general-purpose meters, and it is by far the most convenient means of measuring short intervals of time.

In this article, only general-purpose oscilloscopes are considered. Special types, with facilities for measuring, say, pressure, or for tracing response curves also exist, but are naturally of less widespread interest. The aim here is to give the potential user an idea of the performance of modern oscilloscopes, with some notes on current practice in circuitry.

It is convenient to consider an oscilloscope as a combination of a Y amplifier, a time-base and a cathode-ray tube, and this is what is done below. It should, however, be remembered that these component elements must be used together, and that the ease of operation of the complete instrument may depend to a large extent on how well the designer manages to combine them. For this reason, due attention must be given to factors such as ease of synchronizing or triggering the time-base and the possible effects of the power-supply system. If, for example, the Y-amplifier controls are directly calibrated in volts per centimetre of deflection so that the oscilloscope can be used with a graticule as a direct-reading voltmeter, then the possible effect of mains fluctuations on measurement accuracy must be considered. An instrument with a stabilized h.t. supply might be expected to be more reliable than one without such a refinement, and one with both h.t. and e.h.t. stabilizers better still.

A general requirement of the utmost importance is that an oscilloscope should add nothing to and take nothing away from the signal to be displayed. It is particularly necessary to avoid hum, harmonic distortion and overshoot, since their presence may give a misleading impression of the nature of the input signal.

Y-Deflection Amplifiers

The designer of a general-purpose oscilloscope has no knowledge of the magnitudes of the signals which its ultimate owner will wish to apply to the Y input. Since, in general, the cost of an amplifier is proportional to both its gain and its bandwidth (with an additional price penalty if d.c. amplification is also required) it is not surprising that Y-amplifier performance is one of the most variable features of oscilloscopes. The sensitivity of commercial instruments ranges from a fraction of a millivolt; e.g. 250 microvolts/cm (Nagard DT103) to several volts per centimetre, while a plug-in amplifier gives certain Tektronix instruments a sensitivity of 50μV/cm. At least one oscilloscope (Marconi TF942) has no Y amplifier at all, the sensitivity then being that of the cathode-ray tube itself (about 50 V/cm).

It is always possible (at the expense of some mechanical complexity) to provide a range of sensitivities from the same amplifier by switching-in different sizes of load or negative-feedback resistors. The bandwidth then becomes smaller as the gain is made larger; the shapes of complex waveforms may then change as the gain setting of the amplifier is altered, because of the varying amplification and phase shift to which their high-frequency components are subject. This difficulty is skirted in some instruments by providing two Y amplifiers, a main amplifier of large bandwidth, which is always in circuit, and a pre-amplifier of smaller bandwidth which can be used when high gain is the important criterion. The main amplifier is often direct-coupled, while the pre-amplifier uses *RC* couplings. This reduces the drift problem which would arise if the whole amplifier were capable of passing d.c. Bandwidth can only be made constant if the high-frequency cut-off is allowed to

E.M.I. WM7 wide-band general-purpose oscilloscope. The Y frequency response is d.c. to 50 Mc/s (— 3 dB) using a plug-in Y amplifier giving 100 mV/cm deflection sensitivity

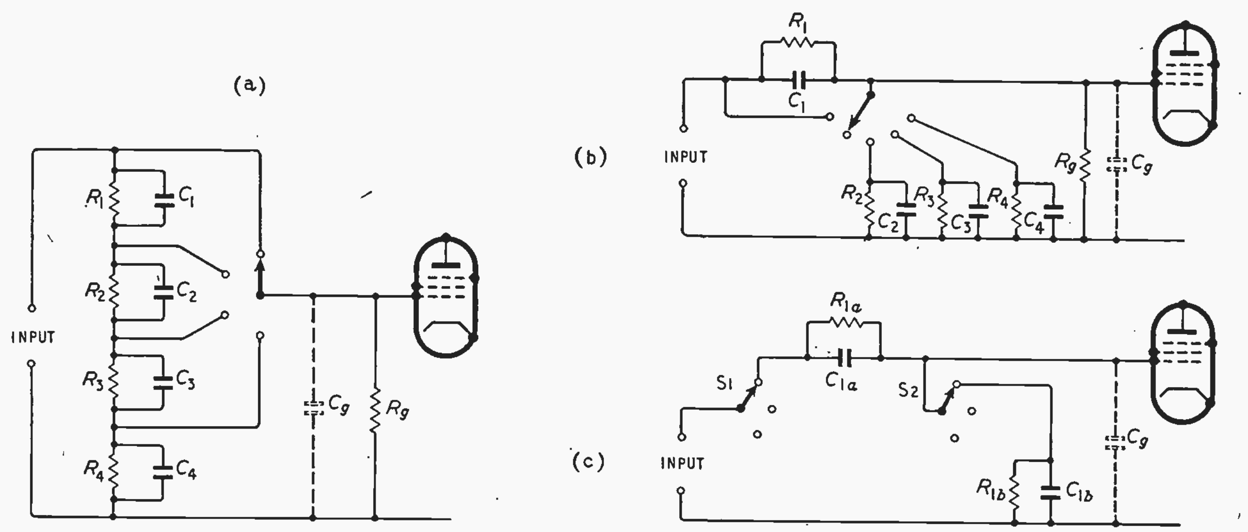
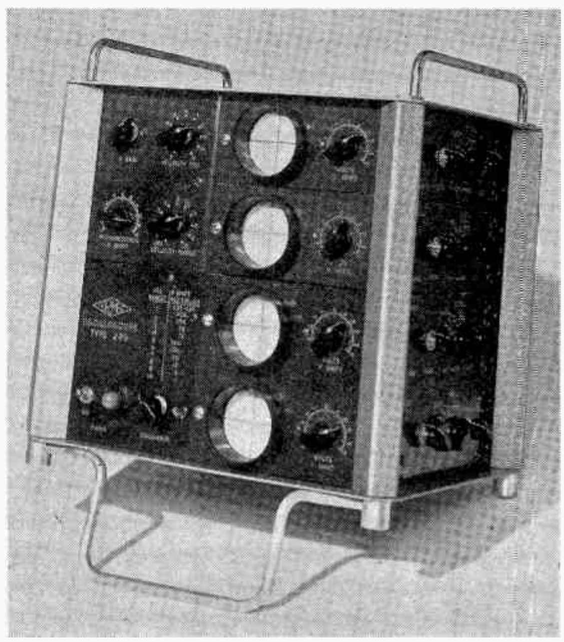


Fig. 1. *Some common attenuator networks. In (c) S_1 and S_2 are ganged*

be determined by the maximum sensitivity condition of the amplifier. The designer must ask himself whether it is better to allow the bandwidth to vary with gain setting, providing the maximum possible bandwidth at the lowest gain setting, or to keep the bandwidth constant, throwing away the possibility of larger bandwidths at lower gain settings. If he thinks the user of the instrument will be misled if wave-shapes change with gain settings, then he may plump for constant bandwidth. On the other hand, he may prefer to offer the user an instrument which will always have a bandwidth of at least, say, 1 Mc/s, but which gives him a bonus in the form of a 10-Mc/s response if he can make do with lower gain. Examples of instruments designed in accordance with these two schools of thought may be cited in the form of two beam-switching double-trace instruments, the Solartron CD. 711 and the Mullard L.101/2. The bandwidth of the former varies from d.c.–7 Mc/s with a sensitivity of 100 mV/cm to 2·5 c/s –200 kc/s (3 mV/cm), while the latter has a constant bandwidth of 5 c/s–4 Mc/s with a maximum sensitivity of 20 mV/cm.

A neat way out of these difficulties, used by Tektronix, is to design a basic oscilloscope unit with a Y output amplifier of large bandwidth, and to provide a selection of Y sub-units, with various gain–frequency characteristics, any one of which can be plugged into the oscilloscope. In this way, the user can purchase whichever Y sub-unit best suits his purposes, with the knowledge that should he at a later time require a different Y amplifier he can continue to use the main part of the oscilloscope, buying only another sub-unit.

The designer's methods of coping with the range of Y-input voltages are rather more standardized. There is always an input attenuator, so that voltages up to, say, 500-V peak can be reduced to manageable proportions at the amplifier input, and there is sometimes another attenuator inside the Y amplifier proper. In all wide-band oscilloscopes, the input attenuator is of the step variety, and consists of *RC* networks designed to give a level frequency response. Actual arrangements vary, and some common ones are shown in Fig. 1. All suffer from the disadvantage that the input impedance is frequency-dependent. The condition for flat frequency response in (a) and (b) is that $R_1C_1 = R_2C_2 = R_3C_3 = R_4C_4 = R_gC_g$, where R_g and C_g are the input resistor and stray capacitance associated with the first valve. These two attenuators have the advantage that a minimum number of components and switch contacts is required. The arrangement of Fig. 1 (b) has the advantage that only two *RC* networks are used at a time, so that errors in the values of the lower networks do not cause cumulative errors in attenuation. Both networks suffer from the disadvantage that the input impedance varies with the attenuator setting. The



Airmec Type 249 *four-channel oscilloscope* arrangement shown at (e), if correctly designed, has a constant input impedance at all settings, but requires more components and contacts. The condition for flat frequency response is that $R_{1a}C_{1a} = R_{1b}\ (C_{1b} + C_g)$, etc.

The input capacitance measured at the Y input terminals often amounts to 20 pF or more. If a cable is used to connect the instrument to the signal source, the cable capacitance must be added to this. The total capacitance is often intolerably large. To reduce it,

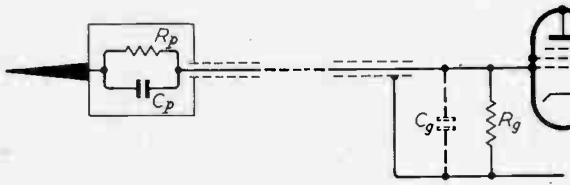


Fig. 2. *Attenuator probe*

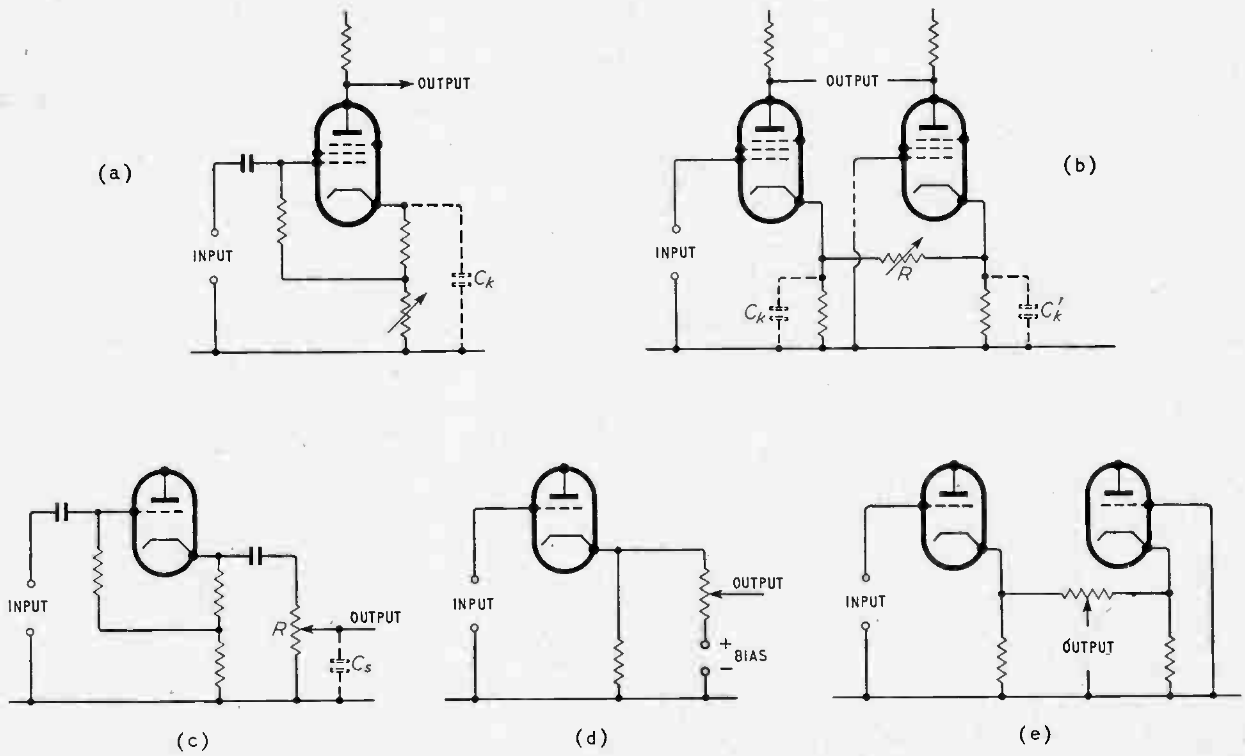
some form of probe may be used. A parallel *RC* network located at the signal-source end of the cable is generally employed. A typical arrangement is shown in Fig. 2. The probe contains a parallel *RC* network R_pC_p and the usual considerations with regard to frequency response apply, the cable capacitance now forming part of C_g. The probe input capacitance can be made as small as necessary at the expense of attenuating the signal transferred to the Y amplifier. Cathode-follower probes, in which the valve is mounted in a holder at the signal end of the cable are occasionally used and have the advantage of negligible attenuation. However, it is questionable whether there is much to be gained from the increased complexity, since the addition of one valve to the amplifier proper would offset the attenuation of an *RC* probe, and the latter is much simpler and more compact. The use in a probe of a high-frequency transistor in the common-collector circuit would appear to offer the possibility of combining some of the advantages of both arrangements.

If step attenuators are included in the Y amplifier, they may take any of the forms of Fig. 1. In addition, the gain can be controlled in steps by variable negative feedback. One circuit which has been used is the 'anode follower', the gain being controlled by switching anode–grid feedback networks.

Continuously-variable control of sensitivity is often possible inside the amplifier. Some arrangements are shown in Fig. 3. The first is only usable in a.c. amplifiers. Increasing the cathode resistor increases negative feedback and reduces the gain. The screen voltage is also reduced, with the same result. Unfortunately, the frequency response varies with the gain setting and, at low gains, may become peaked at the high-frequency end of the passband by virtue of the stray capacitance C_k which reduces feedback at high frequencies.

The arrangement at (b) can be used down to d.c. but, when R is large, the outputs are unbalanced and C_k, C_k' may cause frequency-response errors. The cathode-follower (c) enables a low-resistance potentiometer to be employed, reducing the effect of C_s and, in the form shown at (d), can be used at d.c. In (e), the necessary bias voltage is provided by the drop across the cathode resistor of a similar valve. (This arrangement is

Fig. 3. *Continuously-variable gain controls for oscilloscope deflection amplifiers*



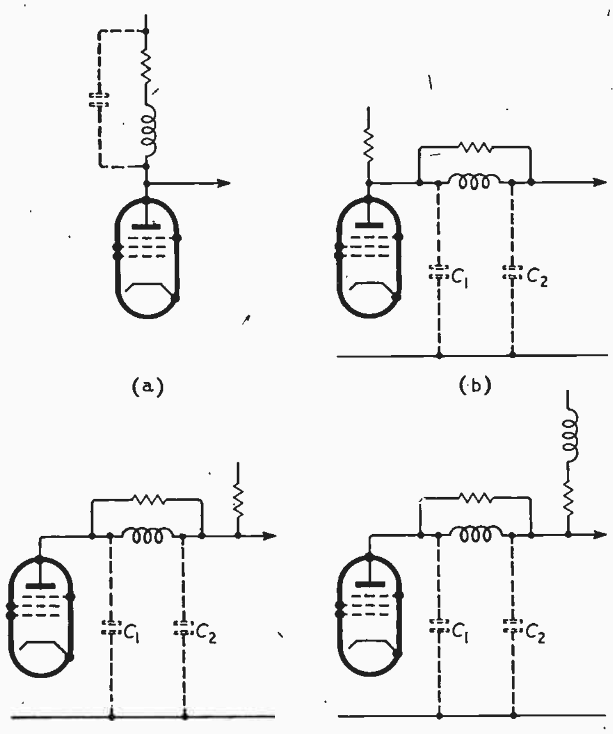


Fig. 4. *Video-coupling networks*

employed in the Röhde & Schwarz oscilloscope type OBF.)

Although continuously-variable sensitivity controls are of use when it is required to keep the size of the displayed waveform constant, fit it to a graticule, etc., not all oscilloscopes incorporate this type of control. One reason is that continuously-variable gain is not compatible with precise voltage measurement. If the accuracy of measurement depends on the amplifier gain, errors in setting such a control will contribute to the total error and this can easily happen if the pointer of the control is not quite aligned with a scale marking. Fortunately, some measuring systems are independent of amplifier gain, provided that the latter does not change while a measurement is made.

The circuitry of the Y amplifier usually follows conventional lines. In general, there is no negative feedback except over single stages, and peaking coils are employed to extend the high-frequency response. Shunt-peaking coils [Fig. 4 (a)] are often employed in the low-level stages, and series-peaking coils [Fig. 4 (b)] in the output stage. It often happens, however, that C_2 is smaller than C_1, in which case the load resistor is placed at the opposite end of the series coil (c). Combination shunt–series peaking is also used, as shown at (d).

A somewhat surprising feature of certain oscilloscopes is the use of triodes in the video stages. The Cossor 1045K (which is sold as a kit of parts and printed circuits to be assembled by the purchaser) uses a 6BQ7A double-triode, with the two sections in cascade, driving a 12BH7 double-triode cathode-coupled output stage (Fig. 5). The high-frequency response is specified as 3 Mc/s (−3 dB). It will be noted that the anode loads (and therefore the stage gains) are low. In addition, the 6BQ7A has a small grid–anode capacitance (1·15 pF). The combination of these two factors no doubt results in a comparatively small Miller-effect component of input capacitance.

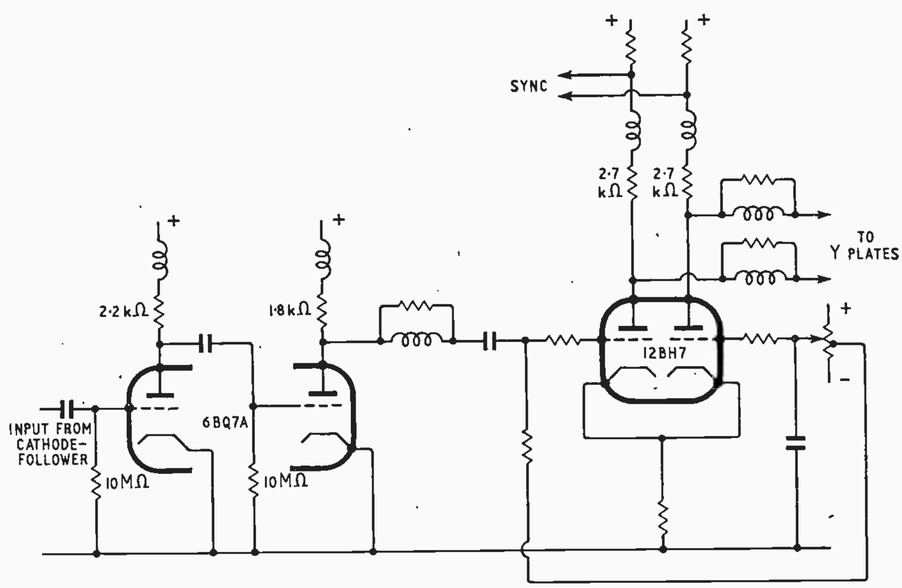
Some other unusual Y-amplifier circuits are a 'ring-of-three' arrangement (two cascaded pentode amplifiers and an output cathode follower, with overall negative feedback), used in the Radar 301 oscilloscope to give a response to 6 Mc/s (−3 dB) with a sensitivity of 100 mV/cm, and a type of stacked valve amplifier (Fig. 6), which is used in the Solartron CD518 and CD568. The advantage claimed for the latter is that, since the two valves are in parallel as far as the signal is concerned, high peak currents can be delivered without undue drain on the h.t. supply.

Some pulse oscilloscopes have a delay line in the Y amplifier. The synchronizing signal is taken from a point before the delay line, so that the time-base is triggered in advance of the Y deflection. The leading edge of a pulse can then be displayed in its entirety. The duration of the delay is generally less than 1 μsec.

Voltage Measurement

It is always possible to calibrate an oscilloscope by applying a known alternating voltage to the Y-amplifier input and noting the deflection obtained. This system is

Fig. 5. *Part of Y amplifier of Cossor 1045K*



employed in some modern oscilloscopes, and is the only one possible if the Y-amplifier response does not extend to d.c. The calibrating reference voltage can be measured, but it is more convenient to have a stable voltage (obtained from a source such as a lamp bridge) permanently available. A square waveform is convenient, since it is easy to align the flat tops with horizontal graticule markings.

A calibrator found in some Tektronix oscilloscopes is shown in a slightly simplified form in Fig. 7. V_1 and V_2 form a 1-kc/s multivibrator, the screen circuit of V_2 being coupled to the grid of V_1. Square waves of unity mark-space ratio are generated, so the anode current

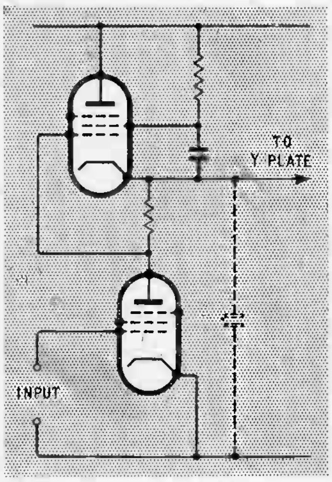


Fig. 6. *Half of a stacked-valve Y output stage (Solartron)*

of V_2 takes the form of square pulses, the valve being cut-off for half the time and switched fully on for half the time. When V_2 is 'off', the potential at the grid of the output cathode follower is determined by the setting of *R* and, when V_2 is 'on', V_3 is cut-off, so that its output voltage is zero.

The value of the output voltage when V_2 is off is the sum of the potential at the slider of *R* and the grid bias of V_3. The setting of *R* is adjusted, with V_2 out of circuit, so that the cathode voltage of V_3 is exactly 100 V. This voltage is determined largely by the h.t. supply voltage, which is stabilized, the only material cause of drift being variations in the working grid bias of V_3. The latter is a triode with a short grid base and is operated well within its ratings. The overall measuring error, which includes any errors in the potential dividers, is specified as less than 3%.

Provided that an oscilloscope is calibrated immediately before use by some such method, so that the effects of changes in amplifier gain with time can be minimized, accurate measurements are possible, but there are at least two sources of error. The first is non-linearity in either the deflection amplifier or the tube. If such non-linearity is present, a calibration carried out in terms of trace deflection on the face of the tube is only strictly valid for deflections of the same amplitude as that of the calibrating waveform. The second source of error is the Y-amplifier frequency response. Unless the calibrating voltage is of the same frequency as the measured voltage, errors are possible.

The first effect is probably much more serious. Few oscilloscope specifications include a linearity figure, and those which do generally quote something between 2% and 5%. Since the instruments for which a figure is quoted are rather better than the average, the errors in linearity to be expected may often be rather larger.

Fortunately, most modern oscilloscopes have direct-coupled Y amplifiers. It is then possible to eliminate the first of the errors mentioned above. The arrangement used is shown in simplified form in Fig. 8. A Y-shift voltage of known magnitude is applied to the Y-amplifier input along with the signal. The distance by which the displayed waveform is shifted on the screen then corresponds to a particular shift voltage, which can be measured on a d.c. meter.

The method of making a measurement is illustrated in Fig. 8 (b). It is required, say, to measure the amplitude of the positive peak of the sine-wave shown on the left. This wave is centred on a base-line A, which is shifted to B (coincident with the original position of the peak) by the Y shift control. The amount of d.c. voltage needed to do this is the same as the peak voltage of the wave. Since the d.c. shift potential is subject to the same non-linearities as the a.c. input, the latter do not affect the accuracy. The display is, in fact, being used merely as a null indicator to show that the zero line at B corre-

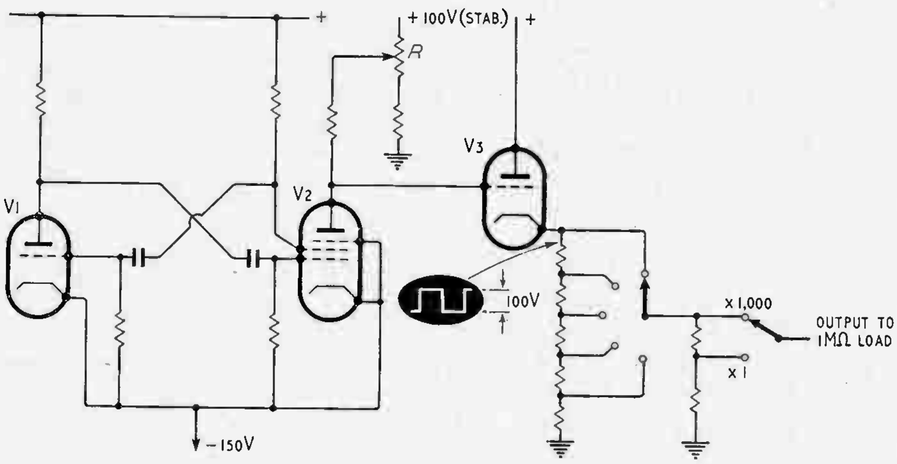
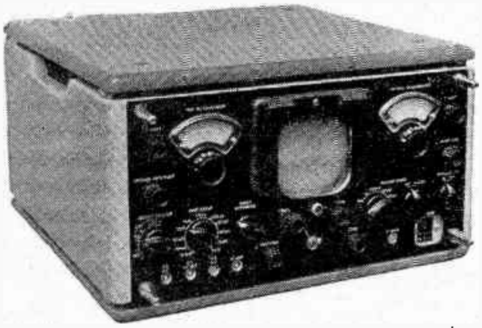


Fig. 7. *Method of generating square-wave calibrating voltage (Tektronix)*



Mullard precision oscilloscope L.140. Time and voltage are measurable within 3%. The Y bandwidth is d.c. to 7·5 Mc/s (−3 dB)

sponds with the original level of a selected portion of the waveform.

There is still a possible source of error, however. The combined signal and shift voltages may overload the Y amplifier, causing a valve to run into grid current. This may have the same effect as an alteration of the Y-shift voltage, so that the actual Y-shift voltage is no longer an accurate indication of the signal amplitude. The effect may occur when large signals are being measured but it can be eliminated, at the expense of some extra trouble, by proceeding as follows. The position on the graticule of the part of the signal to be measured is noted, the signal is switched off and the Y shift operated to transfer the base line to the point of interest. In this way, the amplifier need only be able to handle the shift voltage.

The remaining sources of error are the Y frequency response, the voltmeter calibration, and the accuracy of the potential divider which attenuates the shift voltage before the latter is applied to the amplifier. Most oscilloscopes which use this system have voltage-measurement accuracies of 5% or better, while one (Newport Instruments Precision Pulse Oscilloscope) can be as accurate as 0·1% under certain conditions. Even the worst of these accuracies is about three times better than that obtained when amplifier gain stability and tube linearity are relied on; i.e., the tube graticule is used as a voltage scale, without recourse to a calibration voltage. However, by regulating h.t. supplies and stabilizing amplifier gain by negative feedback, the latter system can be improved; an accuracy of 10% is quoted for one commercial instrument (Solartron CD523S).

One of the difficulties about voltage measurement by graticule only is that changes in deflection sensitivity are brought about by changes in e.h.t. voltage. Unless the e.h.t. voltage is stabilized, mains variations limit accuracy.

If, however, the e.h.t. voltage is stable and capable of adjustment, a means is provided of controlling the deflection sensitivity without doing anything to the Y amplifier. This technique is adopted in the E.M.I. WM5A oscilloscope; the e.h.t. is variable from 1 to 10 kV.

Time-Bases

All commercial oscilloscopes use capacitive time-bases; i.e., the sweep voltage is the voltage across a capacitor which is being charged or discharged through a resistor or equivalent device. It is necessary to restore the state of charge of the time-base capacitor to some initial condition (e.g., zero charge) after each sweep. Thyratrons are no longer used for this purpose, and this is one reason why time-base circuits have become more complicated; a thyratron switches itself on when its striking voltage is reached, but a hard-valve discharge device must be switched on by an external signal.

If the time-base capacitor is charged through a physical resistor, considerations of linearity of sweep require that the capacitor should only be charged to a small fraction of the supply voltage. (In one instrument, the voltage across *C*, which is charged from the h.t. supply, is limited to some 10 V peak-to-peak.) With all practical supply voltages, amplification of the capacitor voltage is necessary in these circumstances. In the Röhde & Schwarz OBF oscilloscope, two voltage-amplifying stages are used after the time-base. A triode

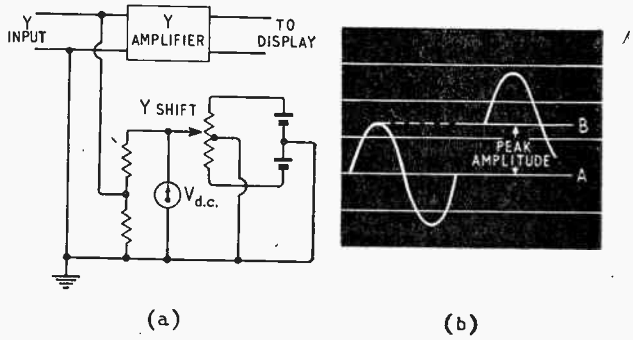
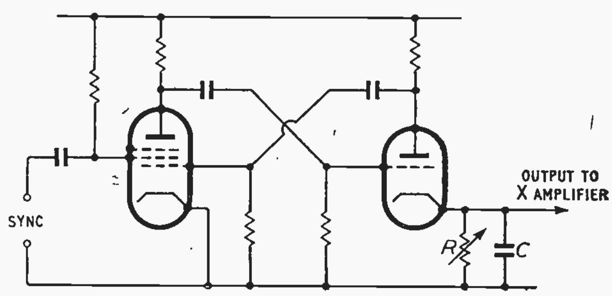


Fig. 8. *Voltage-measurement technique for direct-coupled oscilloscopes; (a) essentials of measuring circuit, (b) measurement of peak amplitude*

is connected across the capacitor to act as a discharge device. It is triggered by pulses from a special multivibrator which, in turn, is synchronized by the signal. In some simpler oscilloscopes, such as the Labgear and the Cossor 1045K, an *RC* timing network is combined with a multivibrator type of relaxation oscillator, as in Fig. 9.

The majority of oscilloscopes use some form of sweep-linearizing device, such as a 'constant-current' charging valve, a bootstrap feedback circuit or a Miller integrator. The last is the most popular, perhaps because it combines good linearity with the ability to work over an enormous range of sweep velocities (e.g., 0·1 μsec/cm to 12 sec/cm in the Tektronix Type 531).

Fig. 9. *Multivibrator time-base*



Nagard Type 301 *wide-band (d.c. to* 40 *Mc/s) oscilloscope*

A disadvantage of the simple Miller circuit is its long flyback time; the timing capacitor charges and discharges through high-resistance circuits. Modified circuits containing a low-impedance charging device, such as a cathode-follower, are generally used.

The Miller integrator itself merely produces a sweep voltage in response to a trigger pulse but, in an oscilloscope, means must usually be provided for making the time-base free-running. To achieve this, some form of regenerative feedback circuit is required. The most economical circuit is the Miller transitron, in which the Miller valve functions both as an integrator and a transitron oscillator.

The conventional circuit is shown in Fig. 10. The *RC* network in the screen and suppressor-grid circuits forms the transitron oscillator which switches the anode current on and off by means of the suppressor grid. When the anode current is switched on, the sweep takes place, its duration being determined by C_2, R_1 and the potentiometer setting. When the anode voltage falls to such a low value that the screen current increases, transitron action takes place and the anode current is abruptly cut-off. It remains cut-off for a period determined by the screen and suppressor-grid circuit component values. This is the flyback time and must be long enough for C_2 to re-charge. On the other hand, the flyback time should preferably not be longer than, say, a tenth of the sweep time. If, therefore, the sweep time is variable, it is necessary to vary the transitron period to keep the sweep/flyback ratio reasonably constant. In practice, C_1 and C_2 are both changed when the coarse sweep-frequency control is operated.

An alternative, which avoids the necessity for two sets of capacitors, is to use a special fast flyback circuit. A typical circuit embodying this and other refinements is used in the Cossor 1058 oscilloscope, and given in simplified form in Fig. 11. The circuit voltages on the diagram refer to the quiescent state. The anode potential of V_4 is then the same as that of V_1, because the diode V_2 conducts and connects the two anodes together. The timing capacitor *C* is charged to approximately this potential by the anode current of the cathode-follower V_5. (The left-hand plate of *C* is clamped at the cathode potential of V_4 by grid current in V_4.) Trigger pulses cause the anode voltage of V_1 to fall considerably. Since V_2 continues to conduct, the grid of V_5 is driven negatively and so is that of V_4, via *C*. *C* therefore discharges through *R*, grid current in V_4 having ceased, the discharge being linearized by Miller feedback via V_5. When the anode voltage of V_4 falls to the same value as the voltage at the slider of R_1, V_3 conducts and no further fall can take place. The setting of R_1 thus determines the sweep amplitude because, once V_3 conducts, Miller feedback ceases, the grid of V_4 is driven positive by the h.t. voltage via *R*, and flyback can take place.

The charging time constant of the circuit containing *C* now depends on the output resistance of the cathode follower V_5 and the input resistance of V_4 under grid-current conditions. Since both of these resistances are small, flyback is always rapid, and a small fixed time constant in the screen–suppressor circuit of V_4 is adequate.

The time-base can be made free-running by reducing the suppressor-grid bias of V_4.

Sync Amplifiers

A feature of many modern oscilloscopes is the comparative complexity of the synchronizing or triggering circuits. Both amplification and pulse-shaping are frequently provided. One reason is that most of the triggered time-bases require a triggering pulse of approximately square waveform with a well-defined duration. Another is that, to make sure that the time-base is rigidly locked, it is advisable to generate sync pulses with short rise-times. If the signal is a sine-wave, it may be necessary to turn it into a square wave by means of a Schmitt trigger circuit or amplifier-limiter in order to produce suitable sync pulses.

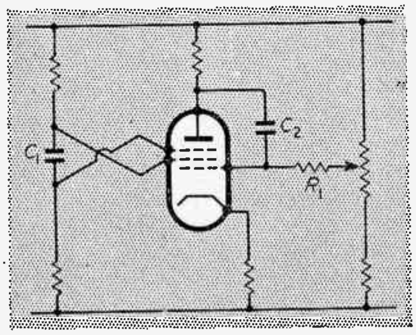
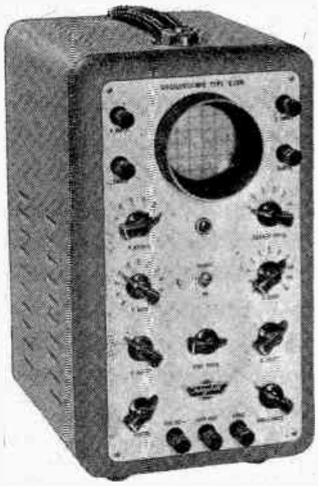


Fig. 10. *Miller-transitron time-base circuit*

It is possible, when the signal to be displayed is also used to synchronize the time-base, to pick off a sync voltage from the Y-amplifier output. This has two advantages: first, less sync amplification is then required and, secondly, if the Y amplifier is push-pull, a sync of either polarity can be readily obtained as required.

However, it is sometimes necessary to use a separate sync signal, in which case sync cannot be obtained in this way. If there is a delay line in the Y amplifier so that the leading edge of a wave can be examined, it is likewise impossible to utilize the full Y gain in the sync path.

Some oscilloscopes incorporate a sync gating circuit which makes a separate amplifier desirable. Suppose it is required to synchronize a time-base of frequency *f* with



This lightweight oscilloscope, the Furzehill O.120, weighs only 18 *lb. The Y response is* 1·5 *c/s to* 100 *kc/s* (−3 *dB*)

a signal 10*f*. It is quite possible that the time-base will lock sometimes to every ninth or eleventh cycle, since the ninth, tenth and eleventh signal cycles all occur at approximately the end of the sweep. To avoid this in timebases in which the sync signal initiates the sweep, the sweep waveform is applied to a gate valve which is so biased that it remains cut off until after the finish of the sweep. The sync signals must pass through the gate to get to the time-base. If the 'open time' of the gate is made short enough, only the required sync signals can pass. A pentode with the sync applied to the control grid and the sweep voltage to the suppressor grid forms a suitable gating valve. Such a device is employed in the Solartron CD518/568 oscilloscopes, in conjunction with a pulse transformer which produces differentiated output pulses (Fig. 12).

A different kind of gate circuit is used in the Telequipment 'Serviscope' to enable sync pulses to be generated from selected portions of the signal waveform. This facility is useful if a detailed examination of a particular portion of a repetitive signal has to be made and is similar in purpose to the expanded sweep and delayed sweep circuits described later in this article. A biased trigger circuit is employed and the bias is adjusted manually so that triggering takes place at a particular input voltage; i.e., a particular phase of the input signal. The resulting pulse initiates the sweep, the duration of which can be adjusted so that the required amount of expansion is obtained.

Time Measurement

The stability and linearity which can be obtained from sweep generators employing Miller-integrator and other feedback linearizing circuits make it a practical proposition to calibrate the time-base controls directly. Accuracy of time measurement is frequently claimed to be better than 5% when this system is used.

Some oscilloscopes incorporate time-calibrating circuits. Calibration can be effected either by displaying a waveform of known period or by using such a waveform to produce brightness modulation. The best frequency standard is a quartz crystal, but it is not easy

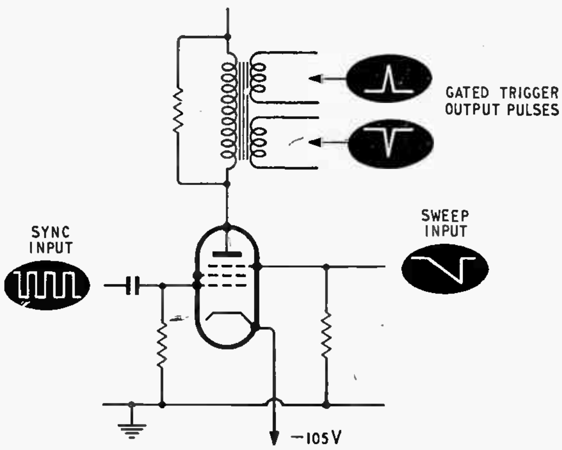


Fig. 12. *Gated sync circuit* (*Solartron*)

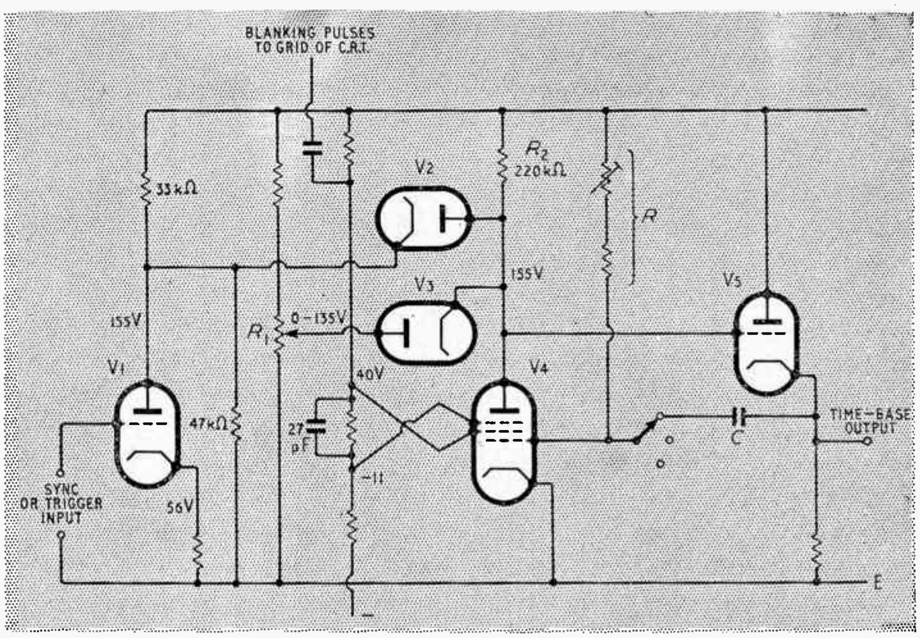


Fig. 11. *Complete Miller-transitron circuit.* (*Cossor*)

to use a crystal oscillator for time measurement. The reason is that, to produce a stable marker display, the phase of the calibrating wave must be the same at the start of each sweep. The usual way of arranging this is to quench the oscillation during flyback and start it up again at the beginning of the sweep but, unfortunately, it is difficult to make a crystal oscillator with good starting characteristics. It might, of course, be possible to synchronize the sweep by means of a free-running oscillator and so obtain stable time-markers, but it is generally required to synchronize the time-base from the Y signal. For this reason, *LC* oscillators or passive ringing circuits are used instead of crystal oscillators.

The commonest time-marker generator is the ringing tuned circuit of Fig. 13. The valve normally conducts, so that the valve cathode impedance is across the tuned circuit, which is heavily damped. The valve is switched off by pulses which bear a fixed phase relation to the sweep waveform, and are usually coincident with it. The energy stored in the field of the inductor causes the tuned circuit to ring, starting with a positive half-cycle. At the end of the pulse, the valve conducts again, damping the oscillation.

The ringing tuned circuit must be of high Q and lightly loaded to avoid excessive diminution of amplitude during the scan. This means that the output voltage must be applied to a valve grid circuit. If the time-markers are to be applied to the Y plates, the Y amplifier is used. If brightness modulation is required, an extra valve is necessary.

It is possible to connect the additional valve as an oscillator, the ringing circuit forming the oscillator tank. The time-marker generator is then a triggered oscillator but, in other respects, operates as before, with the advantage that the amplitude of oscillation does not diminish during the sweep. In either case, the accuracy of time calibration can be no better than the frequency stability of the tuned circuit, which may be 1–5%. A number of tuned circuits may be provided and the appropriate one selected by means of a switch. The frequencies are often 10 kc/s, 100 kc/s and 1 Mc/s, giving 100-μsec, 10-μsec and 1-μsec time marks.

A free-running crystal oscillator can be used to provide time-markers if the time base is of the one-shot variety, since there is then no possibility of time-marks falling in different positions on successive traces to produce a confusing pattern. In general, one-shot sweeps are only useful for examining transients in conjunction with either a long-persistence screen or a camera.

Expanded Trace Working

Facilities for the detailed examination of a small part of an input waveform are provided in many oscilloscopes. This requires stable sweep circuits, since it amounts to sampling the signal for a short interval once every cycle. If the period of the signal is long, and the duration of the sample small, then the accuracy of the necessary timing operation must be high. An extreme case is the selection of one line from each complete television picture. This requires the selection of a sample lasting about 100 μsec once every 40,000 μsec (in a 25-c/s interlaced system). If a 'jitter' of 5% is permissible, a stability of 1 part in 8,000 is necessary. This order of accuracy is beyond all but the best general-purpose oscilloscopes, but it indicates what is desirable in practice.

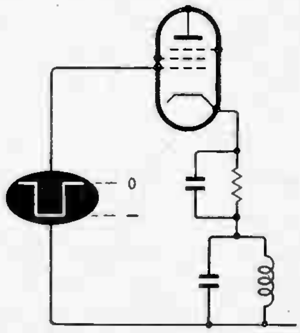


Fig. 13. *Ringing tuned circuit*

Two means of obtaining trace expansion are commonly employed in oscilloscopes. The first (Fig. 14) consists of delaying the start of the sweep until just before the interesting part of the signal and making the sweep duration just long enough to exhibit the interesting part. The time-base must be capable of triggered operation.

Delay lines could be employed but, although they are accurate and stable, they are too expensive and bulky except for very short-duration delays. Some form of time-base circuit is used instead.

The start of a saw-tooth is initiated by the signal and its flyback provides a delayed output pulse for triggering the sweep generator proper. A form of Miller time-base is frequently employed as the delay circuit. Such a device is reasonably stable, cheap, and capable of a very wide range of delays. The delay can readily be made continuously variable. The cathode-coupled phantastron is a suitable delay device and a version of it is employed in the Solartron CD518/568 for this purpose. In the conventional circuit, the anode current of a pentode valve is normally cut off by a suppressor-grid bias, as in the ordinary Miller-transitron time-base. The source of suppressor-grid bias is, however, the voltage drop across a high value cathode resistor. The run-down is initiated by applying a negative pulse to the grid, thereby driving the suppressor positive and switching on the anode current.

The essentials of the circuit are shown in Fig. 15. It will be seen that there are two diodes (V_1 and V_3) in addition to the pentode Miller valve. The purpose of V_3 is to determine the sweep duration, and that of V_1 to isolate the input circuit, once run-down has begun. In the quiescent condition, V_1 is conducting and the grid V_2 is at a positive potential with respect to earth. The suppressor grid is returned to a lower positive potential such that its net bias is sufficiently negative to cut off anode current. The anode is clamped at a high positive potential by V_3. The arrival of a negative trigger pulse causes the anode current to be switched on as described

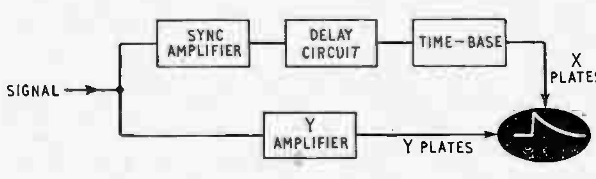


Fig. 14. *Delayed sweep circuit*

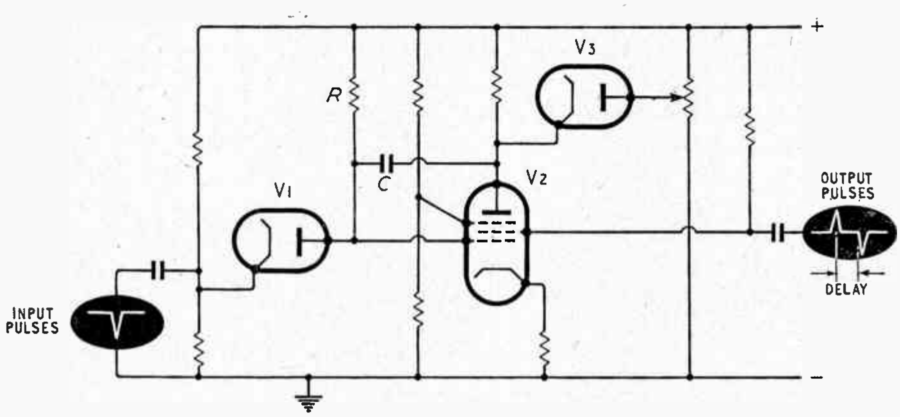


Fig. 15. *Cathode-coupled phantastron delay circuit*

above, and the resulting fall in anode voltage is transferred to the grid via *C*. V_1 is now cut off and remains in this condition until the end of the run-down. This is necessary to prevent the charging resistor *R* from being shunted by the forward resistance of V_1, but it also prevents the arrival of further trigger pulses during the run-down from upsetting the circuit. The circuit can thus be arranged to trigger to every *n*th input pulse, a facility of value when pulse-trains are to be examined. *C* now discharges via *R* and Miller action takes place, the anode voltage falling and the grid voltage rising a little. When the anode voltage becomes such that V_3 conducts, the sweep ends. No further Miller feedback can take place and the grid is driven more rapidly positive via *R*. The resulting drop across the cathode resistor biases the suppressor negative and all the cathode current is diverted to the screen grid, producing a negative output pulse. *C* is switched to provide a coarse delay (actually run-down speed) control, fine control of delay being obtained by adjusting the cathode voltage of V_3 and so fixing the sweep duration.

The delayed output pulses are employed to trigger the time-base proper. 'Sweep expansion' is then usually effected by increasing the sweep velocity while keeping the length of the trace on the cathode-ray tube constant. Assuming that the sweep starts from the same voltage every time, the effect of increasing the sweep velocity will be to spread out the trace in one direction; i.e., from one edge of the c.r. tube screen. In order to put other points of the signal waveform on the screen, the delay time must be altered.

This type of expanded trace operation is best suited to handling pulses spaced well apart in time, because the flyback of the phantastron delay circuit is slow. It is frequently used in radar, but seems to be falling out of favour among oscilloscope designers. It has the great advantage that very large effective sweep expansions can be obtained, the amount depending only on the stability of the delay circuit.

The second possible way of obtaining trace expansion is simply to increase the sweep velocity of a conventional time-base by applying an increased sweep voltage to the X amplifier. The effect can be seen from Fig. 16. A saw-tooth voltage wave of maximum amplitude V_1 is just sufficient to produce a deflection of one screen diameter in a time t_3. A similar saw-tooth wave of amplitude V_2 causes the screen to be swept in time t_2, and one of amplitude V_3 in time t_1. The X amplifier need only be

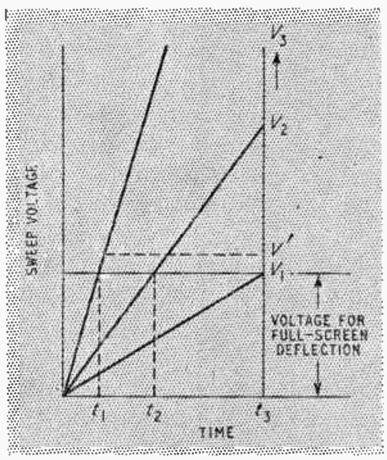
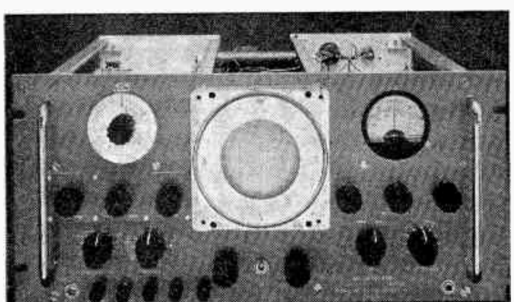


Fig. 16. *Effect on trace duration of increasing saw-tooth amplitude*

able to supply enough output voltage to produce full-screen deflection. Overloading may occur at some output level such as *V′*. A possible amplifier arrangement is shown in Fig. 17. Direct coupling is employed, so that the input to the amplifier is always negative with respect to the cathode of the first valve of the long-tailed pair. If overloading occurs, the anode current of this valve is cut off. If the sweep always starts from a fixed voltage (which is usually the case) increasing the X gain causes the trace to be expanded in one direction only; i.e., from one edge of the c.r. tube screen. In some



*Marconi Instruments rack-mounting oscilloscope TF*1153

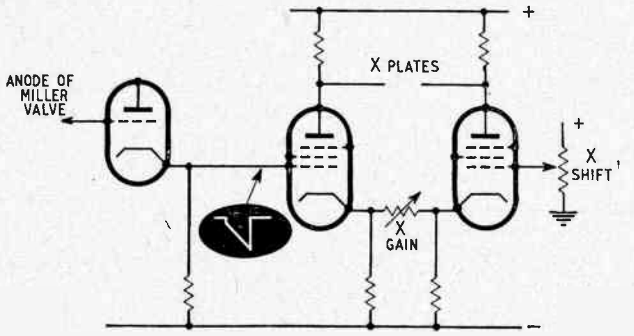


Fig. 17. *Typical X amplifier*

oscilloscopes, the X output stage is driven in a true push-pull manner, and the trace then expands about the centre of the screen. In either case, the part of the Y signal which is of interest can be put in a convenient position by means of the X shift control, provided that too much shift is not required.

The form of trace expansion just described has the advantage over the delayed triggered sweep system that no extra circuitry is required, but only a limited amount of expansion is possible.

The X–Y Oscilloscope

For the purpose of phase measurement and polar co-ordinate displays, it is desirable to have equal X and Y sensitivities and phase–frequency characteristics. At the same time, a sweep generator is not necessary, though, of course, it is still useful to have one available so that the oscilloscope can be used for normal displays.

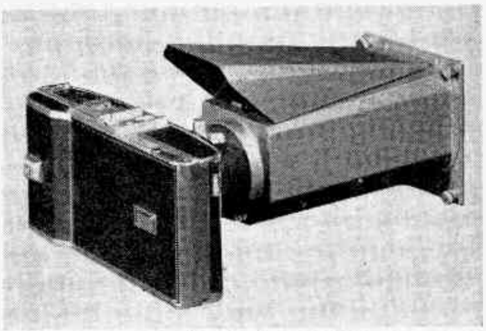
Oscilloscopes with identical X and Y characteristics are now beginning to appear. The Hewlett-Packard 130A, for instance, is a sensitive (1 mV/cm) direct-coupled oscilloscope with a 300-kc/s bandwidth and low drift (1 mV per hour). Tektronix have combined the principle with their system of plug-in units to provide a versatile instrument (Type 536) which can have identical X and Y characteristics, dissimilar characteristics, or a conventional arrangement with a saw-tooth sweep generator. With certain plug-in pre-amplifiers, the relative phase-shift is specified as within 1 degree up to 15 Mc/s.

Multiple Trace Displays

A number of arrangements have been used for displaying more than one waveform on a single cathode-ray tube. In the Cossor split-beam system, a special tube is employed in which the beam, after X deflection, is split into two parts, each of which is subject to the influence of one Y plate. Multiple-gun tubes are made in which a number of complete electrode assemblies are housed in the same glass envelope. In a third type of tube two guns are used, but the X plates are common to both beams. In effect, they are quite separate c.r.ts, except that they share the same screen.

An alternative approach is to use a simple one-gun tube and arrange the associated circuitry so as to produce more than one trace. There are several ways of doing this. One is to employ an electronic switch to connect separate Y inputs alternately to the Y plates, at the same time changing the Y shift voltage in order to separate the traces. A block diagram of the arrangement is given in Fig. 18. Two channels are shown. This is the usual number; more could clearly be used by incorporating a multi-position switch. There are two practical limitations to the number of channels, apart from the increased complexity of the switch. One is that the system involves time-sharing between the inputs and, if the sampling rate is too low, information may be lost. The other is that there is not room on the screen for more than a certain number of traces.

The circuitry in beam-switched oscilloscopes varies but, in general, most of the Y amplification is accomplished prior to the switching in order to minimize the effect of switching 'noise'. The choice of switching frequency is important. One possibility is to display one channel for the whole of one sweep and the next for the whole of the following sweep, and so on. The system becomes unworkable at very low sweep speeds because of flicker. Other objections are that some phenomena of a periodic nature may always coincide with the off period


The Series 400 oscilloscope camera (J. Langham Thompson) embodies a fast processing unit which provides photographic records of traces 60 seconds after exposure, and a safe-light glass in the hood to permit the c.r.t. face to be viewed while recording

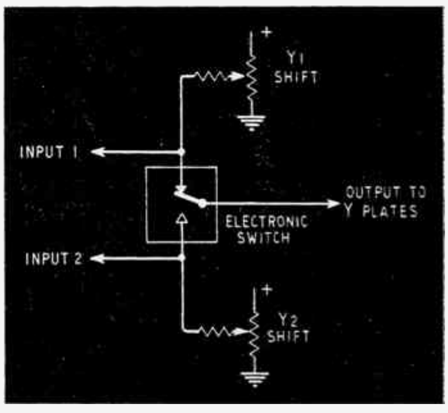
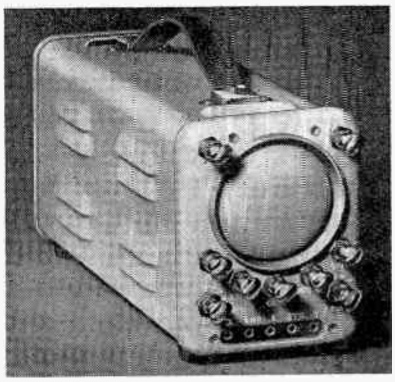


Fig. 18. *Diagram of beam-switching arrangement*



The Cossor miniature oscilloscope 1039M Mk II is intended for use by servicing engineers

of a channel and so never be displayed, and that the true time-relationship of the two signals is not shown.

The switching frequency of the above system is the same as the time-base frequency and the switching operation is performed during flyback. An alternative is to operate the switch continuously at a high frequency not simply related to the sweep or signal frequencies. The traces are then made up of lines of dots, but successive dots do not fall in the same position and the trace appears normal. Ideally, the switching waveform should be square. The upper limit of switching frequency is fixed by the performance of the switch, or the Y bandwidth, and is typically 100 kc/s in a good oscilloscope. All commercial oscilloscopes using the beam switching technique are two-channel instruments.

A third means of obtaining several displays from a single-beam tube is the voltage-coincidence system. This (Fig. 19) involves sampling the Y signals at a rate determined by a free-running oscillator. This oscillator supplies a continuous Y-deflection signal, but the trace is blacked out in the absence of an input signal. Both signal and oscillator voltages are applied to voltage-coincidence detectors (one for each channel), and when coincidences occur output pulses are produced. These are applied to the c.r.t. grid as brightening pulses. The display takes the form of lines of dots, each dot lying on the waveform so that, if the oscillator frequency is high enough, the shape of the wave can be observed accurately. Unfortunately, the dots are spread out into lines at high sweep speeds, so the bandwidth of the oscilloscope is severely limited. A sweep time of 1 μsec would require brightening pulses of the order of 1 mμsec.

Perhaps for this reason, the voltage-coincidence system has found no application in general-purpose instruments. It has, however, the merit that little signal amplification is required, and this can be provided by single-ended amplifiers. The amount of signal amplification required is determined by the threshold stability of the coincidence circuits, since the latter can easily be made to yield output pulses of constant amplitude, irrespective of the input-signal amplitude.

In assessing the merits of different systems of multiple-trace working, cost must be weighed against the facilities available. In many ways, multiple-gun tubes might be expected to give the best all-round performance, because each gun system usually functions independently, and can have its own X shift, focus and brilliance controls, and brightness modulation can be applied to one trace only if required. The only drawback, apart from the cost of the necessary tube, is that variations in linearity and deflection sensitivity may, to some extent, invalidate comparisons between traces. This, however, is a matter of c.r.t. specification and, in general, multiple-gun tubes are capable of good linearity (e.g., 1% between different traces or at different parts of the same trace). The split-beam system is the simplest but may be more liable to cross-talk between channels and it is less easy to secure good focus. The beam-switching system enables a conventional tube to be used, with some sharing of Y amplifiers and complete sharing of focus and brilliance controls.

Photographing C.R.T. Traces

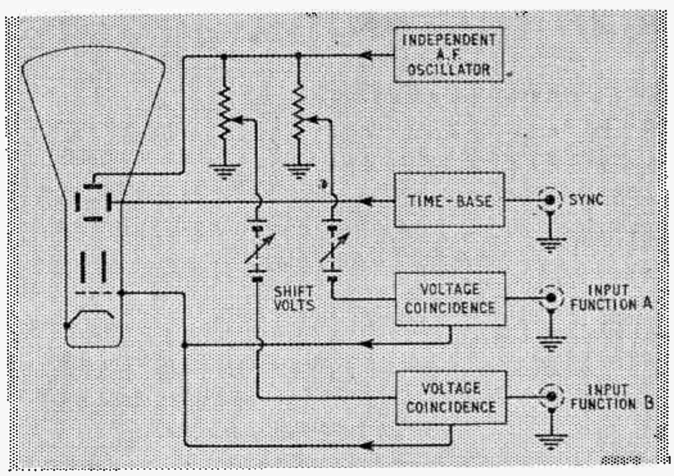
The majority of modern oscilloscopes carry fixtures which enable cameras to be attached, and a number of suitable cameras are available. Oscillograms can be taken either on film or on photo-sensitive paper, the latter being cheaper. Standard lengths of photographic material are obtainable in various widths; typical lengths are 25, 50 and 200 feet, and typical widths are 35, 70 and 120 mm.

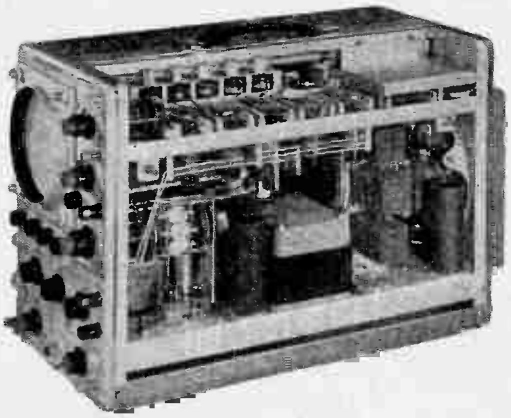
The exposure time required depends on the colour of the light emitted by the screen phosphor as well as the screen brightness and, in general, a blue (calcium tungstate) phosphor is the most suitable.

The film feed mechanism is arranged so that the film (or paper) moves along the X axis. Three ways of taking photographs are in general use. Still pictures (i.e., the film is static while the picture is taken) are used for high-speed traces. In many cases where transients must be recorded, it is necessary to synchronize the start of the sweep with the opening of the camera shutter. Cameras with contacts associated with the shutter-operating mechanism are manufactured, and some oscilloscopes make special provision for the initiation of the sweep by opening or closing contacts.

It is often useful, where comparatively low-frequency waveforms are to be examined over long periods of time, to provide a continuous record of the trace. This can be

Fig. 19. *Arrangement of circuit elements in the coincidence-voltage oscilloscope.* (*Wireless World, February* 1956, *p.* 85)





Tektronix Type 515 *oscilloscope showing the construction. The sides of the cabinet are removable. Cooling fan and air filter are situated at the rear of the instrument. The time-base provides sweep rates of* 0·2 *μsec/cm to* 6 *sec/cm and the Y response is d.c. to* 15 *Mc/s. Time and voltage are measurable to better than* 3%

accomplished by switching off the time-base and moving the film steadily past the screen. The film drive is provided by an electric motor, gears being incorporated in the system to provide film speeds of the order of 1–100 inches/second. Where higher speeds are required, a drum camera can be used. In this, a short length of film is wrapped round a drum which rotates behind the camera lens. Contacts are provided on the drum for initiating and ending the take, and top speeds of thousands of inches per second are obtainable. The oscilloscope is set up as before, with the time-base switched off. However, it is possible to extend the length of the record by applying a slow sweep voltage to the Y input along with the signal, the trace being arranged to start, say, at the top of the drum and finish at the bottom. The usable Y amplitude is then limited by the trace-to-trace spacing.

When photographing continuously, it is useful to be able to initiate the take by means of camera contacts but, since the time-base is not operative, this cannot be done in the way described above. Instead, the beam must be switched on by a brightening signal. This presents some difficulty if, as is usual, the cathode of the tube is at a high negative potential with respect to earth. If long 'takes' are required, coupling a brightening pulse to the cathode-ray tube grid by means of a capacitor becomes impracticable on account of the large physical size and high cost of the capacitor required. The latter must be able to withstand the e.h.t. voltage and, at the same time, provide a long enough time constant. An alternative is to rectify the output of a high-frequency oscillator to provide a d.c. brightening signal. The oscillator valve is switched on by the camera, and the associated rectifier is connected to an insulated winding on the oscillator coil.

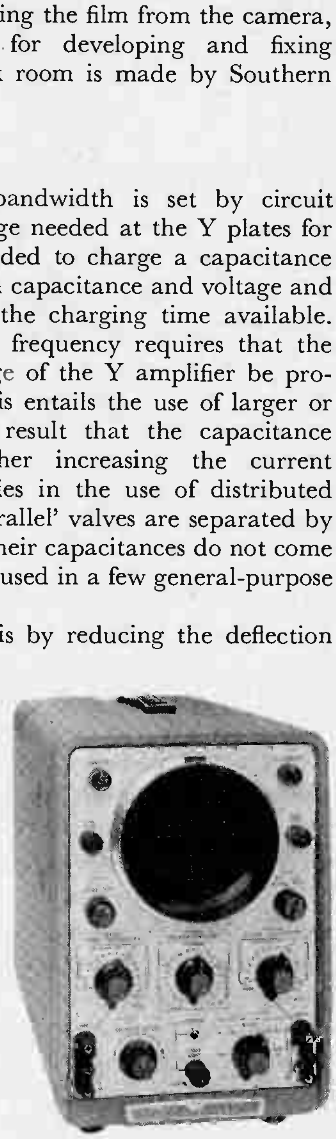
Many cameras are fitted with a film-footage indicator so that the length of film unexposed can be read off. (This is particularly useful for 'still' working with manual film feed.) Another useful refinement is a 'safe-light' viewing hood which enables the tube screen to be observed while the exposure takes place.

The exposed film must, of course, be developed and fixed. A special camera is made by J. Langham Thompson by means of which these processes are carried out rapidly without removing the film from the camera, and an 'auto-processor' for developing and fixing without the use of a dark room is made by Southern Instruments.

Conclusion

The upper limit on bandwidth is set by circuit capacitance and the voltage needed at the Y plates for deflection. Current is needed to charge a capacitance and is proportional to both capacitance and voltage and inversely proportional to the charging time available. An increase of the upper frequency requires that the current in the output stage of the Y amplifier be proportionally increased. This entails the use of larger or parallel valves with the result that the capacitance increases also, still further increasing the current required. One solution lies in the use of distributed amplification in which 'parallel' valves are separated by transmission lines so that their capacitances do not come in parallel. This system is used in a few general-purpose oscilloscopes.

Another line of attack is by reducing the deflection



Hewlett-Packard 130*A oscilloscope. The Y amplifier has a constant bandwidth of d.c. to* 300 *kc/s, and the maximum sensitivity is* 1 *mV/cm. Drift after warm-up is specified as* 1 *mV per hour referred to the input*

voltage needed by the oscilloscope tube. This can be done by using post-deflection acceleration for the electron beam. The deflection sensitivity is then a figure corresponding to a low final anode voltage for an ordinary tube but, as the main acceleration is obtained after deflection, adequate brightness of trace is secured. This is a very promising method, for a doubling of the tube sensitivity enables twice the bandwidth to be obtained from given valves in the Y amplifier. It is a method which is being increasingly used in the more expensive instruments.

In preparing this article, the specifications of a great many commercial instruments were examined. They fell into the price range of £30–£600. It is evident, therefore, that there are enormous differences in commercial oscilloscopes, and they are chiefly in the facilities provided, although quality does enter in some degree.

The prospective purchaser of an oscilloscope must thus consider his requirements rather carefully. The cheapest instruments are usually of restricted bandwidth or sensitivity and have simple time-bases without any sweep-expansion facilities. Quite small tubes are used, too, and for measurement purposes the accuracy is not high. Nevertheless, they are adequate for the examination of a.f., and sometimes of the lower r.f. waveforms. For fault-finding and routine laboratory purposes they are often all that is necessary and, in some cases, the provision of several such instruments may be more useful than a single more elaborate one of equivalent total cost. Even in cases where they are inadequate for some of the work to be done, they can be useful auxiliaries to a better-grade instrument.

Cost goes up rapidly with bandwidth, precision and the inclusion of refinements like sweep-expansion, multiple traces, etc. Not only cost, but size and weight increase, too, so it is as well to think carefully about what one needs the oscilloscope for. The best course is probably to draw up one's minimum specification and then to choose an oscilloscope which easily meets it.

MATHEMATICAL TOOLS **By Computer**

Summation and Manipulation of Series

The possibility of using series expansion has already been mentioned in this series of articles in connection with operational calculus (December 1957). In general, series are so useful for dealing with expressions that cannot otherwise be handled at all easily that we shall here consider their properties generally, and techniques for manipulating them advantageously.

A very useful collection of series whose sums (to n terms or to infinity) are known explicitly has been made by **L. B. W. Jolley**[1]. The proofs of many of these results are difficult, and here we shall not be much concerned with proofs, but rather with making the best use of well-known results.

Let us then first consider the simple geometric series

$$S_n = a + ar + ar^2 + ar^3 + \ldots + ar^{n-1} \quad \ldots \quad (1)$$

in which each term is obtained from the previous term by multiplying by a fixed number r. Both a and r may be complex numbers in Equ. (1). If we multiply both sides of Equ. (1) by r and subtract from Equ. (1) itself, all the terms cancel on the right-hand side except the first and the last, so that

$$S_n(1-r) = a(1-r^n) \quad \ldots \quad (2a)$$

and therefore

$$S_n = a\frac{1-r^n}{1-r} \quad \ldots \quad (2b)$$

The geometric series is unusual in having an explicit sum to n terms, given by (2b) for all values of r, real or complex, with the single exception of $r=1$. If $r=1$, (2b) takes the form $0 \div 0$; one way of dealing with this situation is to go back to (1) which gives immediately that if $r=1$, then $S_n=na$. There is also a useful rule for dealing with expressions like (2) which are, in general, "well-behaved" (i.e., continuous and differentiable any number of times), but become indeterminate at isolated points. To find the limiting value of S_n when r tends to 1, differentiate the numerator $(1-r^n)$ and put $r=1$ after differentiation, giving the value $-n$, and differentiate the denominator $(1-r)$ putting $r=1$ after differentiation, giving -1. Then

$$\lim_{r\to 1} S_n = a\frac{(-n)}{(-1)} = na \quad \ldots \quad (3)$$

in complete agreement with the commonsense result already obtained. In symbols, this rule is usually expressed by saying that if $f(x)$, $g(x)$ are two well-behaved functions such that $f(a)=g(a)=0$, and $f'(x)$, $g'(x)$ are the first derivatives of $f(x)$ and $g(x)$ respectively, then

$$\lim_{x\to a} \frac{f(x)}{g(x)} = \lim_{x\to a} \frac{f'(x)}{g'(x)} \quad \ldots \quad (4)$$

When a series has a sum to n terms for which there is an explicit formula, like Equ. (2), this formula can be differentiated (or integrated) any number of times. (With adequate safeguards, as we shall see later, term-by-term integration of infinite series also is often permissible.) Thus by differentiating Equ. (1) and (2) with respect to r, we find that (for $n \geqslant 2$)

$$S'_n = a+2ar+3ar^2+4ar^3+\ldots+(n-1)ar^{n-2} = \frac{dS_n}{dr}$$
$$= a\frac{d}{dr}\left(\frac{1-r^n}{1-r}\right) = a\frac{1-nr^{n-1}+(n-1)r^n}{(1-r)^2} \quad \ldots \quad (5)$$

and further sums of series may be obtained by repeated differentiation.

We notice that in Equations (2) and (5), if the modulus $|r|$ of r is less than 1 and n is sufficiently large, then the term $-r^n$ in the numerator of Equ. (2), and the corresponding terms $-nr^{n-1}$ and $(n-1)r^n$ in the numerator of Equ. (5) may be omitted. This is formally stated by saying that the series (1) has a 'sum to infinity' of $a/(1-r)$ if $|r|<1$; the series S'_n in Equ. (5) likewise has a 'sum to infinity' $a/(1-r)^2$. If however $|r|$ equals or exceeds 1, no terms can be omitted from Equs. (2) and (5), and S_n, S'_n do not tend to limits as n tends to infinity. A series having a 'sum to infinity' is called 'convergent'; a series which is not convergent is called 'divergent'. Some authors use the word 'divergent' to imply that the sum of n terms of the series increases without limit as n increases, so that they use another word 'oscillatory' for a series like

$$1-1+1-1+1-1+\ldots\text{ to } n \text{ terms } = \frac{1}{2}\left\{1-(-1)^n\right\} \quad \ldots (6)$$

whose sum to n terms does not tend to a limit as n tends to infinity but, nevertheless, does not increase without limit. [Series (6) is of course a special case of Series (1) with $a=1$ and $r=-1$.] Here, however, we shall use the word 'divergent' in the broader sense of 'not convergent', because from the point of view of an engineer, a divergent series is useless. Mathematicians are at liberty in their own world to explore with delight and enjoyment the finer points of difference between various kinds of divergence, but their conclusions are of little assistance to engineers because the borderline cases which give rise to these finer points seldom occur in practice and, when they do, special methods to extract the information of practical value can usually be applied.

For engineering purposes, we need the smallest possible number of simple tests which will show us when a series is convergent (it will be an advantage if rapid convergence is clearly exhibited). We also need computational tricks for an adequate evaluation of the sum (to infinity) of a convergent series. Convergence is no guarantee that there is an explicit formula for the sum to n terms, as in Equ. (2).

For series having only positive terms, there are only two tests of convergence important from the practical point of view, namely :

(i) The 'comparison' test : if u_n is the nth term of the series we wish to use, and $u_n \leqslant v_n$ for all n greater than some value n_0, then the series

$$U_n = u_1 + u_2 + u_3 + \ldots + u_n \quad \ldots \quad (7)$$

is convergent if the series

$$V_n = v_1 + v_2 + v_3 + \ldots + v_n \quad \ldots \quad (8)$$

is convergent.

(ii) The 'ratio test': if in Equ. (7)

$$\lim_{n\to\infty} \frac{u_{n+1}}{u_n} = l \;\; (l<1) \quad \ldots \quad (9)$$

then the series U_n is convergent. Note that this test fails if $l=1$; if $l>1$ then the series is definitely divergent. Rapid convergence is usually associated with small l.

In order to use the comparison test, we need to know a few useful convergent series. Foremost among these is the geometric series, Equ. (1), when $|r|<1$. Using the comparison test with the geometric series as the series of known convergence whose terms are respectively greater than those of the given series, however, is equivalent to using the ratio test. But now consider the series

$$W_n = 1 + \frac{1}{2^p} + \frac{1}{3^p} + \frac{1}{4^p} + \ldots + \frac{1}{n^p} \quad \ldots (10)$$

for which the ratio test fails. If $p=1$, the series (10) is divergent, but if p is any greater number than 1, whether an integer or not, the series (10) is well known to be convergent. If p is an even integer, Jolley[1] gives an explicit formula for the sum to infinity. Hence the series (10) may prove useful in showing the convergence of series for which the ratio test fails, but the convergence of such series is likely to be inconveniently slow.

For series having terms of either sign, or complex terms, the most important convergence test is that of 'absolute convergence'. If the series

$$X_n = |u_1|+|u_2|+|u_3|+\ldots+|u_n| \quad \ldots \quad \ldots (11)$$

is convergent, where the vertical bars denote absolute or modulus values, then the series (7) is called absolutely convergent. An absolutely convergent series is not only convergent ; it is a very 'safe' sort of series, for which the derivative (or integral) of the sum to infinity may be obtained by adding the derivatives (or integrals) of the separate terms. Also, any rearrangement of the order of the terms of an absolutely convergent series is permissible: it does not affect the sum to infinity. It is, however, possible to alter the sum to infinity by rearranging the order of the terms of other series. Grouping terms, without rearranging their order, is always permissible. Thus the series

$$Y_n = 1 - \frac{1}{2} + \frac{1}{3} - \frac{1}{4} + \frac{1}{5} - \ldots + \frac{(-1)^{n+1}}{n} \quad \ldots (12)$$

is not absolutely convergent, because if the signs are all made positive, we obtain the series (10) with $p=1$. It is nevertheless convergent because, if we group the terms in pairs, we have

$$\begin{aligned} Y_{2n} &= \left(1-\frac{1}{2}\right)+\left(\frac{1}{3}-\frac{1}{4}\right)+\left(\frac{1}{5}-\frac{1}{6}\right) \\ &\qquad + \ldots + \left(\frac{1}{2n-1}-\frac{1}{2n}\right) \\ &= \frac{1}{1\cdot 2}+\frac{1}{3\cdot 4}+\frac{1}{5\cdot 6}+\ldots+\frac{1}{2n(2n-1)} \quad (13) \end{aligned}$$

and the nth term of the grouped series is less than, say, $1/(2n^2)$, so that we can apply the comparison test with series (10) having $p=2$ to prove convergence. We must, of course, not overlook the fact that Y_{2n+1} ought also to be considered, but Y_{2n+1} differs from Y_{2n} only by a single term $1/(2n+1)$ at the end, and this term tends to zero as n tends to infinity. Actually, the sum of the series Y_{2n} to infinity is $\log_e 2$. On the other hand, the series

$$Z_n = x - \frac{x^3}{3} + \frac{x^5}{5} - \ldots + (-1)^{n+1}\frac{x^{2n+1}}{2n+1} \quad \ldots (14)$$

is absolutely convergent for $|x|<1$, since if in the series (14) we make all the signs positive, we obtain in this case a series satisfying the conditions of the ratio test. The sum to infinity can be obtained explicitly, for

$$\frac{dZ_n}{dx} = 1 - x^2 + x^4 - \ldots + (-1)^{n+1}x^{2n} \quad \ldots (15a)$$

Now the right-hand side of Equ. (15) is a special case of Equ. (1) with $a=1$ and $r=-x^2$, so that applying Equ. (2), we have

$$\frac{dZ_n}{dx} = \frac{1-(-x^2)^n}{1+x^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots (15b)$$

and if $|x|<1$, Equ. (15b) reduces to

$$\lim_{n\to\infty} \frac{dZ_n}{dx} = \frac{1}{1+x^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots (15c)$$

Since the series (14) is absolutely convergent for $|x|<1$, and the terms of the series (14) are obtained by integrating the corresponding terms of the series (15a), the sum to infinity of the series (14) is given by integrating (15c), that is

$$\lim_{n\to\infty} Z_n = \tan^{-1} x \quad \ldots \quad (16)$$

In fact, (16) is true if $|x|=1$ as well as for $|x|<1$. Since $\tan^{-1} 1 = \pi/4$ the series (14) could be used to calculate π, but its convergence would be very slow, even if the terms were grouped like those of the series (13). There is, however, what might be described as a trick identity which enables π to be calculated much more rapidly by means of the series (14). We have

$$\tan^{-1} a + \tan^{-1} b = \tan^{-1} \frac{a+b}{1-ab} \quad \ldots \quad (17)$$

Using this identity with $a = b = \frac{1}{5}$ gives

$$2 \tan^{-1} \frac{1}{5} = \tan^{-1} \frac{5}{12} \quad \ldots \quad (18)$$

Using the identity (17) again with $a = b = 5/12$ gives

$$4 \tan^{-1} \frac{1}{5} = 2 \tan^{-1} \frac{5}{12} = \tan^{-1} \frac{120}{119} \quad \ldots \quad (19)$$

and finally using the identity (17) with $a=1$, $b=1/239$ gives

$$\tan^{-1} 1 + \tan^{-1} \frac{1}{239} = \tan^{-1} \frac{120}{119} \quad \ldots \quad (20)$$

From these results we have

$$\frac{\pi}{4} = \tan^{-1} 1 = 4 \tan^{-1} \frac{1}{5} - \tan^{-1} \frac{1}{239} \quad \ldots \quad (21)$$

Now the series (13) is very rapidly convergent if $x = 1/5$ or $x = 1/239$, so that the manipulation involved in Equs. (17) to (21) would save an appreciable amount of time if accurate calculation of π to say 20 decimal places was required. It may well be argued that this is a mathematician's job and not an engineer's; the example merely illustrates the power of algebraic and other manipulations to change a mathematical expression into a form much more suitable for the particular task in hand.

A power series

$$P_n = a_0 + a_1 z + a_2 z^2 + \ldots + a_n z^n \quad \ldots \quad (22)$$

usually has a 'radius of convergence'; that is to say, there is a number R such that the series is absolutely convergent when $|z|<R$. For a given power series, R is usually determined easily by replacing z by $|z|$ in the series (22), and applying the ratio test. It is possible for R to be infinite, as in the case of the exponential series, or zero, as in the case where a_n in the series (22) is $n!$ Normally, however, R is finite, so that the power series can be used freely for $|z|<R$, and differentiated and integrated term by term also for $|z|<R$.

Having established the convergence (or, if possible, absolute convergence) of a series, we have to consider obtaining an adequate approximation to the sum to infinity in the general case when neither the sum to n terms nor the sum to infinity are known or obtainable from such a source as Reference 1.

We have already seen, in connection with the series (12) and (13), that grouping the terms, provided that their order is not altered, may transform the given series into one which converges more rapidly. The same kind of improvement can be made if we wish to sum a series which is similar to, but not identical with, one whose sum is known. If, for instance, we wish to sum the series (7) when

$$u_n = 1/(n^2 + 1) \quad \ldots \quad (23)$$

we can consider the difference between the series (23) and the series (10) whose sum to infinity is known to be $\pi^2/6$ when $p = 2$. We have

$$(1/n^2) - u_n = (1/n^2) - 1/(n^2+1) = 1/\{n^2(n^2+1)\} \quad \ldots \quad (24)$$

Now the series associated with the last member of (24) is much more rapidly convergent than that represented by Equ. (23); probably it is sufficient merely to add up a reasonably small number of terms in this case. If we wish, however, we can repeat the process and say

$$u_n - (1/n^2) + (1/n^4) = 1/\{n^4(n^2+1)\} \quad \ldots \quad (25)$$

and the series whose general term is the last member of (25) will be still more rapidly convergent; the series (10) when $p = 4$ has the known sum $\pi^4/90$.

Sometimes a series having terms of the same absolute value as those of the given series has a known sum, but there are discrepancies of sign, or the given series may have terms missing from those of a series with known sum or, more generally, the nth term of the given series can be broken up into a combination of terms each of which is associated with a known series. The techniques available in such circumstances are well illustrated in terms of the series

$$A_n = \frac{1}{2^2} - \frac{2}{3^2} + \frac{3}{4^2} - \ldots + (-1)^{n+1} \frac{n}{(n+1)^2} \quad (26)$$

The sum to infinity is given by Jolley as $(\pi^2/12) - \log_e 2$, and we seek to derive this result using only series already mentioned above.

First of all, the nth term of the series (26) can be written

$$(-1)^{n+1} \frac{n}{(n+1)^2} = (-1)^{n+1} \frac{(n+1)-1}{(n+1)^2}$$
$$= (-1)^{n+1} \left[\frac{1}{n+1} - \frac{1}{(n+1)^2}\right] \quad \ldots \quad (27)$$

The first part of (27) gives us the series

$$\frac{1}{2} - \frac{1}{3} + \frac{1}{4} - \frac{1}{5} + \ldots \quad \ldots \quad (28)$$

and this is the series (13) with the signs reversed and the first term missing. The sum to infinity of (28) is therefore $1 - \log_e 2$. The second part of (27) gives the series

$$-\frac{1}{2^2} + \frac{1}{3^2} - \frac{1}{4^2} + \ldots \quad \ldots \quad (29)$$

which is like the series (10) with $p=2$ except that the first term is missing, and the signs are alternate instead of all positive. We can rearrange (29) in the form

$$(-1) + 1 + \frac{1}{2^2} + \frac{1}{3^2} + \frac{1}{4^2} + \ldots$$
$$- \frac{2}{2^2} \quad - \frac{2}{4^2} - \ldots \quad \ldots \quad (30)$$

and the sum of the top row of (30) is now clearly seen to be $(\pi^2/6) - 1$, while the terms of the bottom row can be taken respectively as

$$-\frac{1}{2}\cdot\frac{1}{1^2}, -\frac{1}{2}\cdot\frac{1}{2^2} \text{ etc.}$$

so that their total contribution is $-\frac{1}{2}\cdot(\pi^2/6)$ or $-\pi^2/12$.

Hence the sum to infinity of the series (22) is

$$(1-\log_e 2)+(\frac{1}{6}\pi^2-1)-(\frac{1}{12}\pi^2) = \frac{1}{12}\pi^2-\log_e 2$$

in agreement with Jolley[1].

The above remarks are intended merely to outline the ways in which series can be manipulated to give results useful to the engineer. The subject of series has rightly received a great deal of attention from mathematicians, and cases of difficulty should be quickly brought to their notice. Here we have merely tried to show that engineers need not be frightened of series.

REFERENCE

[1] L. B. W. Jolley, "Summation of Series". Chapman & Hall, London, 1925.

Wide-Band Transformer Characteristics

By A. C. Hudson*

SUMMARY. *A parameter f_r is defined for wide-band transformers, which represents the series resonance of the leakage inductance and the primary and secondary stray capacitances. It is shown that this parameter may be determined from the low-frequency requirements on the transformer, and then used as a guide to the attainable high-frequency response.*

Wide-band radio-frequency transformers were described by Maurice and Minns[1] in 1947. These transformers usually consist of a relatively small number of turns of wide copper foil wound helically on a closed core, with polytetrafluoroethylene (trade names: I.C.I., "Fluon"; DuPont, "Teflon") tape as insulation. The core material is often a nickel-zinc or manganese-zinc ferrite.

A formula is derived which permits a preliminary estimate of the feasibility of meeting a given specification with a transformer of this type.

Leakage Inductance

Fig. 1 shows the arrangement of the windings, and Fig. 2 the terminologies used in references 1 and 2, and in the present paper.

The leakage inductance in microhenrys given in Equ. (50) of reference 2 (Connelly) after some rearrangement, can be written:

$$L = \frac{8\pi^2 N^2}{10^3 h}\left(C_0 b_0 + \frac{C_1 b_1}{3} + \frac{C_2 b_2}{3}\right) \quad \ldots \quad \ldots \quad (1)$$

where N is the number of turns on the winding to which the leakage inductance is referred, and the other symbols are dimensions in centimetres, as defined in Fig. 2. In order to compare the formula of reference 1 (Maurice and Minns), the terminology of the latter's Equ. (7) is altered and after some rearrangement the leakage inductance in microhenrys is given by:

$$L = \frac{8\pi^2 N^2}{10^3 h}\left[C_0 b_0 + \frac{C_1 b_1}{3} + \frac{C_2 b_2}{3} + \frac{1}{12}(b_1^2 - b_2^2)\right] \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

It can be seen that for practical purposes the formulæ are in agreement; the small term $\frac{1}{12}(b_1^2 - b_2^2)$ can be ignored. For a non-redundant terminology we choose the symbols shown at the extreme right of Fig. 2, and after converting the dimensions from centimetres to inches, the expression takes the following convenient form:

$$L_L \approx \frac{N^2}{30h}[2(r_2^2 - r_1^2) + (r_3^2 - r_0^2)] \quad (\mu\text{H}) \quad (3)$$

A simple derivation of this expression is given in the appendix. Now $2(r_2^2 - r_1^2)$ can very often be made negligible, and then:

$$L_L \approx \frac{N^2}{30h}(r_3^2 - r_0^2) = \frac{N^2}{30h}(r_3 - r_0)(r_3 + r_0) \quad (\mu\text{H}) \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

In a practical case, the factor $(r_3 - r_0)$ will vary much more rapidly with r_3 than the factor $(r_3 + r_0)$; in other words, the leakage inductance is roughly proportional to the winding thickness, $(r_3 - r_0)$. Thus, replacing $(r_3 + r_0)$ by $2r_a$:

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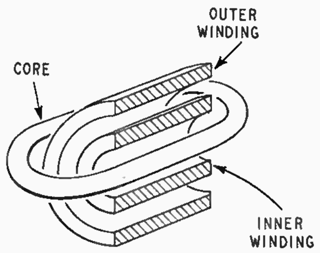

Fig. 1. *Basic arrangement of windings*

$$L_L \approx \frac{N^2 r_a}{15} \cdot \frac{(T_1 + T_2)}{h} \quad (\mu\text{H}) \quad .. \quad .. \quad (5)$$

r_a is the average winding radius of the transformer in inches; T_1 and T_2 are the winding thicknesses in inches (see Fig. 2); and $(r_2 - r_1)$ is assumed to be small.

Stray Capacitance

To calculate the effective stray capacitance of a helical winding, consider a five-turn winding (Fig. 3) with a total applied voltage of 5 volts, to be 'unrolled' as in Fig. 4. It is seen that four sections in parallel result, with a constant potential difference of one volt. In general, N turns yield $N - 1$ sections, with an applied voltage of $1/N \times$ the total applied voltage.

Thus the effective stray capacitance is:

$$C = \begin{bmatrix}\text{Capacitance}\\ \text{of}\\ \text{1 section}\end{bmatrix} \times \begin{bmatrix}\text{Effect of } N-1\\ \text{sections}\\ \text{in parallel}\end{bmatrix} \times \begin{bmatrix}\text{Effect of}\\ \text{voltage}\\ \text{reduction}\end{bmatrix}$$

$$= [C_1 \text{ section}] \times [N-1] \times \left[\frac{1}{N^2}\right]$$

Applying the formula (reference 3, p. 112) for the capacitance of a parallel-plate capacitor gives

$$C = \frac{N-1}{N^2} \cdot 0{\cdot}2244K(A/d) \quad (\text{pF}) \quad .. \quad .. \quad (6)$$

where K is the relative permittivity of the dielectric, A is the area of one plate in square inches, and d is the separation of the plates in inches.

A may be replaced by $2\pi r_a h$, and d may be replaced by T_1/N, for the inner winding, or T_2/N for the outer winding, where the thickness of the conductor is, for the present, considered negligible compared with the thickness of the insulation. $N - 1$ may be approximated by N, then:

$$C_3 \approx 1{\cdot}41Kr_a\frac{h}{T_1} \quad .. \quad .. \quad .. \quad .. \quad (7)$$

where C_3 is the stray capacitance of the inner winding, in picofarads. Similarly C_4 is the stray capacitance of the outer winding.

Capacitance of Two Windings in Series

Let C_5 represent a hypothetical capacitance equal to C_3 and C_4 in series.

$$C_5 = \frac{C_3C_4}{C_3 + C_4},$$

hence,

$$C_5 = 1{\cdot}41Kr_a\frac{h}{T_1 + T_2} \quad (\text{pF}) \quad .. \quad .. \quad (8)$$

With the exception of those transformers which have a very small number of turns, C_5 may be considered to be the stray capacitance of both windings, connected series-aiding. For an autotransformer, C_5 becomes merely the stray capacitance of the complete transformer. C_5 is a convenient capacitance to measure as it does not involve any coupled capacitance. Because readings must be made over a range of frequencies, the measurements are made with the core removed in order to minimize inductance variations.

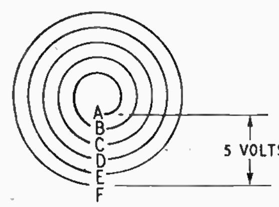

Fig. 3. *Five-turn helical winding*

'Resonance' of the Winding

By examining Equs. (5) and (8) it is seen that the product L_LC_5 takes a particularly simple form, being independent of h, T_1 and T_2.

$$L_LC_5 \approx \frac{N^2{r_a}^2}{10{\cdot}65}K \quad .. \quad .. \quad .. \quad .. \quad (9)$$

Or, in terms of the frequency of series resonance, f_r, of L_L, C_3 and C_4

$$f_r \approx \frac{1040}{DN\sqrt{K}} \quad (\text{Mc/s}) \quad .. \quad .. \quad .. \quad .. \quad (10)$$

where $D = 2r_a$, the average winding diameter, in inches. This equation is the main result of the present paper. (Note that it is not a transformed value of C_4 which is considered to be in series with C_3, but the actual value.)

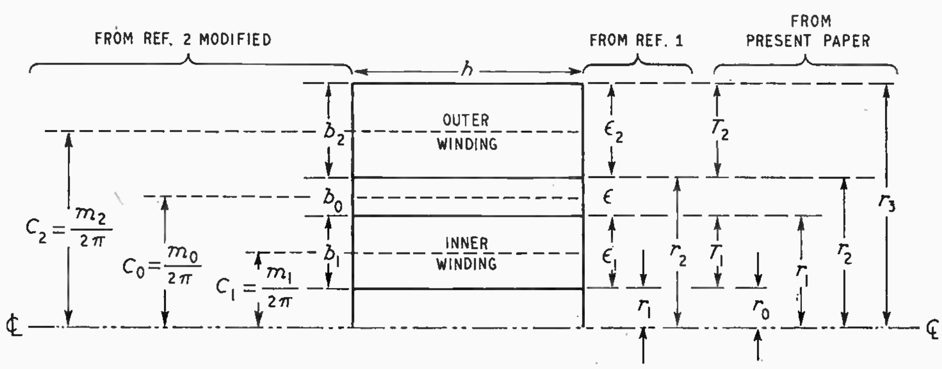

Fig. 2. *Terminology for calculating leakage inductance*

Since the two windings are not in fact in series, the usefulness of the parameter f_r might be doubted. It may, however, be used to evaluate an existing transformer; that is, f_r may be calculated from measured values of leakage inductance and stray capacitance. If the theoretical f_r, as given in Equ. (10), exceeds this value, then an improvement in design is probably possible. Equ. (10) shows the nature of the fundamental limitation on transformers of the type considered. Leakage inductance may be 'traded' for stray capacitance and, independently, the capacitances of the two windings may be chosen; but only in such a way as to keep their series value constant.

It is reasonable to assume that the upper frequency at which a given transformer is useful will be roughly

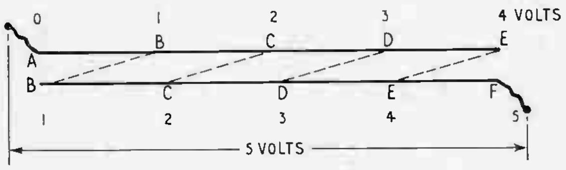

Fig. 4. *Five-turn helical winding, 'unrolled'*

proportional to f_r, and experience shows that a resistively terminated transformer may be used to within about 75% of f_r, f_r being measured from the winding having the largest number of turns. It is hoped to show in a later paper that for the rather specialized case of an unterminated transformer driving a capacitive load, the performance of the optimum transformer may be directly expressed in terms of f_r. The relation also shows that the upper frequency limit is inversely proportional to Nr_a and hence the design at the lower frequency limit, f_b, should be made in such a way as to minimize Nr_a if maximum bandwidth is required.

Equ. (10) may be used to provide a preliminary estimate of the possibility of building a transformer to meet a given specification. Then f_r is essentially determined by the requirements on the transformer at its lower frequency, as is discussed briefly in the next section. On the other hand, tolerable values of leakage inductance and stray capacitance may be deduced from the high-frequency requirements on the transformer. From these, a second value of f_r may be calculated. If this f_r is less than the former value of f_r as determined by direct application of Equ. (10), then the transformer can be built, and conversely. There is, of course, an indeterminate range.

Considerations at Bottom Frequency (f_b)

The transformer must have a certain impedance at f_b, and the core material and shape having been chosen, this imposes a minimum on N^2r_0. Let this minimum be denoted by M_I. A convenient formula for N^2r_0 is

$$N^2r_0 = \frac{17{\cdot}7L_{\mu H}K_s}{\mu_r} \quad \dots \quad (11)$$

$$N^2r_0(_{min}) = M_I \text{ (as determined by impedance)} \quad (12)$$

where $L_{\mu H}$ is the required shunt inductance, in microhenrys,

K_s is the shape factor of the core and is equal to $\dfrac{\text{length of flux path}}{\sqrt{\text{core area}}}$,

r_0 is the effective core radius, in inches,

μ_r is the effective relative permeability of the core,

and N is the number of turns on a given winding.

Further, the transformer must be able to accept a certain r.m.s. voltage E. The maximum flux density having been chosen from considerations of cooling or saturation, the standard transformer equation (reference 3, p. 98)

$$E_{(\text{r.m.s.})} = 4{\cdot}44FNAB \times 10^{-8} \quad \dots \quad (13)$$

shows that Nr_0^2 has a minimum value (where A is the core area in square centimetres, and B is the peak flux density in gauss).

Let this minimum be denoted by M_v. Substituting πr_0^2 for A, and converting r_0 from centimetres to inches:

$$Nr_0^2 = \frac{E}{0{\cdot}9f_bB} \quad \dots \quad (14)$$

where f_b is the bottom frequency in megacycles. Thus,

$$Nr_0^2(_{min}) = M_v \text{ (as determined by voltage)} \quad (15)$$

It is instructive to plot a curve of the type of Fig. 5. Here the abscissa is Nr_0 and the ordinate is r_0. From Equ. (12) it is easily shown that r_0, as a function of Nr_0, has a maximum equal to $(Nr_0)^2/M_I$. This is plotted as locus OA in Fig. 5.

From Equ. (15) it is easily shown that r_0, as a function of Nr_0, has a minimum M_v/Nr_0; this is locus BC in Fig. 5.

Now, Equ. (10) shows that the best high-frequency performance will be obtained when Nr_0 is a minimum; hence we seek the point on Fig. 5 which is in a permissible region and is furthest to the left. This is obviously point P_0. If a safety factor in flux density is desired, a point such as P_1 may be chosen while, if a safety factor in impedance is required, a point such as P_2 could be used.

If one is forced to use some point, say P_3 for example, because only certain discrete values of core radius are available, or because of a desire to use an integral

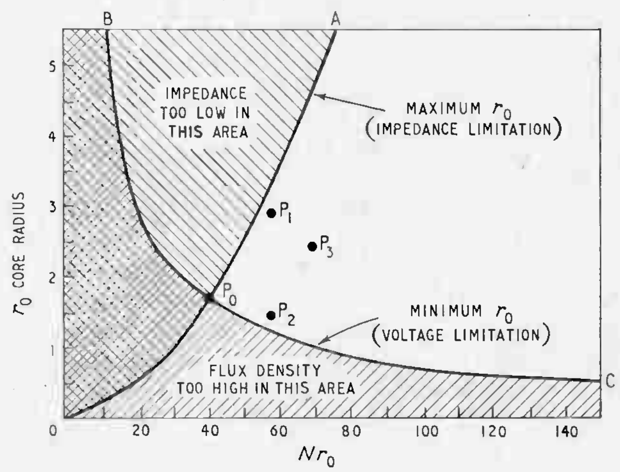

Fig. 5. *Low-frequency limitations on number of turns and core radius (typical example)*

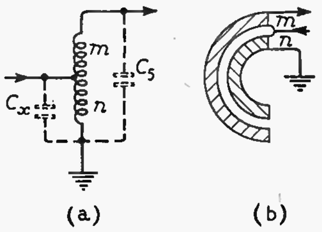

Fig. 6. *Unbalanced autotransformer*

number of turns, then the ratio of the abscissa of P_3 to that of P_0 may be used as a measure of the penalty in the upper frequency limit, since Nr_0 appears linearly in the denominator in Equ. (10).

Point P_0 may be determined analytically; by solving Equs. (12) and (15) it is found that,

$$N = \sqrt[3]{(M_I^2)/M_v} \quad .. \quad .. \quad .. \quad .. (16)$$

and

$$r_0 = \sqrt[3]{(M_v^2)/M_I} \quad .. \quad .. \quad .. \quad .. (17)$$

In the previous discussion r_0 and r_a have been tacitly assumed to be the same. Actually, r_a is the core radius (r_0) plus a suitable estimate of half the thickness of the windings. This does not affect the argument appreciably,

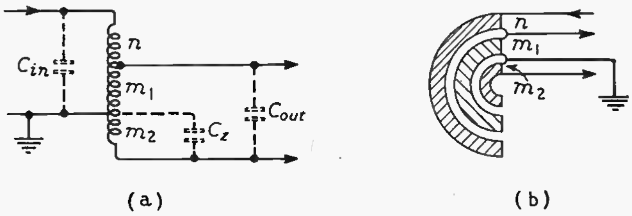

Fig. 7. *Balanced autotransformer*

however. If the leg of the core is not circular in cross-section, r_0 may be taken as 0·564 times the square root of the cross-sectional area of the leg.

Autotransformers

The form of Equ. (10) is unchanged for the autotransformer connection, but the capacitance C_5 has a new significance. Consider first the unbalanced autotransformer of Fig. 6. By reference to Fig. 6(b) and by recalling the derivation of Equ. (10), it is obvious that C_5 now represents not two capacitances in series, but the full capacitance as seen from the $m + n$ side. It is important to note that the input capacitance C_x [see Fig. 6(a)] has no independent existence, which is a further advantage of the autotransformer over the two-winding transformer.

As discussed in the appendix, the autotransformer connection gives a considerable improvement in leakage inductance. For the case of Fig. 6, the leakage inductance may be calculated as if the transformer were a two-winding transformer having the same total winding thickness, and the calculated result divided by the factor $(1 + n/m)^2$ as given in Equ. (19).

Consider now the unbalanced-to-balanced autotransformer of Fig. 7. Here it can be seen that C_5, inherent in Equ. (10), represents C_{in} and C_z in series; or in other words C_5 has a capacitance equal to C_{in} and $2C_{out}$ connected in series.

As in the case of the unbalanced transformer, the leakage inductance may be calculated as if the primary and secondary windings were separate, and had the same total winding thickness; and a factor applied to convert to the autotransformer case. The factor which accounts for the autotransformer connection is given in Equ. (22); f_r is, of course, inversely proportional to the square root of the leakage inductance.

Finite Conductor Thickness

Previously it was assumed that the conducting foil had zero thickness. To allow for its finite thickness, it may be shown that the capacitance will be increased by a factor $(1 + t_{cu}/t_i)$, where t_{cu} is the thickness of the copper foil, and t_i is the thickness of the insulating tape. Thus f_r must be corrected as follows:

$$f_{r\,\text{(corrected)}} = f_{r\,\text{(uncorrected)}} \div \sqrt{1 + t_{cu}/t_i} \quad .. \quad .. (18)$$

Size of Core

It has been shown in Equ. (10) that high-frequency response will always suffer if the core is larger than necessary. Thus, it is conventional to make the core size just large enough to meet the voltage and impedance requirements at f_b, the bottom of the required frequency band.

It is also conventional to group these transformers as either low-power or high-power, a low-power transformer being one where presumably the core size does not matter. It should be appreciated, however, that this distinction is artificial and that, if optimum bandwidth is to be obtained, the core must always be reduced in size until the signal level to be handled becomes a limitation, and then in one sense the transformer is 'high-power' even though it may be handling only a few microwatts.

Examples

While Equ. (10) was derived for a simple tape-wound configuration, it may be used as a standard of compari-

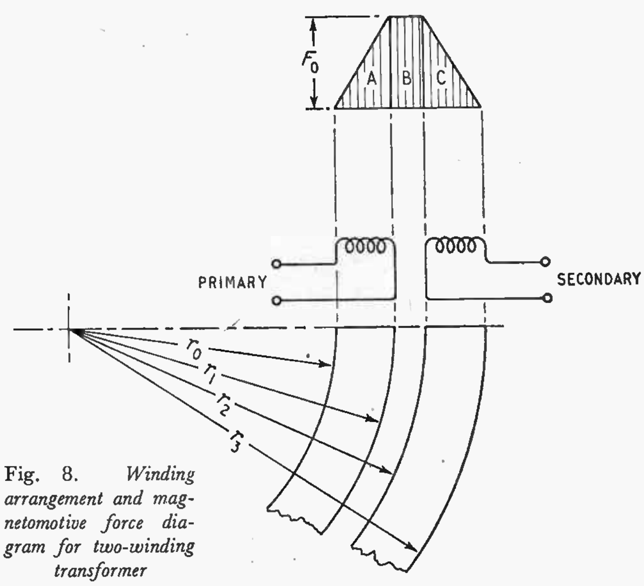

Fig. 8. *Winding arrangement and magnetomotive force diagram for two-winding transformer*

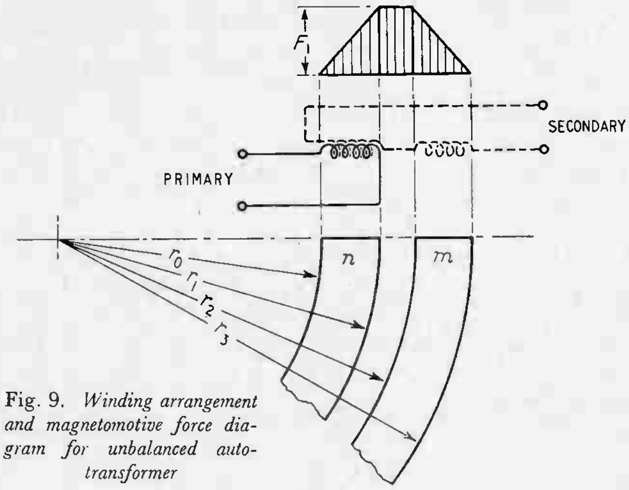

Fig. 9. *Winding arrangement and magnetomotive force diagram for unbalanced autotransformer*

son for other methods of construction. The four examples below, taken from various sources, include several methods of construction.

As a first case, consider the example from reference 1. Maurice and Minns' Fig. 26 shows the average radius to be 0·196 in., $N = 14$, and $K = 2{\cdot}3$. Equ. (10) predicts a basic f_r of 125 Mc/s, which becomes 88·5 Mc/s after applying Equ. (18) to correct for the finite copper thickness. From the dotted curve B of Fig. 13, which is a plot of Equ. (22), the leakage inductance is less than the value calculated above by a factor of 2·4, hence f_r is increased by $\sqrt{2{\cdot}4}$, giving 137 Mc/s.

Consider now their measured values, $C_5 = 10$ and 20 pF in series, while L_L is 0·12 μH; thus $f_{r\text{ (measured)}} =$ 177 Mc/s. This large discrepancy is due to the difference between their calculated and measured values of C.

A similar transformer was constructed in these laboratories, having the core removable to facilitate determination of stray capacitance. The following results were obtained: measured leakage inductance, 0·128 μH; measured value of C_5, 6·6 pF. These resonate at 173 Mc/s, while Equ. (10) predicted 176 Mc/s for this transformer.

As a second example, we consider the transformer of Hibbard *et al.*, reference 4. Core area is 7 sq. in., so D is 2·98 in., the primary N is 8 turns, and $K = 2{\cdot}3$. Equ. (10) gives 28·7 Mc/s for f_r; there is no correction for copper thickness because of the spacing.

To calculate C_5, we note that $C_s + 8^2C_p = 1020$ pF. Now, since the secondary has only one turn, and hence no overlap, C_s will be small compared with 8^2C_p, hence $C_p \approx 16$ pF. But C_s is large compared to C_p, so C_s and C_p in series $\approx C_p \approx 16$ pF. The leakage inductance referred to the primary is 2·02 μH, the resonance is 28 Mc/s, which agrees with 28·7 Mc/s found above.

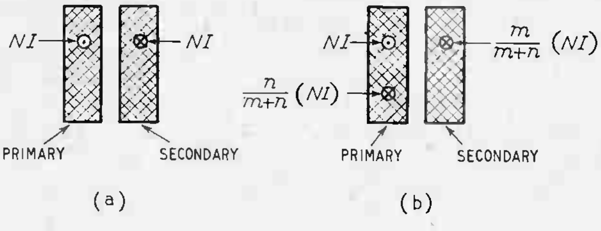

Fig. 10. *Diagrammatic cross-section of windings;* (a) *two-winding transformer,* (b) *autotransformer*

As a third example, we consider an entirely different construction, the 'split screen' type. This is the example from reference 5. By scaling O'Meara's Fig. 8(b), we find that the core area is 0·437 sq. in., hence the equivalent core diameter is 0·745 in., $N = 11$ (considering one-half of the primary), and $K = 2$. Applying Equs. (10) and (18), $f_r = 73{\cdot}2$ Mc/s. The measured value of C_5 is 8·9 and 61·8 pF in series, and that of L_L is 5·5 μH. These resonate at 24·1 Mc/s, which is considerably less than the value 73·2 Mc/s found above. This particular example of a split-screen transformer will be inferior in frequency response to a simple helical tape transformer.

A simple two-winding tape-wound transformer was constructed, having the same core area and number of turns as the above example; the calculated value of f_r was 59·8 Mc/s, while the measured resonance was 62·5 Mc/s. It is not claimed that this particular transformer would necessarily meet all of the requirements of the transformer of reference 5.

The calculated value of f_r for the transformer of reference 6 is 458 Mc/s. (It is assumed that rice paper

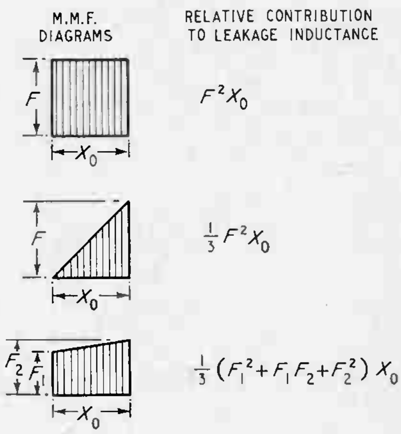

Fig. 11. *Magnetomotive-force diagrams*

was used for insulation.) The quoted leakage inductance is 0·11 μH and the appropriate value of C_5 is that of two 39-pF capacitors in series. These resonate at 109 Mc/s, which is considerably below the calculated value; thus improvement in the design of this transformer is possible.

The above four examples represent all the transformers we have found which are described in sufficient detail to permit comparison of the calculated and measured values of f_r. A number of tape-wound transformers have been constructed in these laboratories; the measured and calculated values of f_r have been found to agree within five per cent.

Conclusions

The main purpose has been to treat wide-band transformers from a slightly novel viewpoint. It is hoped that the ideas developed will supplement the conventional treatment. No reference has been made to aspects, such as the low-pass filter analogy, which have been treated fully elsewhere.

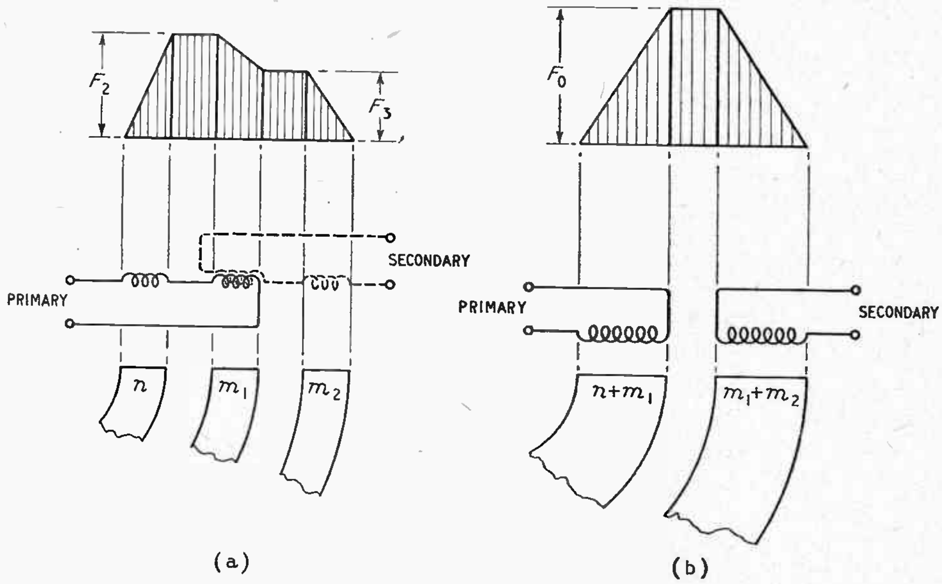

Fig. 12. *Comparison of balanced autotransformer and two-winding transformer;* (a) *balanced autotransformer,* (b) *two-winding transformer*

APPENDIX

Calculation of Leakage Inductance: Two-winding Transformer

Leakage inductance may be calculated by the method described by A. Boyajian (page 72 of reference 7) wherein the primary and secondary are regarded as return circuits to each other from the point of view of their leakage magnetic field. Applying the method to the simple two-winding transformer (Fig. 1 or 2 herein, Fig. 16, p. 72 of reference 7) we note that the leakage m.m.f. diagram is as shown in Fig. 8.

By cancelling I^2 and $2\pi f$ from Boyajian's equation (18) the leakage inductance due to the gap is seen to be

$$0{\cdot}2\frac{rg}{h}N^2 \text{ microhenries}$$

Now r is the mean radius of the gap, which equals $\frac{1}{2}(r_2 + r_1)$ in the terminology of the present article, and also $g = (r_2 - r_1)$ from which the leakage inductance due to the gap is,

$$\frac{0{\cdot}1N^2}{h}(r_2{}^2 - r_1{}^2)$$

The leakage inductance due to the field within the windings themselves may also be calculated easily. The linear reduction of magnetomotive force towards the edge of the windings, which is indicated in regions A and C of Fig. 8, will cause a square-law reduction in the resulting leakage inductance. Since the integral of x^2dx from 0 to a is $1/3a^3$ it is easily shown that the contribution of area A in Fig. 8 is one-third the contribution from a rectangular area of the same base and height.

Thus the contributions to leakage inductance to be added due to regions A and C, Fig. 8, are,

$$\frac{0{\cdot}1N^2}{h}\cdot\frac{r_1{}^2 - r_0{}^2}{3} \text{ and } \frac{0{\cdot}1N^2}{h}\cdot\frac{r_3{}^2 - r_2{}^2}{3}$$

Summing the contributions of all three regions, the leakage inductance becomes expression (3) of the present paper.

Unbalanced Autotransformer

Formulæ for the autotransformer cases will now be derived. For simplicity, the average radius of all windings will be assumed to be the same. Factors of the form $(r_1{}^2 - r_0{}^2)$ may be replaced by $(r_1 - r_0)$ since the factor $(r_1 + r_0)$ represents twice the average radius, which may be assumed to be the same for all windings considered, and hence to cancel out.

Consider the unbalanced autotransformer of Fig. 6 and Fig. 9. The gap $(r_2 - r_1)$ is included for generality; usually in the autotransformer case this gap will equal zero. We are concerned with the factor:

$$\frac{\text{leakage inductance of a two-winding transformer}}{\text{leakage inductance of an equivalent unbalanced autotransformer}}$$

The two transformers which are being compared will have the same transformation ratio, and the same value of h, etc. It is still possible to define 'equivalent' in two ways, however. The total number of turns in the autotransformer (Fig. 6) is $m + n$. For the two-winding transformer of Fig. 8 to be equivalent it must have n primary turns and $m + n$ secondary turns, or a total number of turns of $m + 2n$. If the same thickness of copper and insulating tape is used in both cases, then the total base of the m.m.f. diagram will be larger in Fig. 8 than in Fig. 9. This will be defined as 'complete equivalence'. 'Partial equivalence' will be defined as the condition where the total thickness of the winding, and hence the base of the m.m.f. diagram, is the same for both transformers. This is a rather arbitrary case, where thinner material must be used for the two-winding transformer; but it is convenient for calculation.

As is explained by Boyajian[7] the leakage inductance is proportional to the area under a curve which has the same base as the m.m.f. diagram and with an ordinate proportional to the square of the m.m.f. We wish therefore to consider the relative m.m.f. diagrams in Figs. 8 and 9. Referring to Fig. 9, the ampere-turns of the primary and secondary are equal, therefore the secondary current is $n/(n + m)$ of the primary current. In the autotransformer the secondary current subtracts from the primary, thus the m.m.f. F is $[1 - n/(n + m)]$ of the m.m.f. F_0 in Fig. 8. Hence,

$$\frac{F_0}{F_1} = \frac{m+n}{m} = \left(1 + \frac{n}{m}\right)$$

Considering the gap $(r_2 - r_1)$ to be zero, and squaring,

$$\frac{\text{two-winding transformer leakage}}{\text{autotransformer leakage}} = \left(1 + \frac{n}{m}\right)^2$$

(for partial equivalence) .. (19)

This confirms equation (43) of reference 1, which was obtained by a different method (see their Fig. 22).

For complete equivalence we also consider the advantage of the autotransformer in that its total thickness is less. For zero gap,

$$\frac{\text{two-winding thickness}}{\text{autotransformer thickness}} = \frac{n+2m}{n+m}$$

Combining with Equ. (19) above,

$$\frac{\text{two-winding transformer leakage}}{\text{autotransformer leakage}} = \left(1 + \frac{n}{m}\right)^2\cdot\left(\frac{n+2m}{n+m}\right)$$

$$= \left(1 + \frac{n}{m}\right)\left(2 + \frac{n}{m}\right) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (20)$$

We note that the error due to ignoring the differences in average radius of the primary and secondary is not great, since only ratios have been considered.

It is helpful to provide a physical explanation of the reduction in m.m.f. due to the autotransformer connection. Following Boyajian, we regard the primary and secondary as return circuits to each other (see Fig. 10). The process of converting the two-winding

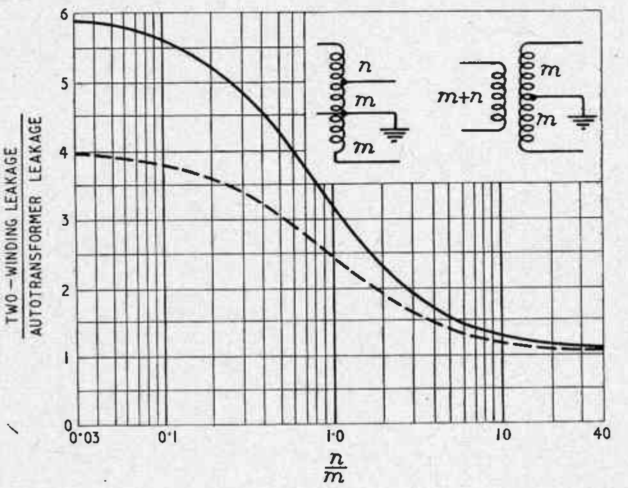

Fig. 13. *Advantage of balanced autotransformer. Solid curve: Same tape thickness, hence autotransformer smaller than two-winding transformer (complete equivalence, Equ.* 21)*; dotted curve: autotransformer winding same thickness as two-winding transformer (partial equivalence, Equ.* 22)

transformer [Fig. 10(a)] to an autotransformer [Fig. 10(b)] consists in moving $n/(m+n)$ of the current 'strands' from the secondary to the primary. Thus the effective ampere-turns are reduced in the ratio $m/(m+n)$ and also the total number of loops available for flux linkage are reduced in the same ratio. Hence the inductance is reduced as the square of this ratio.

Balanced Autotransformer

It is convenient to consider a third shape of m.m.f. diagram. This is shown as the last diagram in Fig. 11. The expression given for relative contribution to leakage inductance is easily derived by integrating F^2dx over the region X_0.

We consider the balanced autotransformer of Fig. 7, redrawn in Fig. 12(a), and it is seen that

$$F_2 = \frac{n}{n+m_1}F_0$$

since only this fraction of the primary turns contribute to the m.m.f. at the junction between the n and m portions of the winding. Similarly,

$$F_3 = \frac{m_2}{m_1+m_2}F_0$$

and introducing now the fact that $m_1 = m_2$

$$F_3 = \frac{1}{2}F_0$$

The gaps between the windings, which are shown for generality in Fig. 12, are assumed to be of zero thickness. The comparison of leakage inductance may now be made, the contributions of the various sections of the appropriate m.m.f. diagrams are added, according to the relations shown at the right in Fig. 11. The result is,

$$\frac{\text{two-winding transformer leakage}}{\text{autotransformer leakage}} = \frac{2\left(1+\frac{n}{m}\right)\left(3+\frac{n}{m}\right)}{2\left(\frac{n}{m}\right)^2 + 2\frac{n}{m} + 1} \qquad (21)$$

for complete equivalence.

For partial equivalence, the result is:

$$\frac{\text{two-winding transformer leakage}}{\text{autotransformer leakage}} = \frac{2\left(1+\frac{n}{m}\right)\left(2+\frac{n}{m}\right)}{2\left(\frac{n}{m}\right)^2 + 2\frac{n}{m} + 1} \qquad (22)$$

Relations (21) and (22) are plotted in Fig. 13. Equ. (22) does not agree with Equ. (44) of reference 1.

The words 'primary' and 'secondary' are used as a convenience: they may always be interchanged.

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[2] F. C. Connelly, "Transformers", Sir Isaac Pitman and Sons, Ltd., 1950.

[3] F. E. Terman, "Radio Engineers' Handbook", First Edition, McGraw-Hill Book Company, Inc., 1943.

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[5] T. R. O'Meara, "Wide-Band Power Amplifiers and Transmitters", *Proc. National Electronics Conference* 1955, Vol. 11, p. 251.

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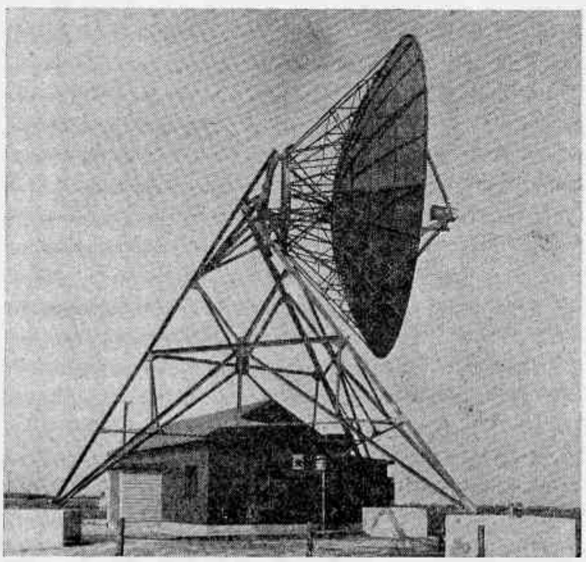
[7] L. F. Blume, A. Boyajian *et al.*, "Transformer Engineering", John Wiley and Sons, Inc., 1951, pp. 72–8.

TROPOSPHERIC-SCATTER LINK

For some years Marconi's have been carrying out experimental work using tropospheric-scatter propagation and, for the past two years, an experimental system between Bromley in Essex and Catterick in Yorkshire has been in operation. It has been used mainly for analysis of transmission characteristics.



In parallel with this work a range of tropospheric-scatter transmitting and receiving equipment has been designed and is now being put into production. The first production units have now been assembled as a multichannel radio link between Start Point, Devon, and Galleywood, Essex, a distance of just over 200 miles. The link, which works on a frequency of 858 Mc/s and has an initial capacity of 24 telephone channels, is a single-way, with the transmitter situated at Start Point and the receiver at Galleywood. The present equipment is designed for a maximum of 60 simultaneous telephone channels, but provision has been made for experimental wide-band transmissions with a view to carrying a much greater number of channels. Later, television pictures are also to be transmitted to ascertain whether an acceptable standard of quality can be realized.

The Start Point transmitter employs a water-cooled four-cavity klystron developing an output power of 10 kW. The associated aerial system, a 30-ft. diameter paraboloid excited by a horn radiator, is mounted with its centre 35 ft. above ground level. This aerial system has a gain of 36 dB over a dipole at the chosen frequency (858 Mc/s) and thus provides an effective radiated power in the direction of maximum intensity of the order of 40 mW.

The receiving equipment is of the dual-diversity type employing two dish aerials each of 30-ft. diameter and spaced 100 ft. apart. This arrangement has been found by previous experience to counteract the characteristic rapid fading of tropospheric-scatter working.

Multichannel tropospheric-scatter links can satisfactorily provide a point-to-point transmission of signals over distances of 200 miles or more as compared with the 40-mile average of direct-beam systems.

Semiconductor Switching Devices

The January issue of *I.R.E. Transactions on Electron Devices* contains articles on some interesting new transistors. These are being developed for use as switches and are all characterized by a sudden transition from a high-resistance to a low-resistance condition. This transition can be controlled by means of a base electrode, and the latter can switch the main path both on and off.

The devices exhibit a regenerative switching-on characteristic which comes about because, at some critical value of collector current, the common-base amplification factor (alpha) exceeds unity, as in a point-contact transistor. The new transistors appear to have considerable advantages over point-contact transistors in that they are large-area devices, capable of passing as much as 25A, and are easier to manufacture.

Though different in form, the new devices appear to work on the same basic principle. The alpha of a transistor-like structure is raised above unity by a flow of minority carriers originating from the collector. This minority-carrier current increases rapidly as the normal collector current is increased, leading to an abrupt transition from the high-resistance to the low-resistance state. Switching off is not regenerative and larger control powers are required. Some details of the switching devices are given below.

The Thyristor[1]

This is a medium-current germanium device with the following characteristics:

Turn-on time	50–100 mμsec
Turn-off time	100 mμsec
Turn-on pulse energy	10^{-4} erg
Turn-off pulse energy	0·1 erg
'On' voltage drop	0·5 V at 100 mA
Hold-off voltage	60 V at room temperature with no base current.

The physical structure of the thyristor is shown in Fig. 1, and its operating characteristics in Fig. 2. It is a p-n-p transistor, with a large-area base–collector junction produced by diffusing arsenic into p-type germanium. An ohmic connection is made to the 'n' side of this slab to form the base connection, and a small-area emitter of indium-aluminium alloy is soldered to the base, creating a local p-n junction.

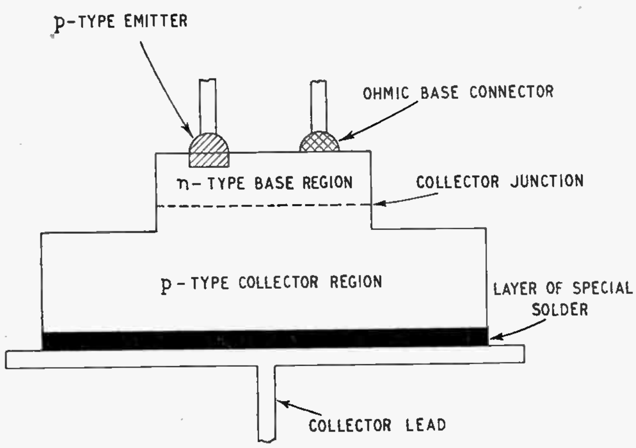


Fig. 1. *Cross-section of the 'thyristor'*

The switching characteristics of the device arise from the nature of the collector contact. This is made by soldering a nickel tab to the collector area, using an alloy of lead, tin and indium. This contact, although ohmic in nature, is capable of injecting electrons into the collector region. These are swept to the emitter, increasing the emitter–collector current. If this process is described in transistor terms, the device has a collector–emitter alpha which has a maximum value in excess of 0·1. This 'electron-alpha', unlike the normal p-n-p 'hole-alpha', is strongly dependent on collector current, and increases about 100 times as the latter is increased from 1 to 10 mA. Since the two alphas add, a critical current is reached at which the total alpha exceeds unity, and a regenerative switching-on action occurs. This critical current is well-defined, because of the way in which the 'electron-alpha' varies with hole current, and the originators of the device state that it is controllable in manufacture by changes in the soldered collector contact.

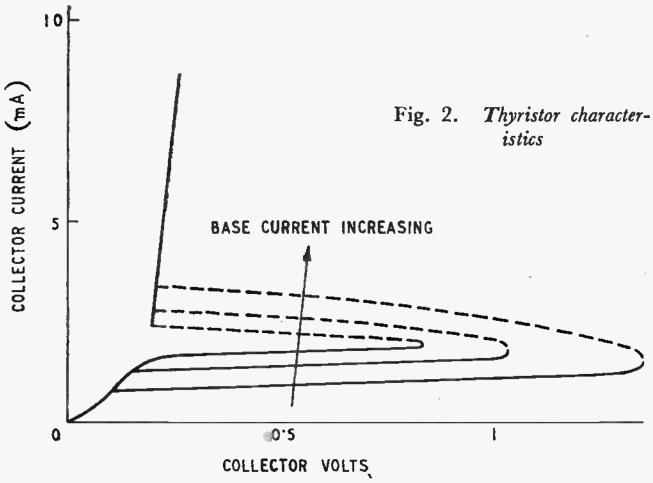


Fig. 2. *Thyristor characteristics*

P-N-P-M Transistor[2]

Although somewhat different physically (Fig. 3), this appears to work in the same way as the thyristor; i.e., the alpha of a diffused-base germanium transistor is raised above unity by injecting minority carriers into the collector by means of a special metallic contact. (The latter accounts for the 'm' in p-n-p-m.) Devices capable of blocking 350 V and passing 10 A or more have been made. The report in *Trans. I.R.E.* gives much information on turn-off and turn-on characteristics. The ratio of output to turn-off power can be as large as 800. Typical operating characteristics for a low-power device are:

Turn-on time	1–10 μsec
Turn-off time	2–10 μsec
Turn-on current	0·6 mA
Turn-off current	15 mA
Hold-off voltage	65 V at zero base current.

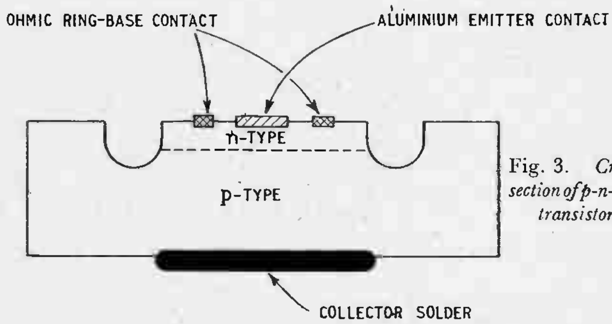


Fig. 3. *Cross-section of p-n-p-m transistor*

P-N-P-N Transistor[3]

This is a development of two-terminal p-n-p-n silicon devices, which can be constructed to behave somewhat like gas-discharge tubes; i.e., with a high resistance below the breakdown voltage and a low resistance when the applied voltage exceeds the breakdown value. In the three-terminal p-n-p-n device, control of the breakdown is effected by means of a base contact. Its characteristics are:

Turn-on base current	about + 20 μA
Turn-off base current	about − 230 μA
Hold-off voltage	70 V (zero base current)
Main-path current	600 μA

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[3] I. M. Mackintosh, "Three-Terminal P-N-P-N Transistor Switches", ibid, p. 10.

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.

Transistor Impedance Matching

SIR—In his article in the April 1957 issue of your journal (p. 128), Dr. H. P. Williams discussed a most useful concept for assessing the transistor matching problem. The quantities which he calls R_{GM} and R_{LM} for the generator and load resistances give precise matching at both input and output terminals of a transistor. Using the hybrid parameters throughout in preference to r_e, r_b, r_c and α, the relationships he gives may also be expressed as

$$R_{GM}^2 = \frac{h_{11}\Delta}{h_{22}} \text{ and } R_{LM}^2 = \frac{h_{11}}{h_{22}\Delta}$$

where Δ is the determinant of the h terms, namely $(h_{11}h_{22} - h_{12}h_{21})$, the matching resistances for the various configurations being correctly given if the appropriate values of the h-parameters are used.

A fairly common need in audio amplifiers is to employ two transistors in cascade, with *RC* coupling for economy of components. It is therefore of interest to compute the matching terminations for a pair of similarly-connected transistors.

Starting with the expressions:

$$R_{in} = \frac{G_L h_{11} + \Delta}{G_L + h_{22}} \quad \ldots \quad \text{(i)}$$

$$R_{out} = \frac{R_G + h_{11}}{R_G h_{22} + \Delta}, \quad \ldots \quad \text{(ii)}$$

and noting that $R_{in2} = R_{L1}$
and $R_{G2} = R_{out1}$,

$$R_{L1} = R_{in2} = \frac{G_{L2}h_{11} + \Delta}{G_{L2} + h_{22}}$$

$$G_{L1} = \frac{G_{L2} + h_{22}}{G_{L2}h_{11} + \Delta}$$

Putting this value of G_{L1} into expression (i)

$$R_{in1} = \frac{G_{L2}h_{11} + h_{11}h_{22} + G_{L2}h_{11}\Delta + \Delta^2}{G_{L2} + h_{22} + G_{L2}h_{11}h_{22} + h_{22}\Delta}$$

$$R_{G2} = R_{out1} = \frac{R_{G1} + h_{11}}{R_{G1}h_{22} + \Delta}$$

Putting this value of R_{G2} into expression (ii)

$$R_{out2} = \frac{R_{G1} + h_{11} + R_{G1}h_{11}h_{22} + h_{11}\Delta}{R_{G1}h_{22} + h_{11}h_{22} + R_{G1}h_{22}\Delta + \Delta^2}$$

$$G_{out2} = \frac{R_{G1}h_{22} + h_{11}h_{22} + R_{G1}h_{22}\Delta + \Delta^2}{R_{G1} + h_{11} + R_{G1}h_{11}h_{12} + h_{11}\Delta}$$

Then for matching terminations,

$$R_{in1} = R_{G1} = R_{GM}$$

$$G_{out2} = G_{L2} = G_{LM} = \frac{1}{R_{LM}}$$

When the resulting simultaneous equations in R_{GM} and G_{LM} are solved, the solutions are relatively simple :—

$$R_{GM}R_{LM} = \frac{h_{11}}{h_{22}}$$

$$\frac{R_{GM}}{R_{LM}} = \frac{h_{11}h_{22} + \Delta^2}{1 + h_{11}h_{22}}$$

or $$R_{GM}^2 = h_{11}\frac{(h_{11}h_{22} + \Delta^2)}{h_{22}(1 + h_{11}h_{22})}$$

$$R_{LM}^2 = \frac{h_{11}(1 + h_{11}h_{22})}{h_{22}(h_{11}h_{22} + \Delta^2)}$$

Taking, for an OC71 in earthed-emitter configuration,

$h_{11}' = 810$ ohms
$h_{12}' = 5 \times 10^{-4}$
$h_{21}' = 46{\cdot}5$
$h_{22}' = 76$ micromhos
$\Delta' = 0{\cdot}0384$,

We obtain $R_{GM}' = 795$ ohms and
$R_{LM}' = 13{\cdot}4$ kilohms.

These values, for two earthed-emitter transistors in cascade, are fairly close to those obtained for a single earthed-emitter OC71, namely

$R_{GM}' = 640$ ohms, $R_{LM}' = 16{\cdot}6$ kilohms

Using the expression for power gain obtained with the h-parameters:

$$\frac{G_L h_{21}^2}{(G_L + h_{22})(G_L h_{11} + \Delta)},$$

the gains of the two-stage amplifier with matching terminations is
$(32{\cdot}1 + 40{\cdot}4)$ dB $= 72{\cdot}5$ dB

There is, of course, a considerable mismatch between the two stages, since $R_{out1} = 16{\cdot}2\,\text{k}\Omega$ and $R_{in2} = 660\,\Omega$ and it is mainly because of this that the suggestion has been made, for example by R. F. Shea ("Principles of Transistor Circuits", Wiley.) of using earthed-collector and earthed-emitter stages in cascade.

It seems profitable to consider this suggestion in terms of equal generator and load resistances, either low in value, say 600 ohms, or high, say 20 kilohms, and to compare the gain available from a combination of earthed-emitter and earthed-collector amplifiers with that available from two earthed-emitter amplifiers.

Taking the terminating resistances as 600 ohms, and hence for reasonable matching an earthed-emitter stage followed by an earthed-collector stage:—computation from the *h*-parameters gives, for OC71 transistors, $R_{in1} = 600$ ohms and $R_{out2} = 360$ ohms indicating fair external matching, while $R_{out1} = 16{\cdot}7$ kilohms and $R_{in2} = 28$ kilohms, indicating fair internal matching. Then the gains of the two stages are 40·1 dB and 16·5 dB respectively, or 56·6 dB total.

Considering now two earthed-emitter stages operating between the same terminating resistances, $R_{in1} = 790$ ohms and $R_{out2} = 13{\cdot}4$ kilohms, while $R_{out1} = 16{\cdot}7$ kilohms and $R_{in2} = 794$ ohms. So there is significant mismatching both internally and externally. Yet the gains of the two stages are 32·8 dB and 31·7 dB respectively, or 64·5 dB total, an advantage of some 8 dB over the previous method.

Such analysis as this serves to explain why in practice the earthed-emitter transistor amplifier is so widely used in *RC*-coupled amplifiers in spite of its apparent disadvantage due to mismatching.

Auckland University College, A. G. Bogle.
Auckland,
New Zealand.
24th April 1958.

Amplifier Low-Frequency Compensation

Sir,—Messrs. J. E. Flood and J. E. Halder—in their article "Amplifier Low-Frequency Compensation" published in your March 1958 issue—state that they have been unable to find any particular case in which a seventh-order compensation of gain-frequency response and a sixth-order compensation of phase-frequency response could be obtained for a single *RC* coupled stage, adding that a rigorous proof of the fact that this compensation is by no means possible is not yet available. Here is, however, a rigorous proof that these compensations cannot be obtained.

For instance, for the gain–frequency response:

Let

$$g_k R_k = k,\ \sigma_s R_s = s,\ \theta = m\eta,\ \eta^2(1+m^2) = N,\ \eta^4 m^2 = P$$
$$P(k+s)(k+s+2) = L \text{ and } \eta^2[2mks+k(k+2)+m^2s(s+2)] = M$$

Then, the equations to be solved for the seventh-order compensation would be:

$$\left.\begin{aligned}(\gamma+\lambda)^2 &= \alpha^2+\lambda^2+M\\ (\gamma+\lambda)^2N &= \alpha^2\lambda^2+(\alpha^2+\lambda^2)(M+N)+L\\ (\gamma+\lambda)^2P &= \alpha^2\lambda^2(M+N)+(\alpha^2+\lambda^2)P(k+s+1)^2\end{aligned}\right\} \quad \ldots \quad (1)$$

or, eliminating $(\gamma+\lambda)^2$:

$$\left.\begin{aligned}(M+N)\alpha^2\lambda^2+L(\alpha^2+\lambda^2)-MP &= 0\\ \alpha^2\lambda^2+M(\alpha^2+\lambda^2)+L-MN &= 0\end{aligned}\right\} \quad \ldots \quad (2)$$

Hence $\alpha^2\lambda^2 = a/b$

where, putting $MN-L = Z$:

$$a = Z^2-ZMN+M^2P \quad \ldots \quad (3)$$
$$b = Z+M^2 \quad \ldots \quad (4)$$

Obviously, a/b must be positive,

$$Z = \eta^4[2ksm(1-m+m^2)+k(k+2)+m^4s(s+2)] > 0$$

Therefore, $b>0$. For a to be >0, we see from Equ. (3) that Z should have a value *outside* the range the limits of which are η^2M and η^2m^2M; i.e.,

$$\eta^2M < Z < \eta^2m^2M \quad (\text{for } m<1)$$
$$\eta^2M > Z > \eta^2m^2M \quad (\text{for } m>1)$$

But

$$Z-\eta^2M = \eta^4m^2s[2k+(s+2)(m+1)](m-1)$$
$$Z-\eta^2m^2M = \eta^4k[2ms+(k+2)(m+1)](1-m)$$

Hence, Z falls *inside* the range and therefore the three simultaneous equations (1) do not admit real roots for all the variables. Q.E.D.

Equation (18) in the article is not correct. When the time-constants of the cathode and screen circuits are very large compared with those of the coupling and anode decoupling circuits, equation (17) is the valid one, not Equ. (18).

Equation (18) is valid when the cathode and the screen are not at all decoupled or when their time-constants are very small compared to those of the coupling and anode decoupling circuits.

Polytechnic Institute Bucarest, Vlad Pauker
Rumania.
12th April 1958.

Sir,—We are very grateful to M. Pauker for providing the proof which eluded us and thus completing the treatment of the *RC* coupled stage. Equation (18) applies of course, when η and θ are very large: the time-constants ($1/\eta$ and $1/\theta$) are then obviously very small. We thank M. Pauker for drawing out attention to this unfortunate error in wording, which did not occur in the earlier paper (ref. 1) and which does not affect the subsequent analysis.

Research Laboratory, J. E. Flood
Siemens Edison Swan Ltd., J. E. Halder
Blackheath, London, S.E.3.
1st May 1958.

Subjective Sharpness of Television Pictures

Sir,—I am most indebted to Dr. N. W. Lewis for his suggestions in the May issue with reference to my paper on the above subject. I would agree that a plot of log $\{p/(1-p)\}$ against log S gives a more linear relationship and to a large extent brings the anomalous curve of Fig. 9 into line with the others. Recalculation of the results represented in Fig. 9 gives five straight lines, four of which are very nearly parallel and of slope of approximately $n = 4$. The line representing the results when the comparison picture was of sharpness factor 0·98 is not parallel to the other four and is of significantly lower slope. Thus, I cannot entirely agree with Dr. Lewis that one single value of n is sufficient to describe all the experimental results contained in Fig. 9.

The definition of the difference limen used in the paper was of a rather elementary nature and could probably be made more precise. A binomial distribution does not, however, describe the actual distribution of observers' opinions in the experiment, so that there would be some difficulty in giving a more sophisticated definition. It is to be noted that Dr. Lewis's calculation gives an answer of 2·63 Mc/s, which is not very different from the figure of 2·67 Mc/s quoted in the paper.

The problem of assessing sharpness in the presence of phase distortion (in addition to amplitude distortion) is a difficult one. It is a doubtful matter whether a single parameter can properly describe this distortion; the use of the criterion of maximum slope of the response to unit step would not distinguish between two distortions where the maximum slopes were equal but one of the distortions had a much more gradual rate of reaching the full response. The slower rate of reaching the full response would show up as a smear and the experiments in which an *RC* circuit was used demonstrated that the eye is particularly sensitive to this distortion.

The British Broadcasting Corporation, W. N. Sproson
Kingswood Warren, Surrey.
12th May 1958.

New Books

Feedback Theory and its Applications

By P. H. Hammond, B.Sc., A.M.I.E.E. Pp. 348. The English Universities Press Ltd., 102 Newgate Street, London, E.C.1. Price 35s.

This book is intended to offer an introduction and review of the subject to post-graduate engineering and physics students. At first sight this seems a contradiction, since most students nowadays learn something about feedback amplifiers and possibly servo systems before graduating, but it is not so. The thoroughness of treatment is such that as presented here the subject will be virtually a new one to the graduate who has no subsequent experience in this field. The book is introductory in the sense that much of the theory is stated without proof but with ample references to original work or more specialized treatises; it introduces the student to all the important results, but leaves proofs and amplification to his further reading. The prospective reader should also realize that this is a book on feedback *theory*; it would be impossible to include in a single volume constructional details of all the types of equipment used in feedback systems, though one or two examples are described, such as a medium-gain d.c. amplifier (gain of 80 dB and drift about 5 mV per hour) which is described in two pages of text and a full circuit diagram.

The author grasps the nettle of trying to define *feedback* and at first states that any system which opposes an applied force can be said to have negative feedback, but then undertakes to limit his discussion in this book to active systems. The difficulty of distinguishing between simple equilibrium systems, such as a bridge under steady load, and negative-feedback systems such as a governed engine, might be eased by defining feedback as a property of closed loops, and further defining a closed loop as a circuit in which information can circulate in one direction only.

A commendable feature of the book is its realistic assessment of the possibilities of dealing with non-linear systems, followed by chapters on the application of phase-plane technique and of the gain-describing-function method of working in terms of the fundamental component only when a sinusoidal signal is applied to a non-linear system. Random fluctuations are mentioned in relation to linear servo systems, but the general problem of optimizing system response in relation to a class of input signals in the presence of noise is barely indicated. Perhaps this would be too extensive a topic for an introductory survey, and in all other respects the book covers its subject very thoroughly. D.A.B.

Problems in Electronics (With Solutions)
By F. A. Benson. Pp. 219. E. & F. Spon Ltd., 15 Bedford Street, Strand, London, W.C.2. Price 36s.

"It is thought that the book covers almost the complete undergraduate electronics courses in engineering at Universities, but it has not been written to match any particular syllabus, and it should be found useful by post-graduate students and research workers as a reference source."

The answers are stated in the text and, in most cases, worked solutions are to be found in the second and larger part of the book. The exceptions are those problems to which the answers are to be found in standard textbooks; in these cases, a reference is given instead.

Basic Television
By A. Schure. Vol. 1. The Transmitter, pp. 112. Vol. 2. Organization of the TV Receiver, pp. 145. Vols. 3, 4 and 5. TV Receiver Circuit Explanation, pp. 137, pp. 121, pp. 135 respectively. Price per volume $2.25, set of 5 $10.00. John F. Rider Publisher Inc., 116 West 14th Street, New York 11, U.S.A.

An elementary non-mathematical description of television.

The Radio Amateur's Handbook 1958
By the Headquarters Staff of the American Radio Relay League. Pp. 584 + 32 pages of valve data and an index. The American Radio Relay League, West Hartford, Connecticut, U.S.A. Price $4.50.

Electrical Research Association: Annual Report 1957
Pp. 120. The Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey.

NATIONAL BUREAU OF STANDARDS

The Measurement of Thickness
By George Keinath. N.B.S. Circular 585. Pp. 79. Price $0.63.

Annual Report 1957
Pp. 143. Price $0.57.

System Design of Digital Computers at the N.B.S.: Methods for High-Speed Addition and Multiplication
N.B.S. Circular 591. Pp. 22. Price 25 cents, post paid.

Methods of Testing Thermocouples and Thermocouple Materials
By Wm. R. Roesen and S. T. Lonberger. N.B.S. Circular 590. Pp. 21. Price 25 cents, post paid.

Electroforming of Waveguide Components for the Millimeter-Wavelength Range
By Albert A. Feldmann. N.B.S. Circular 587. Pp. 16. Price 19 cents, post paid.

The above are available from the Superintendent of Documents, U.S. Government Printing Office, Washington 25, D.C., U.S.A.

BRITISH STANDARDS

Pren Type Electric Cables for Aircraft
B.S.2E.21; 1957. Now fully in line with current practice, and incorporates experience gained in the use of the 1952 issue. Price 7s. 6d.

Cartridge Fuse-Links for Telecommunication and Light Electrical Apparatus
B.S. 2950: 1958. The fuse-links dealt with supersede the Type B fuses which were specified in B.S. 646: 1935. Price 4s. 6d.

Nyvin Type Electric Cables for Aircraft
B.S.E.24: 1957. Specifies cables having an insulation of p.v.c. compound and glass braid with a protective covering of nylon. Price 7s. 6d.

Cartridge Fuse-Links (Rated up to 5 A for a.c. and d.c. service)
B.S. 646: 1958. Deals with non-rewirable fuse-links for two-wire circuits carrying up to 5 A at up to 250 V. Price 4s. 6d.

Loaded and Unloaded Ebonite for Electrical Purposes
B.S. 234: 1957. Price 6s.

Annual Report 1956–57. Price 7s. 6d.

Year Book 1958
Contains information about the B.S.I. and lists the British Standards current on 1st January, 1958, giving a brief description of each. Price 15s.

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

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for April* 1958

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

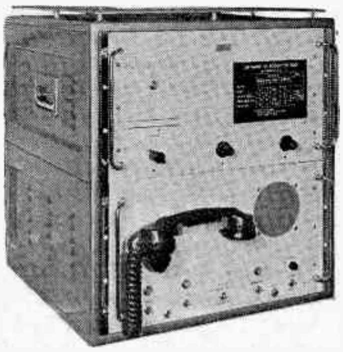
* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 830 *c/s for the N.P.L. caesium resonator. N.M. = Not Measured.*

New Products

Single-Sideband Radio-Telephone

This new 60-W transmitter/receiver (type TRA.55) makes available the effective gain in power which is given by s.s.b. transmission to the user who requires a small, compact and inexpensive equipment. It may also be used for c.w. operation.

It is designed to be extremely simple to operate and to be used easily by entirely unskilled personnel. Four pre-set channels are provided and a single switch selects the



transmitter and receiver channels simultaneously.

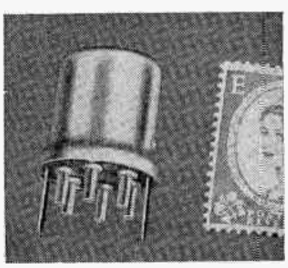
Points from the makers' specification are: Frequency range, 3–12 Mc/s; channels, 4 crystal-controlled (2 in the band 3 to 6 Mc/s; 2 in the band 6 to 12 Mc/s); output power, 60 W; aerial, two outputs (a) 75-ohm for tuned aerials; (b) single-ended output for long-wire untuned aerials; finish, high-grade tropical standard for ambient temperatures of 40°C; power supply, 100–125 V and 200–250 V, 40/60 c/s; a.c. supply consumption—'receive', 95 W; 'transmit', 300 W.

Racal Engineering Ltd.,
Bracknell, Berks.

Miniature Coils

Intended for transistor radio receivers, these are claimed to be the smallest screened coils available on the British market or the Continent. Their dimensions are: height (excluding lugs), $\frac{5}{8}$ in.; diameter (top), $\frac{9}{16}$ in.; diameter (base), $\frac{5}{8}$ in.

Ferrite pot cores and litz windings are employed and the average Q is 180. The turns ratios are designed to match Mullard



OC44/OC45, and Siemens XA101/102 transistors.

There are three types available for covering the medium-wave broadcast band: Type 735, oscillator coil; Type 736, 1st or 2nd i.f. transformer; Type 737, 3rd i.f. transformer.

Channel Electronic Industries Ltd.,
Dunstan Road, Burnham-on-Sea, Somerset.

Trace-Reading Equipment

This is intended to provide comprehensive facilities for the semi-automatic analysis and reduction of analogue trace records.

An extremely wide range of calibrations is possible, both linear and non-linear, enabling the trace data to be presented in digital form in its original units. The presentation is by visual display and a simultaneous permanent record can be made by a typewriter, card punch or tape perforator.

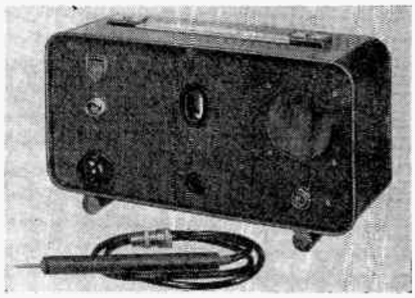
Records up to 6 in. in width are acceptable and the full calibration facilities provided may be applied to traces having between 1 and 6 in. deflection from the datum. The datum may be located anywhere on the record.

Digital output is provided on three ranges of 0–±999, 0–±1999 or 0–±4999. Speed of analogue to digital conversion and visual display is about 0·2 sec. The typewriter control incorporates decimal-point selection for printing-out purposes.

Southern Instruments Computer Division,
Frimley Road, Camberley, Surrey.

Signal Tracer

The Amos model 136 signal tracer consists of a high-gain amplifier with which is associated a simple valve voltmeter (utilizing an electron-beam tuning indicator) and a miniature loudspeaker. The input circuit is arranged so that it operates automatically as a leaky-grid detector when an r.f. signal is applied, or as a grid-current biased linear amplifier when an a.f. signal is present. The valve-voltmeter rectifier is independent of this input circuit and operates on all signals whether modulated or unmodulated, r.f. or a.f. The pencil probe provided may thus be applied to any part of the r.f., local oscillator, i.f. or audio circuits, when the 'magic eye' will indicate the



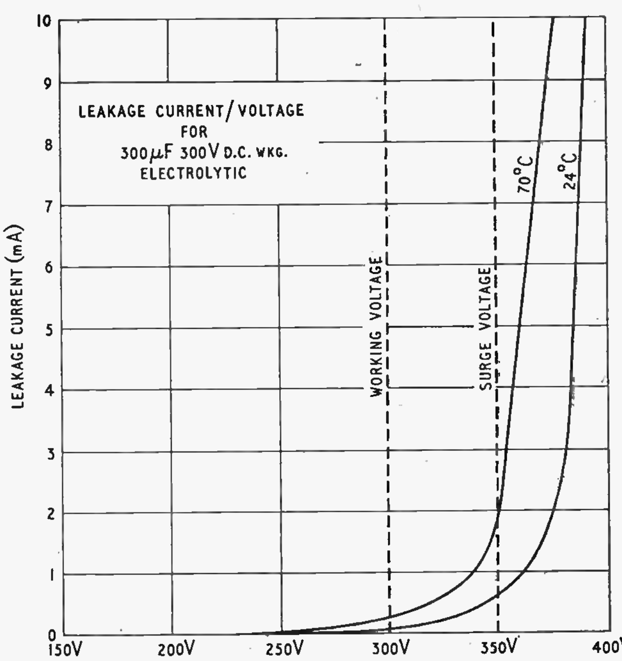
presence and amplitude of the signal and the speaker will render the a.f. component audible.

The instrument is manufactured by Messrs. Amos of Exeter and is available from the distributors at the address below.

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

Electrolytic Capacitors

T.C.C. capacitors for a.c./d.c. receivers are now rated at 300 V peak working, with a surge rating of 350 V. The capacitors may be used for h.t. smoothing in sets operated from 250-V a.c. mains; under these conditions, they may be charged to the peak mains voltage for a short time after switching on. The makers point out that, in designing equipment, attention should be given to the leakage current–voltage curves for electrolytic capacitors, one of which is shown here. A capacitor may be able to withstand a given surge at room temperature but not at a higher temperature. If equipment which has been working some time and has got hot



is switched off for a short time and then switched on again, failures may occur.

The Telegraph Condenser Co. Ltd.,
Radio Division, North Acton, London, W.3.

Miniature Mains-Operated Soldering Iron

This pencil-sized soldering iron for mains operation is made by mounting a fully-insulated element (flash tested at 900 V a.c.)

inside a steel shaft approximately 3 5 mm in diameter.

It is claimed that the insulated element and the exact temperature maintained at the bit should prove very useful for soldering transistors and that live circuits can be soldered without harm.

A.N.T.E.X. Ltd.,
3 Tower Hill, London, E.C.3.

Flash Tester

This instrument is designed for the non-destructive batch testing of components, or other breakdown applications, at voltages of

250, 500, 1,000, 1,500 and 2,000 a.c. These voltages are selected by a high-grade ceramic switch controlled from the front panel.

Labgear Ltd.,
Willow Place, Cambridge.

Test Equipment for Multi-Channel Links

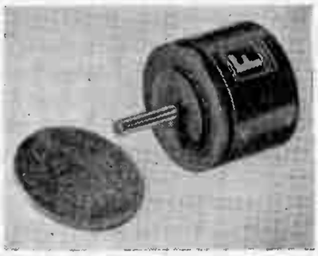
The specialized nature of many of the circuits and techniques used in link systems has created a demand for equally specialized test instruments. To meet this demand, Marconi Instruments Ltd. have introduced a number of new equipments including u.h.f. Test Set OA 1248, which provides comprehensive testing facilities for radio-link equipment in the 1,700–2,300-Mc/s band, and White Noise Test Set OA 1249, used for the measurement of noise and intermodulation distortion in multi-channel telephony equipment.

Other instruments include Precision Slotted Line TF 1233, for standing-wave and impedance measurement in the frequency range 1,700–2,300 Mc/s; an r.f. power meter covering the same range; direct-reading absorption cavity wavemeters (1,700–5,000 Mc/s) and a crystal voltmeter with two measurement ranges—0·5 and 1 volt full-scale—which can be used at any frequency between 1 and 500 Mc/s.

Marconi Instruments Ltd.,
St. Albans, Herts.

Miniature Precision Potentiometers

This type 11 potentiometer, which is manufactured as a single unit, has an overall length excluding shaft of only 0·78 in. and a diameter of 1 in. Ganged potentiometers with up to three units are standard, the overall length, excluding shaft, being increased by 0·281 in. per



gang, the diameter remaining constant at 1 in.

The resistance range is 1 kΩ to 50 kΩ and resolutions of between 2 and 6 turns per degree can be achieved.

The single unit can be supplied with a starting torque of less than 0·5 gm cm and reliable noise-free operation is ensured by the use of double wiping contacts.

Ferranti Ltd.,
Ferry Road, Edinburgh, 5.

Cadmium-Sulphide Photocell

A new type of photo-conductive cell which is claimed to make possible a reduction in the cost of many industrial-control and detection devices will shortly be available in quantity.

The new cell incorporates a specially constructed photo-sensitive element of cadmium sulphide and has extremely high sensitivity. It is said to produce sufficient current when used with weak light sources to operate a large relay directly.

The cell will operate from a low applied voltage. This is made possible by a special form of construction in which the resistance of the cadmium-sulphide element is effectively reduced by an interdigital pattern of copper strips. The cell is mounted on a standard valve base.

Typical performance figures for a cell with a photo-cathode of 1·8 square cm effective area are as follows:

From an illumination of 5 ft.-lamberts, with a colour temperature of 1,500°K, the cell will produce approximately 20 mA of current for an applied voltage of 10 V. From the same illumination, but with a temperature of 2,700°K, the current is approximately 6 mA. Within the limits of permissible power dissipation, doubling the applied voltage gives a four-fold increase in current. Maximum dissipation is 1 W at 25°C and 200 mW at 75°C.

Dark current is extremely small; with 300 V applied to the cell, it is not greater than 2·5 μA at 25°C.

Spectral response range is 4,500 Å to 8,000 Å, covering the visible spectrum and extending into the near infra-red, with maximum response in the yellow/red region.

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

Scintillation Counter

A new scintillation counter, type N612, has been designed for work involving low-energy beta-emitting isotopes.

It is said to provide an easier means of counting, with a higher efficiency than has been obtainable hitherto, without the necessity for lengthy and tedious chemical preparations.

In cases where the sample is soluble in the aromatic hydrocarbon liquids, it may usually be 'counted' simply by adding it directly to the liquid phosphor. Where the sample is in the form of water solution, it can be blended with the phosphor by the addition of absolute ethyl alcohol.

The built-in amplifier has an upper frequency response extending to beyond

1 Mc/s and has switched gains of 100, 250, 500 and 1,000. By adjusting a simple link the gain may be reduced by a factor of 10. The amplifier has a large amount of negative feedback applied to stabilize the gain and linearize the input–output characteristic. A cathode-follower is fitted both at the input and the output; that at the input reduces the capacitive loading on the photo-multiplier collector, and the one at the output will feed up to 2 metres of coaxial cable without worsening the frequency response.

This counter is stated to provide an efficient and simple means of measuring the two isotopes, H3 and C14, which are of great interest in medicine, biology and botany.

Ekco Electronics Ltd.,
Ekco Works, Malmesbury, Wilts.

Abstracts and References

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

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

ACOUSTICS AND AUDIO FREQUENCIES

534.613 **1611**

Radiation Force on Bodies in a Sound Field.—H. Olsen, W. Romberg & H. Wergeland. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 69–76.)

534.75 **1612**

Proposed Laboratory Standard of Normal Hearing.—J. F. Corso. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 14–23.) Report and discussion of laboratory measurements to determine the normal monaural threshold of hearing for pure tones.

534.782 **1613**

Sound Synthesizer with Optical Control.—O. Fujimura. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 56–57.) Brief description of a system based on a bank of ADP crystals, polarized light beams and a photocell. See also 1822 below.

534.84 **1614**

Applications of the Monte Carlo Method to Architectural Acoustics.—J. C. Allred & A. Newhouse. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 1–3.) "The Monte Carlo method of numerical analysis is applied to the determination of mean free paths and acoustic weighting factors relating probability of collision with walls in rectangular parallelopipeds. The method of calculation is discussed and its extension to determination of reverberation times in coupled rooms and auditoria is evaluated."

534.84 : 534.62 **1615**

Output of a Sound Source in a Reverberation Chamber and Other Reflecting Environments.—R. V. Waterhouse. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 4–13.) Expressions are derived for the sound power output for various source/reflector systems of interest in architectural acoustics, including the case of a dipole source near a reflecting plane, and a simple source near a reflecting edge and corner. In general the power output differs significantly from the free-field value if the distance of the source from the reflector is less than one wavelength. See also 2513 of 1955.

621.395.61 + 621.395.62 **1616**

Dynamic Mechanical Stability in the Variable-Reluctance and Electrostatic Transducers.—C. H. Sherman. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 48–55.) An analysis of the stability of these types of transducer for high-power applications. The limiting displacements for typical values of mechanical Q, polarization and frequency are shown.

621.395.625.3 : 778.5 **1617**

Further Data on Infrared Transparency of Magnetic Tracks.—G. Lewin. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1957, Vol. 66, No. 12, pp. 760–763. Discussion, p. 763.) The dependence of the transmittance upon the thickness of the magnetic oxide deposit, and the intermodulation effects on a converted 35-mm reproducer are examined.

AERIALS AND TRANSMISSION LINES

621.372.2.011.1 : 518.4 **1618**

Transmission-Line Calculator.—J. H. Andreae. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 191–193.) A graphical device for finding the length of terminated lines at specific frequencies is described. It has been found useful for calculations on low-loss resonant lines used in ultrasonic research above 50 Mc/s.

621.372.2.011.21 : 517.54 **1619**

The Calculation of Characteristic Impedance by Conformal Transformation.—J. C. Anderson. (*J. Brit. Instn Radio Engrs*, Jan. 1958, Vol. 18, No. 1, pp. 49–54.) The theory is applied to a line with cylindrical outer and strip inner conductors.

621.372.21 **1620**

Calculation of a Lossy Line.—V. S. Mel'nikov. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 28–30.) Analysis for the case where energy absorption per unit length is constant along the line.

621.372.821 : 621.3.049.75 **1621**

The Application of Printed-Circuit Techniques to the Design of Microwave Components.—J. M. C. Dukes. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 155–172. Discussion, pp. 180–181.) The basic theory of strip transmission lines is reviewed and the relative merits of the techniques and materials used are discussed. An adjustable short-circuit and a wide-band precision attenuator are described, together with other components for the band 2 500–4 300 Mc/s. Details are given of an r.f. head for two-way communication equipment.

621.372.821 : 621.3.049.75 **1622**

Developments in Printed Microwave Components.—D. R. J. White. (*Electronic Ind. Tele-Tech*, Nov. 1957, Vol. 16, No. 11, pp. 63–66. . 146.) A general description is given of the mechanical construction and electrical properties of high-Q microwave transmission lines used for connecting printed components.

621.372.821 : 621.372.832.43 **1623**

Broad-Band Slot-Coupled Microstrip Directional Couplers.—J. M. C. Dukes. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 147–154. Discussion, pp. 180–181.) The two strip transmission lines are mounted back to back and coupled through slots in the common ground plane. A 3-dB coupler with 50 transverse slots is described. By grading the phase velocity in one line, a power split equal within 1 dB over the band 2 800–4 300 Mc/s is achieved.

621.372.821 : 621.372.852.1 **1624**

Re-entrant Transmission-Line Filter using Printed Conductors.—J. M. C. Dukes. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 173–179. Discussion, pp. 180–181.) A novel procedure is described for the design of microwave low-pass filters having a relatively high stop-band insertion loss over a 3 : 1 bandwidth. The filters can be produced in microstrip or triplate line using printed circuit techniques.

621.372.83 : 621.372.6 **1625**

Scattering Equivalent Circuits for Common Symmetrical Junctions.—W. K. Kahn. (*Trans. Inst. Radio Engrs*, June 1956, Vol. CT-3, No. 2, pp. 121–127.) Abstract, *Proc. Inst. Radio Engrs*, Oct. 1956, Vol. 44, No. 10, p. 1492.)

621.372.831 **1626**

Design of a Conical Taper in Circular Waveguide System Supporting H_{01} Mode.—L. Solymar. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 618–619.)

621.372.852.1 **1627**

Wide-Band Waveguide Filters with Short Linear Tapers.—G. Craven. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 210–212.) Details are given of broad-band filters which are matched to standard waveguides. In a typical example, the amplitude of the variations in the transmission loss in the band 3 800–4 200 Mc/s is not more than 0·25 dB, with a voltage s.w.r. less than 1·6.

621.372.852.323 : 621.318.134 **1628**

Field-Displacement Isolator in Microwave Communications.—S. Weisbaum & H. Boyet. (*Bell Lab. Rec.*, Nov. 1957, Vol. 35, No. 11, pp. 456–461.) A general summary is given of the theory and uses of ferrite slabs in rectangular waveguides, in which a d.c. magnetic field is applied to the ferrite at right angles to the direction of wave propagation. Reflected waves are dissipated greatly in the ferrite, but forward-going waves only slightly. Typical performance figures in the 10·7–11·7-kMc/s band are: forward loss 1 dB, reverse loss 70 dB, with H = 1 045 oersteds.

621.396.67 : 537.226 **1629**

Polar Diagrams of Surface-Wave Aerials.—K. I. Grinëva. (*Radiotekhnika, Mosk.*, Dec. 1956, Vol. 11, No. 12, pp. 3–14.) An infinite metal surface covered with a thin layer of a dielectric is considered, over which a surface wave is propagated. The polar diagram of the surface-wave aerial is calculated by an approximate method based on Kirchhoff's formula. The effects of attenuation in the dielectric and of aerial length and phase velocity on the polar diagram are discussed. Conclusions are compared with experimental results and a procedure for the design of the aerial is proposed.

621.396.67 : 621.396.712 **1630**

The Planned Medium-Wave and Television Aerial of the N.D.R. at Flensburg.—E. Mohr. (*Rundfunktech. Mitt.* Aug. 1957, Vol. 1, No. 4, pp. 139–142.) Details of an aerial installation erected by the Norddeutsche Rundfunk, which consists of a mast 205 m high divided into two sections by an insulator at a height of 110 m. It incorporates television and v.h.f. transmitting aerials as well as serving as a mast radiator for medium waves in the range 500–1 600 kc/s.

621.396.674.3 : 621.396.11 **1631**

Excitation of Surface Waves on Conducting, Stratified, Dielectric-Clad, and Corrugated Surfaces.—(See 1852.)

621.396.677 **1632**

Antenna Applications in Two-Way Radio Systems.—T. J. McMullin. (*Radio TV News*, Dec. 1957, Vol. 58, No. 6, pp. 37–40.) Practical information is given on the performance of directive aerials in the wave bands for mobile use, 25–50 Mc/s, 72–76 Mc/s and 148–174 Mc/s.

621.396.677.3 : 621.372.51 **1633**

Three Antennas on One Lead.—C. Woodard. (*Radio TV News*, Dec. 1957, Vol. 58, No. 6, pp. 48–49, 157.) Discussion of a method of connecting three Yagi aerials operating on channels 6, 8 and 13, to one lead.

621.396.677.43 **1634**

Investigations of the Large Rhombic Aerial at the Overseas Receiving Station Eschborn.—W. Kronjäger, E. Mark & K. Vogt. (*Nachrichtentech. Z.*, Aug. 1957, Vol. 10, No. 8, pp. 382–384.) Report of comparative tests carried out at 5, 7·5 and 10 Mc/s with the Eschborn aerial which has sides of length 300 m. The advantages over aerials of normal size include a relative gain of about 15 dB at the lower frequencies.

621.396.677.832 : 621.396.96 **1635**

V-Reflex Aerial for an Information Radar Station.—G. von Trentini & E. J. Kirkscether. (*Rev. telegr. Electrónica, Buenos Aires*, Nov. 1957, Vol. 46, No. 541, pp. 637–641.) A discussion of the design and application of V-shaped reflectors with a partially reflecting surface across the mouth of the V; their advantage over curved reflectors lies in greater simplicity and consequent cheapness. Tests on a scale model working at 32 cm λ and coastal installations for operation at 144 cm λ are described. See also *Trans. Inst. Radio Engrs*, Oct. 1956, Vol. AP-4, No. 4, pp. 666–671 (von Trentini).

621.396.677.833 : 523.16 **1636**

The 45-ft Radio Telescope at the Royal Radar Establishment, Malvern.—(*Nature, Lond.*, 7th Dec. 1957, Vol. 180, No. 4597, pp. 1225–1228.) The parabolic reflector has a focal length of 20 ft and may be elevated between 5° and 85° at all azimuths. The telescope may be used to detect galactic radiation at 91 cm λ, or as a radar transmitter and receiver. Radar echoes from the first Russian earth-satellite rocket and also from the moon have been recorded.

AUTOMATIC COMPUTERS

681.142 **1637**

Basic Magnetic Logic Circuits in Computers.—K. Ganzhorn. (*Elektronische Rundschau*, Aug. 1957, Vol. 11, No. 8, pp. 229–234.) Classification of fundamental circuits which can be realized by means of square-loop ferromagnetic cores.

681.142 **1638**

The Design of the Control Unit of an Electronic Digital Computer.—M. V. Wilkes, W. Renwick & D. J. Wheeler. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 121–128. Discussion, pp. 144–146.) The design of the control or sequencing unit is discussed. In one system the order code is determined by the arrangement of the diodes in a diode matrix, and in another by the threading of wires through a matrix of ferrite cores.

681.142 : 537.311.33 : 538.632 **1639**

Analogue Multiplier Based on the Hall Effect.—L. Löfgren. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 158–166.) Discussion of the application of the Hall effect for electronic multiplication of voltages. An investigation of semiconductors shows Si to be the best for accurate multiplication. The choice of crystal dimensions is discussed with reference to the solution of a potential problem with skew boundary conditions which determines the Hall voltage.

681.142 : 621.3.087.6 : 535.376 **1640**

An Accurate Electroluminescent Graphical-Output Unit for a Digital Computer.—T. Kilburn, G. R. Hoffman & R. E. Hayes. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 136–144. Discussion, pp. 144–146.) The unit is fabricated from a uniform layer of electroluminescent phosphor with 512 parallel conducting strips on either side, these sets of strips being at right angles to form a matrix of conductors. A changing electric field between selected strips of both sets causes fluorescence at their intersection. Intersections are selected in turn and the pattern is recorded photographically. Details of some protototype matrices are given.

681.142 : 621.316.86 **1641**

The Design of Function Generators using Silicon Carbide Nonlinear Resistors.—E. Brown & P. M. Walker. (*Electronic Engng*, March 1958, Vol. 30, No. 361, pp. 154–157.) The effective current/voltage characteristics of SiC resistors are modified by linear resistors placed in series and parallel. Such modified resistors are then used as input or feedback components in amplifiers to give nonlinear functions. Design details are given of square-law and sine function generators using this device.

681.142 : 621.318.57 : 538.221 **1642**

Magnetic Switching Circuits for the Representation of Logical Relations.—H. Gillert. (*Nachrichtentech. Z.*, Aug. 1957, Vol. 10, No. 8, pp. 391–402.) A design method is derived for obtaining the number of cores and the number of turns required on each core for a given switching function. Results of a systematic investigation are given for switching functions with two and three variables.

681.142 : 621.318.57 : 621.314.7 **1643**

A New Bistable Element Suitable for Use in Digital Computers: Parts 1 & 2.—C. D. Florida. (*Electronic Engng*, Feb. & March 1958, Vol. 30, Nos. 360 & 361, pp. 71–77 & 148–153.) The development of a trigger circuit using *p-n-p* and *n-p-n* transistors is described. Switch-on and switch-off times of 0·2 μs are achieved with a high current carrying capacity enabling several other circuits to be driven from it. The transient response of the circuit to a voltage ramp input at either the turn-on or turn-off terminals is analysed mathematically. The results thus obtained, assuming certain operating conditions, show good agreement with experimental observations.

681.142 : 621.374.32 **1644**

A Decimal Adder using a Stored Addition Table.—M. A. Maclean & D. Aspinall. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 129–135. Discussion, pp. 144–146.) A description of a serial decimal adder which accepts numbers in binary-coded form. The binary digits are decoded into a set of pulses which actuate a built-in addition table storing all the possible sums.

681.142.001.4 **1645**

Word Generator for Digital Testing.—R. R. Hartel. (*Electronics*, 28th Feb. 1958, Vol. 31, No. 9, p. 71.) "A beam-switching tube supplies arbitrary nine-bit words at pulse rates from a few c/s to 1 Mc/s for testing and evaluating digital systems. Pulse shape can be varied from spike to square wave by changing plug-in capacitors."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.3 : 621.376.32.029.3 : 621.385.1 **1646**

The Reactance Valve at Audio Frequencies.—B. J. Alcock. (*Electronic Engng*, Feb. 1958, Vol. 30, No. 360, pp. 86–88.) "The reactance-valve circuit is analysed to derive circuits producing an inductive impedance at a.f. with particular reference to obtaining a large Q factor."

621.3.012 : [621.3.015.3+621.3.018.4 **1647**

Approximate Relations between Transient and Frequency Response.—H. H. Rosenbrock. (*J. Brit. Instn Radio Engrs*, Jan. 1958, Vol. 18, No. 1, pp. 57–64.) An existing graphical method for deriving the transient response of a linear network from the sinusoidal frequency response and vice versa is extended to cover certain difficult cases. Typical examples are illustrated.

621.3.049.75 : 621.372.821 **1648**

The Application of Printed-Circuit Techniques to the Design of Microwave Components.—Dukes. (See 1621.)

621.3.049.75 : 621.372.821 **1649**

Developments in Printed Microwave Components.—White. (See 1622.)

621.3.072.6 : 621.372.5 **1650**

The Locking Band for the Automatic Phase Control of Frequency.—M. V. Kapranov. (*Radiotekhnika, Mosk.*, Dec. 1956, Vol. 11, No. 12, pp. 37–52.) Formulae are derived relating the locking band to the filter parameters and the delay time of the circuits.

621.314.2 : 621.373.431.2 **1651**

Designing Transformers for Blocking Oscillators.—R. D. McCartney. (*Electronics*, 28th Feb. 1958, Vol. 31, No. 9, pp. 78–80.) Design procedure, based on four common circuits, is given; the pulse shape is calculable, and a 6 : 1 size reduction over previous units is obtained with a Type-L-1 core.

621.314.6 **1652**

An Extended General Network Theorem on Rectification.—H. Stockman. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 615–616.) Extension of Gewartowski's theorem (48 of January).

621.318.435.3.042.143 **1653**

A New Type of Construction for High-Grade Transductor Cores.—U. Krabbe & G. H. Giesenhagen. (*Elektrotech. Z, Edn A*, 1st Oct. 1957, Vol. 78, No. 19, pp. 712–716.) Improvements of laminated-core design and assembly are discussed which permit a better utilization of the qualities of the material.

621.318.57 : 621.373.431 **1654**

Calculation of the Duration of a Quasi-Equilibrium State in a Phantastron Circuit.—G. I. Perov. (*Radiotekhnika, Mosk.*, Dec. 1956, Vol. 11, No. 12, pp. 61–74.) A formula is derived, and typical operating conditions of phantastron circuits are considered.

621.318.57 : 621.387 : 621.314.7 **1655**

Transistor Circuits for Use with Gas-Filled Multicathode Counter Valves.—J. B. Warman & D. M. Bibb. (*Electronic Engng*, March, 1958, Vol. 30, No. 361, pp. 136–139.) Transistors are used in conjunction with dekatrons to provide a logical circuit; the combination is run off a low-voltage supply, the h.v. to the dekatrons being supplied by a transistor d.c. converter. A block of these circuits is made to function successfully in a telephone exchange register.

621.319.4 **1656**

Ceramic Capacitors—a Complete Substitute for Paper and Mica Capacitors.—C. V. Ganapathy, R. Krishnan & T. V. Ramamurti. (*J. Instn Telecommun. Engrs, India*, Dec. 1957, Vol. 4, No. 1, pp. 2–11.) The electrical characteristics of paper, mica and ceramic capacitors are compared. Performance tests on receivers before and after substituting ceramic for paper capacitors show no significant difference.

621.372.4 **1657**

The Sense of Rotation and Curvature of the Impedance Loci of Real and Ideal Two-Poles.—H. Wolter. (*Arch. elekt. Übertragung*, Sept. 1957, Vol. 11, No. 9, pp. 359–365.) The impedance of a two-terminal network, as represented on the Gauss number sphere, describes a closed curve revolving completely, at least once, in a clockwise direction for a frequency variation from $-\infty$ to $+\infty$. A method is given for synthesizing two-poles which have impedance loci turning anticlockwise in places.

621.372.414 **1658**

Design of Complex Resonators.—A. I. Zhivotovskiĭ. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 22–27.) An investigation of resonators consisting of several uniform coaxial-line sections with different characteristic impedances.

621.372.414 : 621.372.8 **1659**

Microwave High-Power Simulator.—H. Heins. (*Electronic Ind. Tele-Tech.*, Nov. 1957, Vol. 16, No. 11, pp. 78–81, 155.) "By periodically injecting energy from a directional coupler into a waveguide cavity made in the form of a closed loop much higher power levels can be achieved than are available from microwave power generators. Energy is stored in a circulating or travelling wave." See also 1325 of 1956 (Sferrazza).

621.372.5/.6 **1660**

The Scattering Matrix in Network Theory.—H. J. Carlin. (*Trans. Inst. Radio Engrs*, June 1956, Vol. CT-3, No. 2, pp. 88–97.) Abstract, *Proc. Inst. Radio Engrs*, Oct. 1956, Vol. 44, No. 10, pp. 1491–1492.)

621.372.54 **1661**

Frequency Transformations in Filter Design.—A. Papoulis. (*Trans. Inst. Radio*

Engrs, June 1956, Vol. CT-3, No. 2, pp. 140–144.) Abstract, *Proc. Inst. Radio Engrs*, Oct. 1956, Vol. 44, No. 10, p. 1492.)

621.372.54 **1662**
Optimum Filters with Monotonic Response.—A. Papoulis. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 606–609.) The amplitude characteristic has no ripple in the pass band and a high rate of attenuation in the stop band and combines the desirable features of the Butterworth and Tchebycheff response.

621.372.54 **1663**
A Twin-T Variable-Slope Filter.—G. B. Miller. (*Electronic Engng*, March 1958, Vol. 30, No. 361, pp. 143–145.) "By suitable modification a twin-T notch filter is converted into a low-pass filter in which the rate of attenuation up to a specified upper frequency is controlled by a single potentiometer. Design data are given and a detailed circuit is described."

621.372.54 : 621.315.212 **1664**
Filters, Built from Coaxial Conductors.—T. J. Weijers. (*Philips Telecommun. Rev.*, Nov. 1957, Vol. 18, No. 4, pp. 186–206 & Jan. 1958, Vol. 19, No. 1, pp. 23–54.) Zobel's method of filter design is applied to coaxial-line filters considering the impedance of the line as a function of frequency. Characteristics can then be determined for the entire frequency range. Design data are shown in a series of tables, and four types of filter section are discussed in detail.

621.372.542.2 **1665**
A Filter with Characteristics Approximating to those of an Ideal Low-Pass Filter.—J. Remer. (*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 151–158.) The design and construction of a filter are described whose characteristics conform closely to those of the ideal low-pass filter specified by Küpfmüller. The variations of its amplitude response up to cut-off at 5·5 kc/s are less than $\pm 0{\cdot}5\%$. At cut-off the response is 9% of its maximum value, and the fluctuations of group delay are less than 0·025 ms.

621.372.543 : 621.375.126 **1666**
***RC* and *LC* Resonant Filters and their Application in Selective Amplifiers.**—H. H. Rabben. (*Elektronische Rundschau*, Sept. & Oct. 1957, Vol. 11, Nos. 9 & 10, pp. 265–268 & 314–318.) The design of a narrow-band amplifier with a centre frequency of about 200 c/s is discussed following a comparison of circuits using different types of filter elements. A two-stage circuit with negative feedback and incorporating a *RC* band-stop filter was the design adopted for use in the measurement of the solar magnetic field.

621.372.552 **1667**
An Approach to the Design of Constant-Resistance Amplitude Equalizer Networks.—J. S. Bell. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 185–189.) A method of designing constant-resistance amplitude equalizers to give a desired frequency response characteristic over a given range is suggested. A typical example illustrates the method of adjusting the response of a velocity-type pick-up over part of the a.f. spectrum.

621.372.57 : 621.314.7 **1668**
A Treatment of Cascaded Active Four-Terminal Networks, with Application to Transistor Circuits.—H. L. Armstrong. (*Trans. Inst. Radio Engrs*, June 1956, Vol. CT-3, No. 2, pp. 138–140.) Abstract, *Proc. Inst. Radio Engrs*, Oct. 1956, Vol. 44, No. 10, p. 1492.)

621.373.42 **1669**
Self-Oscillator with Large Circuit Attenuation.—A. Z. Khaĭkov. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 63–72.) The optimum operating conditions for power transfer to the load and the efficiency of the oscillator are discussed. Equations are derived and graphs show the grid voltage variations for different values of attenuation.

621.373.42.029.42 **1670**
Amplitude-Stabilized Low-Frequency Oscillator.—A. K. Choudhury & B. R. Nag. (*J. Instn Telecommun. Engrs, India*, Dec. 1957, Vol. 4, No. 1, pp. 36–45.) The performance of a v.l.f. oscillator for the range 0·01—10 c/s with the output stabilized by biased diodes or lamps, is analysed. Formulae for the harmonic content with different initial damping are deduced and means of reducing it are indicated. A circuit for measuring the harmonic content is described.

621.373.421.029.62/.63 **1671**
Double-Tetrode Oscillator.—J. H. Andreae & P. L. Joyce. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 173–177.) The oscillator covers a frequency range of 150–500 Mc/s and can be anode modulated with 3-μs 3-kV pulses repeated at a rate of about 300/s. Constructional details are given.

621.373.421.13 **1672**
Crystal Oscillator has Variable Frequency.—G. A. Gedney & G. M. Davidson. (*Electronics*, 14th Feb. 1958, Vol. 31, No. 7, pp. 118–119.) A maximum frequency deviation of 5 c/s is obtained from a two-stage crystal feedback amplifier operating at 9·1 kc/s with a long-term frequency stability within a few parts per million.

621.373.421.13 **1673**
Frequency Stabilization of U.S.W. Oscillators by using Harmonics for the Excitation of the Quartz Crystal.—M. M. Pruzhanskiĭ. (*Radiotekhnika, Mosk.*, Dec. 1956, Vol. 11, No. 12, pp. 15–27.) Modern methods of direct stabilization of u.s.w. oscillators by quartz crystals are reviewed. Circuits are described in which the static capacitance of the crystal is compensated by inductance. Simple bridge circuits for operation over a range of frequencies are also described, in which the crystal is excited by high-order harmonics. Uncompensated circuits are also considered, and comparison is made between the various types of circuit.

621.373.431.1 **1674**
A Three-Phase Three-Valve Multivibrator.—W. F. Lovering. (*Electronic Engng*, Feb. 1958, Vol. 30, No. 360, pp. 94–95.) A three-valve circuit in which each anode is capacity coupled to the other two grids is described. Frequencies up to 50 kc/s are obtained using miniature r.f. pentodes.

621.373.44 **1675**
Generation of Bell-Shaped Pulses.—L. I. Kastal'skiĭ. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 73–75.) A pulse generating circuit is described for producing pulses 2·5 to 8 ms long at a repetition frequency of 25 kc/s.

621.373.44 : 621.314.7 **1676**
Transients in Pulse Circuits with Point-Contact Transistors.—O. G. Yagodin. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 43–57.) Analysis based on the dynamic characteristics of transistors. Transient processes in relaxation oscillators and triggering circuits are examined.

621.373.5 **1677**
A Carrier-Energized Bistable Circuit using Variable-Capacitance Diodes.—E. O. Keizer. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 475–485.) A variable-capacitance junction diode [see 2559 of 1956 (Giacoletto & O'Connell)], when used in a simple circuit driven from a high-frequency source, can cause that circuit to have a bistable characteristic suitable for dynamic storage, or to have a sensitive output/input characteristic suitable for control or detection purposes.

621.374.3 **1678**
Adjustable Electronic Delay Circuit for the Microsecond Range.—J. F. Vervier & P. C. Macq. (*J. Phys. Radium*, Oct. 1957, Vol. 18, No. 10, p. 603.) A modification of a pulse discriminator circuit of a type described by Moody et al. (2730 of 1952) using Type-EFP60 secondary-emission pentodes is described. By means of an adjustable capacitance coupling dynode and grid, the time interval between successive negative and positive pulses derived in the circuit can be adjusted between 2×10^{-8} and 2×10^{-5} s. For an application of the method in a delayed coincidence circuit, see *Rev. sci. Instrum.*, Oct. 1957, Vol. 28, No. 10, pp. 843–844 (Macq & Vervier).

621.374.32 **1679**
Binary Frequency Divider with Junction Transistors.—A. R. Luminskiĭ & N. M. Trakhtenberg. (*Elektrosvyaz'*, April 1957, No. 4, pp. 33–39.)

621.375.13 **1680**
Limited-Gain Operational Amplifiers.—A. W. Keen. (*Electronic Radio Engr*, April 1958, Vol. 35, No. 4, pp. 141–143.) The effect of finite gain can be allowed for by assuming infinite amplifier gain and then adding fictitious elements to the feedback network to reduce the gain to its actual value. This leads to a more convenient equivalent network, of which simple examples are given.

621.375.13 : 621.372.54 **1681**
Parallel-T *RC* Selective Amplifiers.—J. J. Ward & P. V. Landshoff. (*Electronic Radio Engr*, April 1958, Vol. 35, No. 4, pp. 120–124.) The operation of a less familiar

form of selective amplifier, with low-impedance input, is analysed, in which the signal is injected at the null point of the parallel-T network in the feedback loop. Design equations are derived and applied as an example to a 50-c/s fixed-tuned amplifier with a 10-c/s square-wave input.

621.375.2.029.3 : 621.372.552 **1682**

Electronic Equalizer.—S. Subramanian. (*J. Instn Telecommun. Engrs, India*, Dec. 1957, Vol. 4, No. 1, pp. 12–17.) Description of an a.f. amplifier with variable boost or attenuation at both the low- and high-frequency ends of the response characteristic. Typical response curves of the amplifier are shown.

621.375.2.029.3 : 621.396.82 **1683**

Hum in Audio Stages.—R. S. Babbs & M. E. Mason. (*Mullard tech. Commun.*, Dec. 1957, Vol. 3, No. 27, pp. 209–213.) A design procedure, based on measurements on typical valve types, for reducing hum in the audio stages of a.c./d.c. equipment.

621.375.23 **1684**

Design of Feedback Amplifiers for Prescribed Closed-Loop Characteristics.—J. L. Stewart. (*Trans. Inst. Radio Engrs*, June 1956, Vol. CT-3, No. 2, pp. 145–151.) Abstract, *Proc. Inst. Radio Engrs*, Oct. 1956, Vol. 44, No. 10, p. 1492.)

621.375.23 : 621.396.621.55 **1685**

Modified Rice Neutralization.—B. C. Das. (*J. Instn Telecommun. Engrs, India*, Dec. 1957, Vol. 4, No. 1, pp. 46–48.) The theory of Rice and modified Rice neutralization are discussed and formulae are derived for maximum stability with no feedback voltage on the grid. The conditions for positive and negative feedback are also derived.

621.375.3 : 621.318.435 **1686**

Subminiature Magnetic Amplifiers.—A. H. Argabrite. (*Radio TV News*, Dec. 1957, Vol. 58, No. 6, pp. 70–71, 175.) A description of circuits using a new 'ferristor' fast saturable reactor which may be used as a magnetic amplifier, or as a bistable ferro-resonant element.

621.375.4 **1687**

Transistor Bias Circuits.—R. P. Murray. (*Electronic Ind. Tele-Tech*, Nov. 1957, Vol. 16, No. 11, pp. 75–77..148.) A detailed comparison of four bias circuits with reference to performance and the operating-point stability. A practical stability factor is derived together with general design formulae for bias components.

621.375.4.001.2 **1688**

The Use of Universal Curves in the Design of Transistor Amplifier Stages.—E. De Castro. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, Supplement to No. 3, pp. 1–24.) Design curves are derived from the principal transistor parameters for linear operation at low frequencies.

621.375.4.024 **1689**

D.C. Transistor Amplifier for High-Impedance Input.—D. Schuster. (*Electronics*, 28th Feb. 1958, Vol. 31, No. 9, pp. 64–65.) A short discussion and circuit diagram of an amplifier using a double-emitter follower and grounded-emitter voltage amplifier. Temperature compensation gives short-term drift stability. The input impedance is 0·4 MΩ, and the current gain 1 000.

621.375.9 : 538.221 : 538.569.4.029.6 **1690**

Theory of Parametric Amplification using Nonlinear Reactances.—S. Bloom & K. K. N. Chang. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 578–593.) "The parametric amplifier is analysed phenomenologically in terms of an equivalent-circuit model. The model consists of a signal circuit resonant at ω_1, an idling circuit at ω_2, and a pumping circuit at $\omega_3=\omega_1+\omega_2$, these three circuits being coupled across a nonlinear inductance. The analysis is general enough to delineate the conditions on the signal level and circuit parameters which lead to distortionless amplification. Expressions are derived for power gain, bandwidth, and noise factor for the case in which the signal and idling frequencies are well separated and for the degenerate case in which these two frequencies are equal."

621.375.9 : 538.569.4.029.6 **1691**

Solid-State Maser Amplifier.—A. L. McWhorter & J. W. Meyer. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 312–318.) The operation of a solid-state maser amplifier at 2 800 Mc/s is described. A dual-frequency cavity containing paramagnetic potassium chromicyanide in an isomorphous cobalt diluent is used at 1·25°K. The upper three of the four energy levels of the Cr+++ ion are used. Spin state populations are inverted by saturating the resonance absorption at 9 400 Mc/s. The experimental observations of the maser, both as an amplifier and as an oscillator are compared with theory.

621.375.9 : 538.569.4.029.63/.64 **1692**

New Approaches to the Amplification of Microwaves.—J. P. Wittke. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 441–457.) The basic principles governing the operation of two new types of 'molecular' microwave amplifier—the maser and the parametric amplifier—are described. Both types of amplifier have relatively narrow bandwidths, but excellent noise properties.

621.376.239 : 621.318.134 **1693**

Ferrite Microwave Detector.—D. Jaffe, J. C. Cacheris & N. Karayianis. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 594–601.) It is shown theoretically that second-order terms in one component of the magnetization of a ferrite under the action of a r.f. field may be used to produce magnetostriction in the ferrite and so to detect an a.m. microwave signal. An experimental technique is described in which the detector consists of a long thin ferrite rod in a waveguide. Magnetostriction vibrations are observed by means of a polarized $BaTiO_3$ ceramic rod bonded to the ferrite. The qualitative results agree with theory, and factors which are expected to improve performance are discussed.

621.376.332 : 621.396.822 **1694**

Noise Output of Balanced Frequency Discriminator.—D. Slepian. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, p. 614.) A mathematical analysis for Gaussian noise input, assuming no limiter action.

621.376.54 : 621.314.7 **1695**

A Conductivity-Storage Transistor Pulse-Width Modulator.—J. C. Price. (*Electronic Engng*, Feb. 1958, Vol. 30, No. 360, pp. 88–90.) "Transistors of both point-contact and junction types exhibit in certain pulse applications, prolonged conduction generally ascribed to 'hole storage'. This can be controlled in a circuit to provide a simple means of pulse width modulation for time-division multiplex. The technique is described for use with selected point-contact transistors."

GENERAL PHYSICS

537.32 **1696**

Thermoelectric Effects.—F. E. Jaumot, Jr. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 538–554.) The basic principles of thermoelectricity are reviewed, recent achievements are outlined in terms of specific practical applications, and the present status of the more detailed theoretical treatments is discussed in a nonmathematical fashion. Useful equations describing important parameters are tabulated. Over 100 references.

537.533 **1697**

Optical Theory of Thermal Velocity Effects in Cylindrical Electron Beams.—G. Herrmann. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 127–136.) "The present theory is based on a nonlaminar optical model which treats thermal velocities as an integral part of the motion. A Maxwellian distribution of initial transverse velocities is assumed at the cathode, and a first-order focusing theory is applied in order to calculate trajectories at any point in the beam. It is shown that whenever a long beam is confined by a focusing field, images of the cathode are formed repeatedly along the axis. When applied to uniform magnetic focusing fields, the theory predicts the periodic formation along the axis of cathode images and crossovers, and a relative rotation of successive images. Such effects have been reported."

537.533 **1698**

Effect of Variation of D.C. Current in a Modulated Electron Beam.—I. P. Shkarofsky. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 222–223.) Experiments indicate that existing theory needs correction. An empirical formula satisfying all experimental observations is difficult to obtain.

537.56 : 538.56 **1699**

Oscillations in Plasma: Part 1.—S. Kojima, K. Kato & S. Hagiwara. (*J. phys. Soc. Japan*, Nov. 1957, Vol. 12, No. 11, pp. 1276–1281.) A sensitive super-regenerative detector was used to study oscillations in a Looney-Brown tube without auxiliary electron beams. Oscillations were detected by a small external aerial.

537.56 : 538.56 **1700**

Nonlinear Effects in Electron Plasmas.—P. A. Sturrock. (*Proc. roy. Soc. A*, 5th Nov. 1957, Vol. 242, No. 1230, pp. 277–299.) A mathematical study of three groups of effects, namely, the excitation of harmonics, coherent interaction, and incoherent interaction. Complications of finite boundaries, external fields, non-zero temperature, multistream flow, and collisions, are ignored. The main interest is in the incoherent interaction which results in spectral decay, and damping effects are compared with experimental data.

537.56 : 538.56 : 538.6 **1701**

Waves in a Plasma in a Magnetic Field.—I. B. Bernstein. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 10–21.) An analysis of the small-amplitude oscillations of a fully ionized, quasi-neutral plasma in a uniform, externally produced magnetic field. No self-excitation of waves around thermal equilibrium is predicted, in contrast to the results of Gordeyev. For longitudinal electron oscillations propagated perpendicularly to the constant magnetic field, there are gaps in the spectrum of allowed frequencies at multiples of the electron gyration frequency, but zero Landau damping. When the ion dynamics are included, two classes of low-frequency oscillations are found and the results for the propagation of e.m. waves in an ionized atmosphere are also derived.

537.56 : 538.566 **1702**

Conductivity of Plasmas to Microwaves.—H. Margenau. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 6–9.) The complex conductivity is calculated for a neutral plasma with electrons having a distribution in space and velocity which does not change over a time interval long compared with the period of the microwave field. The conductivity is derived for velocity distributions given by the Dirac δ function, a step function, and a Maxwellian distribution.

537.56 : 538.6 **1703**

Plasma Diffusion in a Magnetic Field.—M. N. Rosenbluth & A. N. Kaufman. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 1–5.) "The equations governing the diffusion of a fully ionized plasma across a magnetic field are derived. It is assumed that macroscopic quantities vary slowly across an ion radius of gyration, and that the interparticle collision frequency is much less than the gyration frequency. The relevant transport coefficients—electrical resistivity, thermal conductivity, and thermoelectric coefficient—are derived. Some similarity solutions of the equations are found."

538.3 **1704**

Nonlinear Electromagnetism and Photons in the Functional Theory of Particles.—F. Aeschlimann & J. L. Destouches. (*J. Phys. Radium*, Nov. 1957, Vol. 18, No. 11, pp. 632–637.)

538.566 : 535.43 **1705**

New Tables of Total Mie Scattering Coefficients for Spherical Particles of Real Refractive Indexes ($1 \cdot 33 \leqslant n \leqslant 1 \cdot 50$).—R. B. Penndorf. (*J. opt. Soc. Amer.*, Nov. 1957, Vol. 47, No. 11, pp. 1010–1015.) K, the total Mie scattering coefficient, is defined as the total flux scattered by the particle divided by the flux incident on its cross-section πr^2; the size parameter $\alpha = 2\,\pi\,r/\lambda$. The theory applies only to a very dilute aerosol of randomly arranged spheres and does not allow for coherence in the field or for multiple scattering. K has been calculated on an electronic computer for $0 \cdot 1 \leqslant \alpha \leqslant 30$ in steps of $0 \cdot 1$ and for $n = 1 \cdot 33$, $1 \cdot 40$, $1 \cdot 44$, $1 \cdot 486$, and $1 \cdot 50$.

538.566 : 535.43 **1706**

Scattering of Electromagnetic Waves by Long Cylinders.—A. W. Adey. (*Electronic Radio Engr*, April 1958, Vol. 35, No. 4, pp. 149–158.) The field scattered by a metal or dielectric cylinder, when excited by a wave propagated in a direction normal to the cylinder axis, is discussed theoretically for plane and cylindrical waves. The radius of the cylinder is comparable with the wavelength. The dielectric cylinder is a resonant structure.

538.569.3 **1707**

Propagation through a Dielectric Slab.—T. B. A. Senior. (*Electronic Radio Engr*, April 1958, Vol. 35, No. 4, pp. 135–137.) Propagation from a point source through an infinite slab is considered theoretically, with particular reference to the apparent source and its modified polar diagram. An appreciable change in polar diagram can be produced by even a thin sheet of dielectric.

538.569.4 : 535.34 **1708**

The Microwave Spectrum and Structure of Trichloracetonitrile.—J. G. Baker, D. R. Jenkins, C. N. Kenney & T. M. Sugden. (*Trans. Faraday Soc.*, Nov. 1957, Vol. 53, Part 11, pp. 1397–1401.) The microwave spectrum of $C\,Cl_3^{35}\,CNh$ as been studied in the range 16–27 kMc/s using a Stark modulation microwave spectrometer.

538.569.4 : 539.14 **1709**

Nuclear Magnetic Resonance.—M. Lipsicas. (*Brit. Commun. Electronics*, Nov. 1957, Vol. 4, No. 11, pp. 686–690.) The phenomenon is briefly discussed and the Pound system for its detection is described. For high-resolution work the Bloch spectrometer is particularly suitable.

538.569.4 : 539.14 **1710**

Audio-Frequency Nuclear-Resonance Echoes.—J. G. Powles & D. Cutler. (*Nature, Lond.*, 14th Dec. 1957, Vol. 180, No. 4598, pp. 1344–1345.) A spin-echo technique for determining nuclear magnetic resonance by experiments in the earth's magnetic field is described. It is possible to derive the true spin-spin relaxation time even if the field is not homogeneous.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 : 621.396.677.833 **1711**

The 45-ft Radio Telescope at the Royal Radar Establishment, Malvern.—(See 1636.)

523.164.3 + 551.594.6 **1712**

Radio Noise from Planets.—F. Horner. (*Nature, Lond.*, 7th Dec. 1957, Vol. 180, No. 4597, p. 1253.) A comparison is made of h.f. radio noise from Jupiter and Venus and from terrestrial lightning. The hypothesis that radio noise from the two planets is due to electrical discharges analogous to terrestrial lightning requires modification. See also 3357 of 1956 (Kraus).

523.164.32 **1713**

The Distribution of Radio Brightness over the Solar Disk at a Wavelength of 21 Centimetres: Part 4—The Slowly Varying Component.—W. N. Christiansen, J. A. Warburton & R. D. Davies. (*Aust. J. Phys.*, Dec. 1957, Vol. 10, No. 4, pp. 491–514.) The emitting regions were studied individually using a 32-element interferometer producing fringes 3′ of arc in width. During 1952–1953 the radio sources were found to lie about 22 000 km above and to have the same size as 'plages faculaires'. The angular distribution of radio flux follows approximately a cosine law suggesting that the source is a thin shell parallel to the sun's surface. The correlation between radio flux and sunspot areas is discussed. Part 3: 1706 of 1956 (Christiansen & Warburton).

523.53 : 621.396.96 **1714**

Meteor Radiant Determination from High-Echo-Rate Observations.—C. S. L. Keay. (*Aust. J. Phys.*, Dec. 1957, Vol. 10, No. 4, pp. 471–482.) "A simplified analysis is given of the Clegg method for delineating meteor radiants from radar observations [see 1031 of 1949]. A further analysis reveals a new and faster method of interpreting the data contained in meteor echo records. This method is applicable when sensitive equipment is employed and the resulting echo rate is very high."

523.75 : 523.164.32 **1715**

The Association of Solar Radio Bursts of Spectral Type III with Chromospheric Flares.—R. E. Loughhead, J. A. Roberts & M. K. McCabe. (*Aust. J. Phys.*, Dec. 1957, Vol. 10, No. 4, pp. 483–490.) Simultaneous optical and radio observations were made for over 300 flares 85% of which were of Class 1, and 20% of which are associated with Type-III events. 60% of the bursts were recorded during the life of the flare, usually near the beginning. The probability of a burst accompanying a flare is greater for large flares, and is the same for flares on the east limb as for those in the centre of the disk. This implies a wide cone of escape for Type-III radiation.

523.75 : 621.396.812.5.029.51 **1716**

A New Effect of Chromospheric Eruptions.—Waldmeier. (See 1865.)

550.389.2 **1717**

Bureau of Standards Role in I.G.Y.—(*Radio TV News*, Dec. 1957, Vol. 58, No. 6, p. 41.) A short summary of the work undertaken.

550.389.2 : 621.396.11 **1718**

Radio Research and the I.G.Y.—C. M. Minnis. (*Brit. Commun. Electronics*, Nov. 1957, Vol. 4, No. 11, pp. 691–693.)

The objectives and organization of the I.G.Y. are outlined. The programs of observations which are of particular interest to radio engineers and the users of rockets and satellites are described.

550.389.2 : 629.19 **1719**

Artificial Earth Satellites.—V. Vakhnin. (*QST*, Nov. 1957, Vol. 41, No. 11, pp. 22–24, 188.) English version of 3860 of 1957.

550.389.2 : 629.19 **1720**

Radio Observations on the Russian Satellites.—(*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 81–115.) The following short contributions were presented at a meeting of the Radio and Telecommunication Section of the I.E.E., London, 22nd November 1957 :—

(*a*) Observations at Cambridge.—J. R. Shakeshaft (pp. 83–84).

(*b*) Apparatus used at the Royal Aircraft Establishment.—A. N. Beresford (pp. 85–88).

(*c*) Some Direction-Finding Observations on the 20-Mc/s Signal.—F. A. Kitchen, E. R. Billam, W. R. R. Joy, R. F. Cleaver, D. L. Cooper-Jones & J. M. Beukers (pp. 89–91).

(*d*) Observations of Bearing and Angle of Elevation of Satellite I.—W. C. Bain & R. W. Meadows (pp. 91–93).

(*e*) Estimating the Height of the First Satellite from Radio Interferometer Records. —G. B. Longden (pp. 93–95).

(*f*) Precise Frequency Measurements on First Russian Satellite.—H. Stanesby (pp. 96–99).

(*g*) Analysis of Doppler Data from Earth Satellites.—D. E. Hampton (pp. 99–100).

(*h*) Radio Observations of the Signal Characteristics of Satellite I.—P. J. Brice & P. N. Parker (pp. 101–104).

(*i*) Radar Observations of the Russian Earth Satellites and Carrier Rocket.—J. Davis, J. V. Evans, S. Evans, J. S. Greenhow & J. E. Hall (pp. 105–107).

(*j*) Observations at the Royal Radar Establishment.—J. S. Hey (pp. 107–108).

Discussion (pp. 95–96, 108–115).

550.389.2 : 629.19 **1721**

Ranging the Satellite by Doppler-Shift Observation.—J. M. Osborne. (*Short Wave Mag.*, Nov. 1957, Vol. 15, No. 9, pp. 459–462.) An experimental method using simple apparatus.

550.389.2 : 629.19 **1722**

Unusual Propagation at 40 Mc/s from the U.S.S.R. Satellite.—H. W. Wells. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, p. 610.) Interferometry recordings were obtained in Washington on a few occasions during October 1957 when the satellite was on the opposite side of the earth, near an antinodal point.

550.389.2 : 629.19 **1723**

A Note on Some Signal Characteristics of Sputnik I.—J. D. Kraus & J. S. Albus. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 610–611.)

550.389.2 : 629.19 **1724**

Detection of Sputniks I and II by C.W. Reflection—J. D. Kraus. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 611–612.) The method described is based on the reception of the 20-Mc/s transmission of WWV at a distance of about 330 miles.

550.389.2 : 629.19 **1725**

The Last Days of Sputnik I.—J. D. Kraus. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 612–614.) Deductions are based on radio reflection records using WWV transmissions on 20 Mc/s. See also 1724 above.

551.510.535 **1726**

Proceedings of the Polar Atmosphere Symposium Held at Oslo, 2–8 July 1956: Part II—Ionospheric Section.—(*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, 212 pp.) The text is given, with ensuing discussions, of 21 papers presented at the symposium. Abstracts of some of these are given individually. Titles of others are as follows :—

(*a*) Results of Ionospheric Drift Measurements in the United States.—V. Agy (pp. 23–25).

(*b*) Results of Ionospheric Drift Measurements in Norway.—L. Harang (pp. 26–32).

(*c*) Movements of Ionospheric Irregularities Observed Simultaneously by Different Methods.—I. L. Jones, B. Landmark & C. S. G. K. Setty (pp. 41–43). See 2750 of 1957.

(*d*) Turbulence in the Ionosphere with Applications to Meteor Trails, Radio-Star Scintillations, Auroral Radar Echoes, and Other Phenomena.—H. G. Booker (pp. 52–81). See 1441 of 1957.

(*e*) Geographic Distribution of Geophysical Stations on the Polar Cap.—A. H. Shapley (p. 108).

(*f*) A Theory of Long-Duration Meteor Echoes based on Atmospheric Turbulence with Experimental Confirmation.—H.G. Booker & R. Cohen (pp. 171–194). See 1417 of 1957.

551.510.535 **1727**

Theoretical Views on Drift Measurements.—I. L. Jones. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 3–11.) The conversion of the amplitude-pattern drift at the ground, as sampled by three closely spaced receivers, to the true drift in the ionosphere requires the determination of auto- and cross-correlation functions. Problems in relating ground-pattern drift to ionospheric drift are stated.

551.510.535 **1728**

The Drift of an Ionized Layer in the Presence of the Geomagnetic Field.—K. Weekes. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 12–19.) The efficiency of air winds in causing ionization drift decreases with atmospheric pressure. Thus, in the F region, electric fields may be the main cause of drift ; in the E region electric fields may contribute equally with air winds ; in the D region air winds would be the main cause. The effect of winds and fields in causing drifting of cylindrical irregularities in each region is given.

551.510.535 **1729**

The Height Variation of Horizontal Drift Velocities in the E Region.—I. L. Jones. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 20–22.) The drift was measured on two adjacent frequencies corresponding to heights of reflection differing by about 5 km. The diurnal change in the N–S and E–W velocity components differs in phase at the two heights.

551.510.535 **1730**

Large-Scale Movements of the Layers.—H. W. Wells. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 33–40.) Moving wavelike disturbances, with $\lambda \approx$ 200–400 km, appear to travel with inclined wavefronts—having both horizontal and vertical motions which are often of similar magnitudes—and with varying direction. The effects are independent of magnetic activity, and are probably not caused by atmospheric winds. A travelling compressional wave is suggested as an alternative to motion caused by electric fields.

551.510.535 **1731**

Electron Distribution in a New Model of the Ionosphere.—H. K. Kallmann. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 82–87.) Rocket measurements between 90 and 200 km for pressure, density and temperature are used to derive a new empirical model for temperature and electron density in the ionosphere. This model agrees well with the theory of temperature variation and radio measurements of electron density. See also 436 of 1957 (Kallmann et al.).

551.510.535 **1732**

Some Implications of Slant E_s.—E. K. Smith & R. W. Knecht. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 195–204.) Slant E_s is common in polar regions, the virtual height of traces rising from 100 up to 500 km. A linear characteristic passing through the origin relates virtual height and frequency. It is suggested that slant E_s is due to back-scatter from E_s. Four modes of propagation are possible: (*a*) direct back-scatter, (*b*) reflected back-scatter, (*c*) single reflected scatter, (*d*) double scatter. Experiments are suggested which would distinguish between them.

551.510.535 **1733**

An Easily Applied Method for the Reduction of *h-f* Records to *N-h* Profiles including the Effects of the Earth's Magnetic Field.—E. R. Schmerling. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 8–16.) The method depends on sampling h' at fixed sub-multiples of the frequencies f_n at which true heights are required. The latter are then obtained by averaging the values of h'. A complete computation is presented for Washington, D.C., and the results checked against an analysis of $h'(f)$ curves for known profiles.

551.510.535 **1734**

The Electron Distribution in the Ionosphere over Slough: Part 1—Quiet Days.—J. O. Thomas, J. Haselgrove & A. Robbins. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 46–56.) The distribution of electron density with height is calculated from $h'(f)$ records using an electronic computer. The method assumes that the $N(h)$ curve increases monotonically and allows for the effect of the earth's magnetic field. The height of maximum electron density in the F_2 layer is found to be considerably lower than previously supposed, particularly in the summer and equinox months.

551.510.535 **1735**

The Height Variation of Drift in the E Region.—I. L. Jones. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 68–76.) Simultaneous observations at 2·0 and 2·5 Mc/s of drift velocities show significant differences and it is deduced that they refer to different heights. The velocities rotated through 360° at a nonuniform rate during daylight hours; the nonuniformity is due to the variation of reflection height. In the winter months the rotation at the smaller height lagged with respect to that at the larger. Rapid changes in drift were observed during September and October. The results agree with those deduced by Greenhow and Neufeld (1415 of 1957) from meteor trail observations, if allowance is made for the height differences.

551.510.535 **1736**

Solar Tidal Effects in the F_2 Region of Ionosphere over Delhi.—C. S. R. Rao. (*Indian J. Phys.*, Oct. 1957, Vol. 31, No. 10, pp. 516–525.) Harmonic analysis of h_pF_2 and f_0F_2 shows the existence of semi-diurnal and seasonal solar tidal effects in the F_2 layer at Delhi. Vertical drifts of 20 km/h in summer, 18 km/h in winter, and 33 km/h at the equinoxes are calculated from ionization density variations. Attachment coefficients, including these tidal effects, worked out for different seasons are in agreement with those obtained by Ratcliffe et al. (2724 of 1956).

551.510.535 : 523.164 **1737**

A Study of the Ionospheric Irregularities which Cause Spread-F Echoes and Scintillations of Radio Stars.—B. H. Briggs. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 34–45.) A discussion of the correlation between the occurrence of spread-F echoes at Slough, Inverness and Oslo. The irregularities responsible for the spreading occur in bands which lie along parallels of latitude, and have a width of the order of 450 km. It is suggested that the irregularities occur at heights near 300 km.

551.510.535 : 523.78 **1738**

Ionospheric Records of Solar Eclipses.—G. H. Munro & L. H. Heisler. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 57–67.) The horizontal gradients of ionization produced during an eclipse introduce curvature into the isoionic contours so that the ionosonde soundings are effectively oblique. The resultant errors in the deduced ion distributions, assuming vertical soundings, give rise to many of the apparent abnormalities in eclipse records.

551.510.535 : 551.55 **1739**

Travelling Disturbances in the Ionosphere: Changes in Diurnal Variation.—G. H. Munro. (*Nature, Lond.*, 7th Dec. 1957, Vol. 180, No. 4597, pp. 1252–1253.) The mean diurnal variation in the direction of horizontal movement of travelling ionospheric disturbances is plotted for January 1957 and for January in the years 1951–1954. The main feature of the former is an unusually large change of direction about midday possibly associated with high sunspot activity.

551.510.535+551.510.52] : 621.396.11 **1740**

Progress in the Field of Ionospheric Research and Tropospheric Wave Propagation.—B. Beckmann. (*Nachrichtentech. Z.*, Aug. 1957, Vol. 10, No. 8, pp. 369–376.) Report based on papers presented at a conference held at Kleinheubach, Germany, 11th–13th October 1956 and organized by the German U.R.S.I. Committee and the Nachrichtentechnische Gesellschaft.

551.510.535 : 621.396.11 **1741**

Ionospheric True Height and M.U.F. Calculations.—J. O. Thomas & A. Robbins. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 77–79.) M.u.f.'s calculated from the more accurate determinations of the electron density distribution (see 1734 above) are consistently higher than the Slough Bulletin values.

551.510.535 : 621.396.11 **1742**

The Spectrum of the Electron Density Fluctuations in the Ionosphere.—R. M. Gallet. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 165–170.) The fluctuations are assumed to arise from the vertical transport of small pockets of air in the presence of (*a*) a non-adiabatic temperature gradient and (*b*) vertical gradients of mean electron density. A preliminary account of the theory of the spectrum of the fluctuations is given with references to the application of the results to ionospheric and tropospheric scattering of radio waves. See also 234 of 1956.

551.510.535+550.385](98) **1743**

Polar Disturbances.—J. H. Meek. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 120–128.) Disturbances in the ionosphere and geomagnetic field, and also auroral behaviour, are usually treated statistically. In this paper an attempt is made to investigate what occurred in a few specific cases. There is some evidence for a spiral auroral zone.

551.510.535(98) **1744**

Statistical Results and their Shortcomings concerning the Ionosphere within the Auroral Zone.—R. W. Knecht. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 109–119.) The difficulties of obtaining significant data from high-latitude $h'(f)$ curves are discussed together with the statistical treatment of such data. These difficulties could be minimized by the adoption of the recommendations of the U.R.S.I. High Latitude Committee.

551.510.535(98) **1745**

Measurements of Irregularities and Drifts in the Arctic Ionosphere using Airborne Techniques.—G. J. Gassman. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 44–51.) An $h'(f)$ ionospheric recorder was installed in an aircraft and flights were made in North Polar regions. The interpretation of the experimental data in terms of ionospheric structure is discussed. Some drift measurements were made at the North Pole. See also 2382 of 1956.

551.510.535(98) : 621.396.11 **1746**

Quantitative Measurements of Absorption in the Auroral Zone.—F. Lied. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 135–146.) The various alternative methods of measuring ionospheric absorption are discussed and the results compared with particular reference to the North Polar region.

551.510.535(98) : 621.396.11 **1747**

Echoes from the Lower Ionosphere during Polar Blackouts.—B. Landmark. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 79–80.)

551.510.535(98) : 621.396.11 **1748**

Polar Blackout Occurrence Patterns.—V. Agy. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 129–134.) A brief survey of the geographical distribution of the probability of the occurrence of blackouts in North Polar regions. Difficulties in interpretation of the data are discussed. See also 1039 of 1955.

551.594.6 **1749**

Classification of Atmospheric Waveforms.—F. Hepburn. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 1–7.) The desirability and requirements of a systematic classification of atmospheric waveforms are discussed. An observational scheme is suggested and interpreted in terms of known properties of the discharge and propagation mechanisms. These data are reviewed to clarify application of the scheme to individual waveforms. The relation between the classification and previous inadequate groupings is indicated.

LOCATION AND AIDS TO NAVIGATION

621.396.663 : 550.372 **1750**

Phase Distortion due to Ground Inhomogeneities.—K. Baur. (*Nachrichtentech. Z.*, Aug. 1957, Vol. 10, No. 8, pp. 385–389.) Formulae are derived for assessing the quality of the ground with regard to fluctuations of conductivity and dielectric constants, thereby assisting in the selection of a site suitable for d.f.

621.396.933 **1751**

An Improved Medium-Range Navigation System for Aircraft.—C. G. McMullen. (*Trans. Inst. Radio Engrs*, Sept. 1956, Vol. ANE-3, No. 3, pp. 103–107. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1956, Vol. 44, No. 12, p. 1897.) A survey of existing systems leads to the conclusion that range-measurement equipment is more precise than bearing-measurement equipment. The use of a computer with existing equipment is suggested to give range-range measurements in terms of elliptical or hyperbolic coordinates.

621.396.933 : 621.396.676 **1752**

Azimuth Errors for the TACAN System.—D. W. T. Latimer, Jr. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. ANE-3, No. 4, pp. 150–156. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 570.)

621.396.96 : 621.396.677.832 **1753**

V-Reflex Aerial for an Information Radar Station.—von Trentini & Kirkscether. (See 1635.)

621.396.96.001.362 **1754**

Radar Simulators.—L. J. Kennard & C. H. Nicholson. (*J. Brit. Instn Radio Engrs*, Jan. 1958, Vol. 18, No. 1, pp. 17–30. Discussion, pp. 30–31.) The characteristics required for an accurate presentation of radar signals for air traffic control, surface vessel movement and air and naval warfare, including various types of jamming are discussed. An accurate aerial-pattern function generator is necessary and two complete computing systems are described.

621.396.96.001.362 **1755**

The Use of Radar Simulators in the Royal Navy.—P. Tenger. (*J. Brit. Instn Radio Engrs*, Jan. 1958, Vol. 18, No. 1, pp. 33–47.) A detailed description of an instrument developed to meet specialized requirements.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 : 546.482.41 : 539.23 **1756**

High-Voltage Photovoltaic Effect.—L. Pensak. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, p. 601.) Vacuum-evaporated films of CdTe have been prepared that show unusually high photovoltages across their ends. The effect is independent of the electrode material and the voltage is proportional to the length of the film.

535.215 : 546.482.41 : 539.23 **1757**

Properties of Photovoltaic Films of CdTe.—B. Goldstein. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 601–603.) The photoelectric properties of a CdTe film of the type described by Pensak (1756 above) are analysed. Results are consistent with the film containing a series of *p-n* junctions having a distribution of barrier heights.

535.37 **1758**

Influence of Activator Environment on the Spectral Emission of Phosphors.—G. R. Fonda. (*J. opt. Soc. Amer.*, Oct. 1957, Vol. 47, No. 10, pp. 877–880.) For isomorphous basic compounds the spectral emission is governed by the field strength to which the activator is subjected by its environment.

535.37 : [546.321.31 + 546.472.21 **1759**

Nature of Luminescent Centres in Alkali Halide and Zinc Sulphide Phosphors.—F. E. Williams. (*J. opt. Soc. Amer.*, Oct. 1957, Vol. 47, No. 10, pp. 869–876.) "The theory of the excitation and emission spectra of Tl-activated KCl is reviewed and extended to the problem of oscillator strengths for luminescent transitions. Semiconductor theory is applied to ZnS phosphors and the acceptor and donor nature of activator and coactivator, respectively, is thereby revealed."

535.376 **1760**

Electroluminescence.—S. T. Henderson. (*Brit. J. appl. Phys.*, Feb. 1958, Vol. 9, No. 2, pp. 45–51.) The present state of research in field-controlled and carrier-injection luminescence is briefly reviewed. Intrinsic luminescence is considered in relation to the behaviour of various phosphors, mainly ZnS with Cu as impurity activator, the voltage/brightness relation, temperature effects and the dependence of the emission waveforms upon the waveform of the applied field. Differences in the theoretical explanations of the behaviour are discussed and practical applications of the phenomenon are outlined. 69 references.

535.376 : 546.26–1 **1761**

A Surface Electroluminescence Effect in Diamonds.—H. J. Logie & R. R. Urlau. (*Nature, Lond.*, 7th Dec. 1957, Vol. 180, No. 4597, pp. 1254, 1271.) Diamonds coated with a thin layer of graphite have been found to luminesce with bright green spots of light when a potential difference is applied to them. See also 2468 of 1957 (Wolfe & Woods).

535.376 : 546.472.21 **1762**

Trapping Action in Electroluminescent Zinc Sulphide Phosphors.—C. H. Haake. (*J. opt. Soc. Amer.*, Oct. 1957, Vol. 47, No. 10, pp. 881–887.) Discussion of the results of brightness measurements showing that electron traps are more important in electroluminescent phosphors than in ordinary photoluminescent phosphors.

537.226/.227 **1763**

Optical and Dielectric Investigation of Boracite.—Y. Le Corre. (*J. Phys. Radium*, Nov. 1957, Vol. 18, No. 11, pp. 629–631.) An abrupt change in the dielectric constant of $Mg_3B_7O_{13}Cl$ at 265° and results of measurements support the hypothesis that the material is ferroelectric.

537.226/.227 **1764**
: [546.431.824-31 + 546.42.824-31

Phase Equilibria in the System $BaTiO_3$-$SrTiO_3$.—J. A. Basmajian & R. C. DeVries. (*J. Amer. ceram. Soc.*, Nov. 1957, Vol. 40, No. 11, pp. 373–376.) Report on investigations above 1 200° C.

537.227/.228.1 : 546.32.882.5 **1765**

Method for Growing Single Crystals of Potassium Niobate.—C. E. Miller. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 233–234.) Crystals approximately 1 in. on each side have been obtained.

537.311.33 **1766**

Problems of the Metallurgy and Physics of Semiconductors.—(*Ak. Nauk S.S.S.R. Special Publication*, 1957, 152 pp.) The texts are given of the following papers presented at the Second Conference on Semiconductors held in January 1956 at the A. A. Baĭkova Institute of Metallurgy of the U.S.S.R. Academy of Sciences, Moscow.

(*a*) Problems closely related to the Development of the Metallurgy of Semiconductors.—D. A. Petrov (pp. 5–11). See also 2168 of 1957.

(*b*) Preparation of High-Purity Silicon by an Iodine Method.—I. A. Inozemtseva (pp. 12–17).

(*c*) Preparation of Pure Silicon by the Method of Reduction of Silicon Chloride by Zinc.—D. A. Petrov & L. K. Zhukova (pp. 18–23).

(*d*) Preparation of Silicon Single Crystal by the Method of Drawing from the Melt.—B. P. Mitrenin, Sh. S. Burdiashvili, N. A. Shamba, V. P. Volkov, V. K. Kovyrzin & L. K. Solov'ev (pp. 24–34.)

(*e*) Use of Zone Melting for Obtaining Silicon Single Crystals.—B. P. Mitrenin, S. P. Lalykin, Yu. P. Savrasov & L. K. Radaĭkin (pp. 35–40).

(*f*) Preparation of Silicon Single Crystals.—D. A. Petrov, M. G. Kekua, V. D. Khvostikova, Yu. M. Shashkov & A. D. Suchkova (pp. 41–46).

(*g*) The Problem of Growing Single Crystals of Germanium from a Melt.—A. P. Izergin (pp. 47–49).

(*h*) Developments in the Purification of Germanium by Crystallization Methods, and in the Preparation of Single Crystals of Germanium with Uniform Longitudinal Properties.—D. A. Petrov, M. G. Kekua, M. Ya. Dashevskiĭ, V. S. Zemskov & P. L. Petrusevich (pp. 50–58).

(*i*) Investigation of the Possibility of Obtaining a Homogeneous Alloy of Germanium with Silicon by means of Zone Melting.—B. P. Mitrenin, N. E. Troshin, K. P. Tsomaya, V. A. Vlasenko & Yu D. Gubanov (pp. 59–69).

(*j*) Preparation of Single Crystals of AlSb and Investigation of their Properties.—D. A. Petrov, M. S. Mirgalovskaya, I. A. Strel'nikova & E. M. Komova (pp. 70–79).

(*k*) Synthesis of Aluminium Antimonide (AlSb) and Some of its Properties.—G. N. Nikolaenko (pp. 80–90).

(*l*) Investigation of the System Bi-Te.—N. Kh. Abrikosov, V. F. Bankina & G. A. Fedorova (pp. 91–96).

(*m*) Investigation of Thermoelectric Properties of Cobalt Antimonide.—L. D. Dudkin & N. Kh. Abrikosov (pp. 97–109).

(*n*) Vitreous Semiconductors.—N. A. Goryunova & B. T. Kolomiets (pp. 110–120).

(*o*) Diffusion Coefficient of some Impurities in Germanium.—B. I. Boltaks (pp. 121–129). See also 484 of 1957.

(*p*) Diffusion of Antimony and Germanium in Silicon.—D. A. Petrov, Yu. M. Shashkov & I. P. Akimchenko (pp. 130–132).

(*q*) Influence of Impurities on the Lifetime of Excess Charge Carriers in Germanium.—A. V. Rzhanov (pp. 133–137). See 2190 of 1957.

(*r*) The Problem of Using the Photoelectric Method of Measuring the Diffusion Length of Minority Current Carriers in Silicon.—I. D. Kirvalidze (pp. 138–141).

(*s*) New Etchants for Silicon and Germanium.—B. I. El'kin (pp. 142–151).

537.311.33 **1767**

Magnetic Field Effects on Electron Populations in Semiconductors.—P. T. Landsberg. (*Proc. phys. Soc.*, 1st Jan. 1958, Vol. 71, No. 457, pp. 69–76.) The change in population of the states of the conduction band when a magnetic field is applied is calculated. It is assumed that the conduction band states are shifted up or down by an energy *d* due to the interaction of their spins with the magnetic field and that the donor states are also shifted up or down in energy by the same amount d. The more important general properties of the model are discussed and a numerical example is worked out.

537.311.33 **1768**

The Influence of Interelectronic Collisions on Conduction and Breakdown in Covalent Semiconductors.—R. Stratton. (*Proc. roy. Soc. A*, 5th Nov. 1957, Vol. 242, No. 1230, pp. 355–373.) The breakdown field strength and variation of mobility with field strength are calculated for conditions of high, moderate, and low electron densities; results for all three cases are similar. The energy and momentum distributions of electrons (or holes) are used; at high electron densities, both distributions are largely determined by interelectronic collisions, but at lower densities only the energy distribution is affected. Acoustic and optical lattice mode scattering are considered for various temperature ranges.

537.311.33 **1769**

On the Theory of Surface Recombination in Semiconductors for Large Potential Differences between Surface and Bulk.—F. Berz. (*Proc. phys. Soc.*, 1st Feb. 1958, Vol. 71, No. 458, pp. 275–280.) When the potential variation over a carrier mean free path is of the order of kT/e, the theoretical results show hole-electron capture cross-section ratios different from those to be expected on a simple Maxwell-Boltzmann distribution picture.

537.311.33 : 535.215 **1770**

Comparison of the Semiconductor Surface and Junction Photovoltages.—R. O. Johnson. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 556–577.) It is shown analytically that the surface and junction photovoltages are almost identical phenomena. The functional dependence of the two photovoltages on the light-injected carrier densities is exactly the same, except in the region of saturation. Charge changes in traps can have a marked effect on the surface photovoltage, however, but no direct effect on the total junction photovoltage.

537.311.33 : 537.311.4 **1771**

Simplified Treatment of Electric Charge Relations at a Semiconductor Surface.—E. O. Johnson. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 525–555.) Simple, graphical representation is used to describe the electric charge and potential relations at a semiconductor surface. The balance between trapped and mobile charge at the surface is considered for both equilibrium and non-equilibrium conditions. The treatment is extended to cover metal/semiconductor, gas/semiconductor and *p-n* junction interfaces.

537.311.33 : 537.312.8 **1772**

Isotropic Approximation to the Magnetoresistance of a Multivalley Semiconductor.—R. W. Keyes. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 43–46.) A multivalley model is suggested which describes approximately the magnetoresistance phenomena of a polycrystalline semiconductor. The results provide a means for determining the valley anisotropy of a multivalley semiconductor from magnetoresistance data without reference to a particular model of the band structure. See also 1480 of 1957.

537.311.33 : 537.312.9 **1773**

An Apparatus for Measuring the Piezoresistivity of Semiconductors.—R. F. Potter & W. J. McKean. (*J. Res. nat. Bur. Stand.*, Dec. 1957, Vol. 59, No. 6, RP 2814, pp. 427–430.) A detailed description is given of an apparatus and procedure designed to measure the piezoresistive effect in semiconductors over an extended temperature range. A tensile force up to 1 kg can be applied to the sample by means of a calibrated beam balance. The apparatus has been used for measurements on InSb over the range 78°K–300°K, and tensile stresses of the order of 5×10^7 dynes/cm² can be applied to samples that are cut in a special manner.

537.311.33 : 538.63 **1774**

The Theory of the Nernst Effect in Semiconductors.—J. E. Parrott. (*Proc. Phys. Soc.*, 1st Jan. 1958, Vol. 71, No. 457, pp. 82–87.) The effects of phonon-electron drag on the Nernst effect are examined using general methods previously described (3511 of 1957). A quantitative agreement between theory and experiments on germanium is reported. It is shown that the sign of the Nernst coefficient can be altered by a phonon current.

537.311.33 : [546.28 + 546.289 **1775**

Growth of Silicon and Germanium Disks.—J. R. O'Connor & W. A. McLaughlin. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, p. 222.) A preliminary investigation has been made of the growth of large disks of semiconductor materials by slowly withdrawing a disk-shaped seed from a melt.

537.311.33 : [546.28 + 546.289 **1776**

Electron Mobility in the Germanium-Silicon Alloys.—B. Goldstein. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 458–465.) The method of measurement involves the determination of charge-carrier density from the voltage dependence of the depletion-layer capacitance of a *p-n* junction formed on the material under test. Similar qualitative behaviour has been found for the conductivity and Hall mobilities as changes are made in alloy composition.

537.311.33 : 546.28 **1777**

Arrangements of Dislocations in Plastically Bent Silicon Crystals.—J. R. Patel. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 170–176.) "Dislocations introduced into single crystals of Si by plastic bending at an elevated temperature have been studied quantitatively by the etch-pit technique. The average etch-pit density after deformation is approximately two to three times higher than the calculated dislocation density."

537.311.33 : 546.28 **1778**

Distorted Layers of Silicon Produced by Grinding and Polishing.—W. C. Dash. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 228–229.)

537.311.33 : 546.28 **1779**

Experimental Study of Spin-Lattice Relaxation Times in Arsenic-Doped Silicon.—J. W. Culvahouse & F. M. Pipkin. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 319–327.) Relaxation times for the stable As^{75} were measured by using fast passage techniques to observe the relative amplitudes of the electron resonance signals as a function of time. The relaxation times for the radioactive As^{76} were measured by observing the formation and decay of the nuclear alignment. The measurements were made at 8 500 gauss and 1·3° K.

537.311.33 : 546.28 **1780**

Preparation of Pure Silicon by the Hydrogen Reduction of Silicon Tetraiodide.—G. Szekely. (*J. electrochem. Soc.*, Nov. 1957, Vol. 104, No. 11, pp. 663–667.) Investigation of this method shows that the reaction is heterogeneous, taking place on a hot surface. Dense Si layers as well as crystals can be deposited. The reaction of Si and I_2 to form SiI_4 and the fractional distillation of this compound are discussed.

537.311.33 : 546.28 : 621.314.7 **1781**

Transistor-Grade Silicon: Part 1—The Preparation of Ultra-pure Silicon Tetraiodide.—B. Rubin, G. H. Moates & J. R. Weiner. (*J. electrochem. Soc.*, Nov. 1957, Vol. 104, No. 11, pp. 656–660.) "A stepwise method of preparing and purifying SiI_4 has been found involving the direct combination of the elements, recrystallization of the product, followed by sublimation and zone purification steps. The values of the segregation coefficients of several impurity elements have been determined."

537.311.33 : 546.289 **1782**

Relaxation-Time Anisotropy in *n*-Type Germanium.—C. Goldberg. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 331–335.) The anisotropy parameter $K = K_m/K_\tau$ is determined from magnetoconductance measurements in the temperature range 45°K to 300°K. Using the cyclotron resonance value for K_m, the measurements give $K_\tau = 1{\cdot}0$ for lattice scattering and $K_\tau > 1$ for moderate amounts of impurity scattering.

537.311.33 : 546.289 **1783**

Surface Properties of Semiconductors.—W. H. Brattain. (*Science*, 26th July, 1957, Vol. 126, No. 3265, pp. 151–153.) A *p-n* junction in a crystal of Ge and a Ge surface in an ambient gas are considered. The surface properties of the Ge in the latter case depend primarily on the surface treatment and on the nature of the gas, not on the type, *p* or *n*, of the body material.

537.311.33 : 546.289 **1784**

New Phenomenon in Narrow Germanium *p-n* Junctions.—L. Esaki. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 603–604.) Dynatron-type current/voltage characteristics have been observed which are tentatively explained on the basis of field emission across the junction.

537.311.33 : 546.289 **1785**

Oxygen as a Donor Element in Germanium.—G. Elliott. (*Nature, Lond.*, 14th Dec. 1957, Vol. 180, No. 4598, pp. 1350–1351.) Single crystals of Ge grown in both pure N_2 or pure A showed sudden changes of resistivity when small quantities of air were added to the atmosphere.

537.311.33 : 546.289 : 537.534.9 **1786**

Etching of Germanium Crystals by Ion Bombardment.—G. K. Wehner. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 217–221.) "A study is made of the etch effects produced by sputtering Ge crystals and bicrystals under normal incident low-energy (100 eV) Hg^+ ion bombardment in a low-pressure plasma (1 μ gas pressure)."

537.311.33 : 546.289 : 538.63 **1787**

Galvanomagnetic Effects in Oriented Single Crystals of *n*-Type Germanium.—W. M. Bullis. (*Phys. Rev*, 15th Jan. 1958, Vol. 109, No. 2, pp. 292–301.) Measurements of the magnetoresistance, Hall, and planar Hall coefficients on oriented single crystals of *n*-type Ge at 77°K and 300°K are described. Effective values for the anisotropy factor and the mean free time τ are deduced from the 77°K data assuming τ anisotropic but energy-independent. The 300°K data are interpreted on the basis of the low-field approximation for the conductivity tensor and various simplifying approximations for the energy dependence of τ.

537.311.33 : 546.289 : 538.632 **1788**

Measurement of the Hall Mobility in *n*-Type Germanium at 9 121 Mc/s.—Y. Nishina & W. J. Spry. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 230–231.) A single crystal of 16-Ω. cm *n*-type Ge at room temperature gave a microwave value of 2 900 cm²/V.sec compared with a measured d.c. value of 2 670 cm²/V.sec.

537.311.33 : 546.289 : 539.16 **1789**

Low-Temperature Irradiation of *N*-Type Germanium.—J. W. Cleland & J. H. Crawford, Jr. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 149–151.) Studies of irradiation effects in Ge at temperatures well below that of liquid nitrogen have been conducted to examine the thermal stability of radiation-induced defects and the importance of minority-carrier trapping processes.

537.311.33 : 546.289 : 539.4 **1790**

Some Effects of Environment on Fracture Stress of Germanium.—P. Breidt, Jr, J. N. Hobstetter & W. C. Ellis. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, p. 226.)

537.311.33 : 546.289 : 621.396.822 **1791**

Generation Recombination Noise in Intrinsic and Near-Intrinsic Germanium Crystals.—J. E. Hill & K. M. van Vliet. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 177–182.) Measurements are made on single crystals at temperatures between 300° K and 450° K. Results are compared with theory considering generation and recombination by means of direct transitions or via recombination centres located either in the bulk or at the surface. Reasonable agreement is found in most cases. In some the spectrum falls off rather weakly at higher frequencies indicating that the recombination centres do not lie at a single sharp energy in the forbidden gap but are distributed within a small range.

537.311.33 : 546.47.19 **1792**

Zn_3As_2, Semiconducting Intermetallic Compound.—G. A. Silvey. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 226–227.) Preparation and properties are discussed.

537.311.33 : 546.682.18 **1793**

Decomposition Method for Producing *p-n* Junctions in InP.—K. Weiser. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 229–230.) A simple method is described to produce *p-n* junctions in compound semiconductors which decompose before melting and contain a volatile constituent such as InP or GaAs.

537.311.33 : 546.817.221 **1794**

Intrinsic Optical Absorption and the Radiative Recombination Lifetime in PbS.—W. W. Scanlon. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 47–50.) A description of measurements of the absorption coefficient in the region of intrinsic absorption in PbS. The coefficients range from about 10 cm^{-1} to 10^5 cm^{-1}, and the radiative recombination lifetime for PbS calculated from the data is 63 μs at 300° K.

538.22 : 621.318.1 **1795**

Magnetism in Materials.—D. H. Martin. (*Wireless World*, Jan.–April 1958, Vol. 64, Nos. 1–4, pp. 28–30, 70–74, 126–131 & 178–180.)

Part 1—The Physical Basis of Dia-, Para-, Ferro- and Ferri-magnetism.

Part 2—Ferromagnetic Domains and their Influence on Magnetic Properties.—Hysteresis, coercivity and magnetostriction are explained physically in terms of domain wall movement, rotation of the axis of magnetization within a domain and impurity content. Fine powder magnets in which each grain of powder is a single domain are discussed.

Part 3—Commercial Magnetic Materials and Domain Theory.—Hysteresis and associated parameters are considered, and characteristic values for magnetically soft materials including ferrites are given, with details of their composition and preparation.

Part 4—Rectangular - Hysteresis - Loop Materials. Permanent Magnets.—A note on the applications of these materials and a review of the development of permanent-magnet materials leading to Co and Ba ferrites and powdered alloys.

538.221 **1796**

Ferromagnetism of a Zirconium-Zinc Compound.—B. T. Matthias & R. M. Bozorth. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 604–605.)

538.221 **1797**

Studies on the Magnetic Anisotropy Induced by Cold Rolling of Ferromagnetic Crystal: Part 1—Iron-Nickel Alloys.—S. Chikazumi, K. Suzuki & H. Iwata. (*J. phys. Soc. Japan*, Nov. 1957, Vol. 12, No. 11, pp. 1259–1276.) Magnetic domain patterns and torque measurements were made with single crystals and polycrystals of Ni_3Fe, the direction of easy magnetization being found for different crystallographic orientations. The results are theoretically explained by assuming that slip deformation induces the directional order observed after rolling.

538.221 : 538.569.4 **1798**

Microwave Resonance in Nickel at 35 Gc/s.—G. S. Barlow & K. J. Standley. (*Proc. Phys. Soc.*, 1st Jan. 1958, Vol. 71, No. 457, pp. 45–48.) From measurements on a Ni single crystal, the anisotropy constants K_1 and K_2 have been determined from 20 to 150° C. At 20° C, $K_1 = K_2 = -6{\cdot}06 \times 10^4$ ergs cm^{-3} gives the best fit to the experimental data. Measurements on Ni-Cu and Ni-Mn alloys are also reported.

538.221 : 539.215.1 **1799**

Loss of Exchange Coupling in the Surface Layers of Ferromagnetic Particles.—F. E. Luborsky. (*Phys. Rev.*, 1st Jan. 1958, Vol. 109, No. 1, pp. 40–42.) Experiments with spherical iron particles 20 Å to 265 Å in diameter show that the suggested nonferromagnetic surface layer on an iron particle must be less than 1 Å thick.

538.221 : 539.23 **1800**

Thin Ferromagnetic Films.—S. J. Glass & M. J. Klein. (*Phys. Rev.*, 15th Jan. 1958, Vol. 109, No. 2, pp. 288–291.) "The spontaneous magnetization of thin films of ferromagnetic materials has been studied by means of spin-wave theory. Results have been obtained for the magnetization as a function of temperature and film thickness for body-centered and face-centered cubic materials, generalizing earlier calculations by Klein & Smith [1675 of 1951]. The approximations in the theory are critically discussed, and the relevant experimental material is briefly reviewed."

538.221 : 539.23 : 538.569.4 **1801**

Ferromagnetic Resonance at U.H.F. in Thin Films.—R. H. Kingston & P. E. Tannenwald. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 232–233.) Experimental results are compared with theoretical values.

538.221 : 548.0 **1802**

The Crystal Structures of a New Group of Ferromagnetic Compounds.—P. B. Braun. (*Philips Res. Rep.*, Dec. 1957, Vol. 12, No. 6, pp. 491–548.) A description of four new compounds which are structurally related to magnetoplumbite. O and Ba atoms form a slightly expanded closely packed arrangement, with the Ba atoms in certain selected positions, and the smaller ions in certain of the holes between the large ones. 'Plates', either four or six O layers thick, can be distinguished. The relation between the structures is discussed. An appendix describes a calculator for performing Fourier syntheses with indices up to 60.

538.221 : 621.318.2 **1803**

Extremely Small Permanent Magnets.—(*Tech. News Bull. nat. Bur. Stand.*, Nov. 1957, Vol. 41, No. 11, pp. 179–180.) Details are given of the processing and magnetic properties of cold-drawn cunife wire 0·005 in. in diameter.

538.221 : 621.318.134 : 537.226 **1804**

The Dielectric Behaviour of Magnesium Manganese Ferrite.—J. Peters & K. J. Standley. (*Proc. phys. Soc.*, 1st Jan. 1958, Vol. 71, No. 457, pp. 131–133.) Measurements on a ferrite having the approximate composition 0·9 MgO, 0·1 MnO, 0·8 Fe_2O_3 in the frequency range 30 c/s–100 Mc/s and in the temperature range 20–220°C, are reported and discussed.

538.221 : 621.318.134 : 548.0 **1805**

Crystal Distortion in Ferrite-Manganites.—G. I. Finch, A. P. B. Sinha & K. P. Sinha. (*Proc. roy. Soc. A*, 8th Oct. 1957, Vol. 242, No. 1228, pp. 28–35.) The origin of the distortion of spinels from cubic to tetragonal symmetry is examined using copper ferrite and a series of manganite-ferrites. The degree of distortion depends on temperature and on the fraction of cations forming appropriately orientated dsp^2 bonds in octahedral sites. There is agreement with experiment.

538.221 : 621.318.134 : 548.73 **1806**

An Improved X-Ray Method for Determining Cation Distribution in Ferrites.—L. P. Skolnick, S. Kondo & L. R. Lavine. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 198–203.) The method is of general value in distinguishing between two elements distributed over non-equivalent positions in a crystal lattice when their X-ray scattering factors are almost equal.

538.221 : 621.318.134
: [621.317.335.3 + 621.317.41 **1807**

Measurement of Ferrite Loss Factors at 10 Gc/s.—Srivastava & Roberts. (See 1820.)

538.569.4 : 666.94 **1808**

Debye-Type R.F. Absorption in Cements.—J. Le Bot. (*J. Phys. Radium*, Nov. 1957, Vol. 18, No. 11, pp. 638–639.) The values of ϵ^1 and ϵ^{11} in Portland and non-normalized aluminous cement for temperatures from 4° to 350°K and frequencies 0·1, 1·0, 10 and 100 kc/s were determined in an investigation of the process of water fixation.

621.315.612 + 621.318.124
+ 621.318.134 **1809**

Recent Developments iu Ceramic Materials for the Electronic Industries.—P. Popper. (*Brit. Commun. Electronics*, Nov. 1957, Vol. 4, No. 11, pp. 694–701.) A review of the properties and applications of ceramic materials with a list of insulators, resistors, ferrites and dielectrics giving their main characteristics and the names of British manufacturers.

621.319.2 : 535.215 **1810**

Theory of Photoelectrets.—V. M. Fridkin, N. T. Kashukeev & I. S. Zheludev. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Dec. 1957, Vol. 117, No. 5, pp. 804–807.) An expression for the electron concentration in trapping levels in single-crystal sulphur is derived. Results show a direct ratio between the charge of an illuminated specimen and the intensity of the polarizing field. No saturation was observed even with fields of 20 kV/cm. See also 3627 of 1955 (Chatterjee & Bhadra).

MATHEMATICS

517.6 : 517.52 **1811**

On the Calculation of the Function j_0 (z, θ) for Large Values of 'z'.—H. E. Fettis. (*J. Math. Phys.*, Oct. 1957, Vol. 36, No. 3, pp. 279–283.) Asymptotic and integral expansions of the function j_0 (z, θ) $= \int_0^\theta e^{iz \cos \phi}\, d\phi$ are considered when z is large, e.g. of the order of $1/\lambda$ in optical diffraction or radar scattering problems.

517.946.4 : 530.145.6 **1812**

The Reduced Wave Equation in a Medium with a Variable Index of Refraction.—W. L. Miranker. (*Commun. pure appl. Math.*, Nov. 1957, Vol. 10, No. 4, pp. 491–502.)

517.946.4 : 530.145.6 **1813**

On Solutions of Nonlinear Wave Equations.—J. B. Keller. (*Commun. pure appl. Math.*, Nov. 1957, Vol. 10, No. 4, pp. 523–530.)

517.949 : 681.142 **1814**

On the Solution of the Schroedinger and the Klein-Gordon Equations by Digital Computers.—H. F. Harmuth. (*J. Math. Phys.*, Oct. 1957, Vol. 36, No. 3, pp. 269–278.)

MEASUREMENTS AND TEST GEAR

621.3.018.41 (083.74) : 538.569.4 **1815**

Experimental Evaluation of the Oxygen Microwave Absorption as a Possible Atomic Frequency Standard.—J. M. Richardson. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 137–145.) "Theoretical design, actual design, and results for an oxygen microwave spectrometer for use either in observing the line frequency or as a discriminator in a frequency control loop synchronizing an oscillator are described. Essential characteristics are the rate of change of spectrometer output signal with frequency and the output noise level. General expressions for these quantities for a wide range of experimental arrangements are obtained, and may be used to predict the attainable frequency precision."

621.3.018.41(083.74)
: 621.373.421.13 **1816**

A Frequency Standard Stable to 2 parts in 10^{10}.—(*Brit. Commun. Electronics*, Nov. 1957, Vol. 4, No. 11, p. 681.) A brief description of a quartz crystal oscillator using a servo system of frequency control.

621.317.1 : 621.395.625.3 **1817**

Magnetic Tape Recorders in Measurement Techniques.—H. Wehde. (*Elektrotech. Z., Edn A*, 1st Nov. 1957, Vol. 78, No. 21, pp. 792–796.) Advantages and some applications are discussed.

621.317.2 : 621.373.4.029.3 **1818**

Calibrated Audio Oscillator.—G. C. Fox. (*Short Wave Mag.*, Dec. 1957, Vol. 15, No. 10, pp. 514–519.) Practical details of the design and construction of an instrument useful for Doppler shift measurements and a.f. testing. Frequency is controlled by a nonlinear potentiometer.

621.317.2 : 621.373.421.13 **1819**

Wide-Range Crystal Marker Generator.—H. Pollack. (*Radio TV News*, Dec. 1957, Vol. 58, No. 6, pp. 68–69, 179.) Design detail of a unit providing 100-kc/s and 1-Mc/s check-points up to 150 Mc/s.

621.317.335.3 + 621.317.41]
: 538.221 : 621.318.134 **1820**

Measurement of Ferrite Loss Factors at 10 Gc/s.—C. M. Srivastava & J. Roberts. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 204–209.) "The analysis is presented of the losses arising in a rectangular cavity containing ferrite slabs which extend the full length of the cavity. The loss factors associated with the dielectric constant and the scalar and tensor permeabilities are deduced from Q-factor measurements on the cavity. The methods are particularly suited to low-loss ferrites."

621.317.337 : 621.385.029.6 **1821**

'Cold' Methods of Measuring Magnetron Quality.—W. Schmidt. (*Elektronische Rundschau*, Aug. 1957, Vol. 11, No. 8, pp. 235–241.) Three groups of methods are distinguished and compared and examples of each are given.

621.317.35.082.5 : 621.372.412 **1822**

Piezo-optic Frequency Analyser.—T. Ogawa. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 46–47.) Note of a method of frequency analysis in which the birefringence induced in one of a set of ADP crystals at resonance is detected using a beam of polarized light passing through the crystal.

621.317.39.088.7 **1823**

Extending Transducer Transient Response by Electronic Compensation for High-Speed Physical Measurements.—F. F. Liu & T. W. Berwin. (*Rev. sci. Instrum.*, Jan. 1958, Vol. 29, No. 1, pp. 14–22.) Electronic compensators are described which automatically and continuously correct for dynamic errors of electromechanical transducers during transient and steady-state measurements, even in regions beyond the transducer's natural frequency. Transient phenomena with rise time a fraction of a microsecond can be measured directly with a minimum of amplitude and phase distortion.

621.317.42 : 550.380.87 **1824**

An Electrical Recording Magnetometer.—P. H. Serson. (*Canad. J. Phys.*, Dec. 1957, Vol. 35, No. 12, pp. 1387–1394.) The equipment described, which is of the saturated-transformer type, is used for recording at a fixed station variations in three orthogonal components of the earth's magnetic field. See also 229 of 1957 (Serson & Hannaford).

621.317.44 : 621.375.2 **1825**

Integrator-Amplifier for Core Measurements.—C. E. Goodell. (*Electronics*, 14th Feb. 1958, Vol. 31, No. 7, pp. 110–113.) Instantaneous flux is measured as the time integral of voltage. Details of design and complete circuits are given.

621.317.619 : 621.316.825 **1826**

Measurement of some Characteristic Parameters of Thermistors with Low Time Constant based on their Harmonic and Transient Characteristics.—V. Andresciani & A. Lepschy. (*Note Recensioni Notiz.*, May/June 1957, Vol. 6, Supplement to No. 3, pp. 25–48.) Three methods are described, two dealing with the determination of the impedance locus curve and one with obtaining the time constant of a thermistor from an analysis of its transient response characteristic.

621.317.7 : 621.314.7 **1827**

A 1-kc/s Junction-Transistor *T*-Parameter Measurement Set.—R. A. Hall. (*Electronic Engng*, Feb. 1958, Vol. 30, No. 360, pp. 82–85.) A description of the set and experimental procedure is given. Frequency-dependent errors which occur in the case of low frequency transistors can be avoided by reducing the measurement frequency to 200 c/s.

621.317.7 : 621.397.62 : 621.373.444.1 **1828**

Television Timebase Measurements: Anode Dissipation of Line Output Valves.—A. Ciuciura. (*Mullard tech. Commun.*, Dec. 1957, Vol. 3, No. 27, pp. 228–231.) A method of measuring dissipation is described involving tests under both dynamic and d.c. conditions; bulb temperature, measured by means of a thermocouple, is used as a reference for both tests.

621.317.725 : 621.314.7 **1829**

A New Transistorized Voltmeter.—H. Malamud. (*Radio TV News*, Nov. 1957, Vol. 58, No. 5, pp. 66–67, 174.) Description of an instrument with two transistors in a bridge circuit, with sensitivity 100 000 Ω/V and voltage ranges 1, 10, 100 and 1 000 V.

621.317.73.029.64 **1830**

An Instrument for the Measurement of Surface Impedance at Microwave Frequencies.—A. E. Karbowiak. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 195–203.) The surface impedance is deduced from the resonant conditions of the cavity operated simultaneously in the H_{01} and E_{11} modes. Experimental details are given of instruments for use at 6 kMc/s and 34 kMc/s, together with experimental results.

621.317.74 **1831**

Measuring Equipment with Automatic Frequency Sweep and Visual or Graphic Presentation of Measurements.—W. Bürck. (*Elektrotech. Z.*, *Edn A*, 1st Nov. 1957, Vol. 78, No. 21, pp. 782–785.) Description of commercial-type frequency-sweep oscillator equipment, including a wide-band wobbulator covering 500 kc/s–400 Mc/s with 100-Mc/s maximum sweep, and a recording-type a.f. analyser.

621.317.755 : 621.318.57 **1832**

A Simple Three-Channel C.R.O. Beam Switch.—W. F. Lovering & M. P. Hearn. (*Electronic Engng*, March 1958, Vol. 30, No. 361, pp. 134–135.) The arrangement is suitable for both high- and low-frequency switching operations. The design incorporates a three-phase multivibrator and requires only six valves.

621.317.757 **1833**

Instrument for Viewing and Measuring Group Delay Frequency Characteristics, Phase Shift and Modulus of Propagation Coefficient. (Frequency Curve Tracer).—I. T. Turbovich, A. V. Knipper & V. G. Solomonov. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 31–42.) The operation and instrumental errors of the equipment are described.

621.317.772 **1834**

Measurement of Small Phase Shifts with a Phase-Sensitive Voltmeter.—D. J. Collins & J. E. Smith. (*Electronic Engng*, March 1958, Vol. 30, No. 361, pp. 146–147.) Phase shifts of the order of 1° are measured by 'backing-off' the in-phase component to reduce the reference meter indication. This allows increased voltmeter sensitivity with a consequent increase of the indication of the quadrature component.

621.317.794 **1835**

Bolometers for Radiation Measurements.—G. Barth. (*Arch. tech. Messen*, Sept. 1957, No. 260, pp. 201–204.) The principal parameters are defined and the construction of various types of bolometer as well as some typical bridge circuits for measuring infrared radiation are described. 28 references.

621.317.794 : 535.61-15 **1836**

A Possible Infrared Detector using Thermal Expansion.—R. V. Jones. (*Proc. phys. Soc.*, 1st Feb. 1958, Vol. 71, No. 458, pp. 280–283.) The detector has a r.m.s. noise level only twice that of the external noise fluctuations. It makes use of the linear expansion of a constantan strip which causes the deflection of an optical lever.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

551.508.71 : 535.325 **1837**

A New-Type Refractive-Index Variometer.—K. Hirao & K. Akita. (*J. Radio Res. Labs, Japan*, Oct. 1957, Vol. 4, No. 18, pp. 423–437.) It is shown that the variation of refractive index can be expressed as a linear combination of wet and dry bulb temperature variations for the purposes of radio meteorology. Small bead-type thermistors having a response time <1 sec and an accuracy within 0·1°C are used in the sounding head which is raised by captive balloon.

551.508.71 : 535.34-15 **1838**

Improved Infrared-Absorption-Spectra Hygrometer.—R. C. Wood. (*Rev. sci. Instrum.*, Jan. 1958, Vol. 29, No. 1, pp. 36–42.) A hygrometer is described which utilizes an infrared light beam as the principal sensing element. The beam contains narrow bands of radiation around 2·45 and 2·60 μ, the latter only being attenuated by water vapour. The instrument measures the ratio of the transmitted energies and this gives the concentration of water vapour.

551.508.71 : 538.569.4 **1839**

Recording Microwave Hygrometer.—J. B. Magee & C. M. Crain. (*Rev. sci. Instrum.*, Jan. 1958, Vol. 29, No. 1, pp. 51–54.) "This paper describes a rapid-response microwave hygrometer for continuously recording the water vapour pressure of atmospheric air over a wide ambient range. The principle employed involves the measurement by means of a cavity resonator of the contribution of water vapour to the refractive index of atmospheric air."

621-52 : 621.373 **1840**

A Function Generator.—N. Hambley. (*Electronic Engng*, Feb. 1958, Vol. 30, No. 360, pp. 91–94.) A graph of the function is converted to a proportional voltage by means of a closed-loop feedback system involving a c.r. tube and photocell.

621.365.5 **1841**

Radio-Frequency Hardening and Brazing.—F. Viart. (*A.C.E.C. Rev., Charleroi*, 1957, Nos. 3/4, pp. 2–21.) The influence of frequency, and of the permeability and resistivity of the material under treatment, on process efficiency is investigated with particular reference to skin effect. Practical operating data are derived and some industrial applications are described.

621.384.6 **1842**

Particle Accelerators.—R. Wideröe. (*VDI Z.*, 11th Dec. 1957, Vol. 99, No. 35, pp. 1743–1754.) The operating principles of the various types of accelerators are described, and a survey of existing and projected installations is made. 59 references.

621.387.462 : 546.28 **1843**

Silicon Crystal Counters.—W. D. Davis. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 231–232.) Single crystals of Si doped with impurities producing energy levels deep within the forbidden region make excellent crystal counters. Results with gold-doped Si are described.

621.398 : 550.389.2 : 629.19 **1844**

Telemetering in Earth Satellites.—W. Matthews. (*Elect. Engng, N.Y.*, Nov. 1957, Vol. 76, No. 11, pp. 976–981.) Square-hysteresis-loop magnetic cores with switching transistors form the basis of a 48-channel telemetering system weighing 3·2 oz. The power consumption is 4 mA at 2·7 V.

PROPAGATION OF WAVES

621.396.11 **1845**

Wave Propagation over an Irregular Terrain: Part 2.—K. Furutsu. (*J. Radio Res. Labs, Japan*, Oct. 1957, Vol. 4, No. 18, pp. 349–393.) A multiple series has been obtained for propagation over an earth having discontinuities in both height and electrical properties. When the spherical sections of irregular terrain are large the series converges rapidly and one term is sufficient, but when the sections are small and substantially plane supplementary formulae are required. These formulae are applied to diffraction over a ridge to derive obstacle gain. Part 1: 3985 of 1957.

621.396.11 : 551.510.52 **1846**

Phase Relations in the Diffraction Shadow in the Stratified Troposphere.—E. Berg. (*Arch. elekt. Übertragung*, Sept. 1957, Vol. 11, No. 9, pp. 366–378.) The solution of the wave equation for a magnetic dipole in a troposphere with overcritical refraction is given in a form from which the solutions for a homogeneous or nearly homogeneous atmosphere can be derived as special cases. For infinite ground conductivity an approximation is obtained showing the dependence of the cut-off wavelength on the refractive index and the height of a tropospheric duct. See also 3385 of 1955.

621.396.11 : [551.510.535+551.510.52 **1847**

Progress in the Field of Ionospheric Research and Tropospheric Wave Propagation.—Beckmann. (See 1740.)

621.396.11 : 551.510.535 **1848**

The Fading of Radio Waves Reflected at Oblique Incidence.—J. W. King. (*J. atmos. terr. Phys.*, 1958, Vol. 12, No. 1, pp. 26–33.) An analysis of the diffraction pattern formed on the ground by an m.f. wave indicates that the ionospheric irregularities are larger in the horizontal than in the vertical plane, that the speed of fading of a wave of frequency f is proportional to $f \cos i$, where i is the angle of incidence, and that the amplitude probability distribution tends to be 'log-normal' in form.

621.396.11 : 551.510.535 **1849**

Consideration of M.U.F. on the Basis of Results of Measurements of Field Intensity on Commercial Circuits.—S. Ishikawa & I. Kasuya. (*J. Radio Res. Labs, Japan*, Oct. 1957, Vol. 4, No. 18, pp. 439–443.) Comparison of m.u.f.'s, calculated by the two-point control method, and fade-in times, shows a possible variation of transmission path from the great circle in winter during the sunspot-cycle minimum.

621.396.11 : 551.510.535 **1850**

Forecasting of Disturbed H.F. Communication Conditions.—R. C. Moore. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 147–156.) Forecasts are produced for the North Atlantic and North Pacific areas and are issued in advance by approximately 3 weeks, 1 week, 1 day and 6 hours. The forecasts are based on the 27-day recurrence tendency, ionospheric behaviour, radio transmissions and other relations which are described.

621.396.11 : 551.510.535(98) **1851**

Prediction Techniques at High Latitudes.—J. H. Meek. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 101–107.) Critical frequencies and m.u.f. factors are expressed as linear functions of 12-month running-mean sunspot number. The intercept and slope of these relations are plotted as contour maps on gnomonic projections centred on the North pole. A permanent set of maps avoids the need for making new prediction maps every month.

621.396.11 : 621.396.674.3 **1852**

Excitation of Surface Waves on Conducting, Stratified, Dielectric-Clad, and Corrugated Surfaces.—J. R. Wait. (*J. Res. nat. Bur. Stand.*, Dec. 1957, Vol. 59, No. 6, RP 2807, pp. 365–377.) "An expression for the field of an electric dipole located over a flat surface with a specified surface impedance Z is derived from the formal integral solution by a modified saddlepoint method." 27 references.

621.396.11.029.62 : 551.510.535 **1853**

Results of Scatter Measurements at 36 Mc/s over a 1 200-km Path.—T. Hagfors. (*J. atmos. terr. Phys.*, 1957, Special Supplement, Part II, pp. 205–209.) The results refer to a north-south path in Norway and include data on the diurnal changes in signal strength, the correlation of field strength with ionospheric and magnetic data, and the angular dependence of the scattering cross-section.

621.396.11.029.63 **1854**

Broad- and Narrow-Beam Investigations of S.H.F. Diffraction by Mountain Ridges.—K. Nishikori, Y. Kurihara, M. Fukushima & M. Ikeda. (*J. Radio Res. Labs, Japan*, Oct. 1957, Vol. 4, No. 18, pp. 407–422.) The results are given of measurements of the horizontal angle of arrival, signal strength and spatial variation of 2 980-Mc/s radio waves transmitted over ridges and mountains. The received signals consisted of multiple components arriving from directions up to seven degrees apart. Spatial variation was caused by interference between these multiple components. The increase of loss with diffraction angle was found to be greater than that predicted by the Fresnel-Kirchhoff theory.

621.396.11.029.64 **1855**

Microwave Propagation over the Sea beyond the Line of Sight.—M. Onoue, K. Nishikori, M. Nenohi, A. Takahira, H. Irie & R. Usui. (*J. Radio Res. Labs, Japan*, Oct. 1957, Vol. 4, No. 18, pp. 395–406.) The vertical angle of arrival of 3-cm waves over an 80-km path and the variation of field intensity with distance up to 120 km have been compared with the vertical distribution of refractive index. Four types of diurnal and seasonal variation were distinguished depending on the season. A qualitative correlation between meteorological factors and variation in field strength was found with ducting playing an important part. Leakage from the duct and variation in the profile with time have also been studied.

621.396.11.029.65 **1856**

Propagation of Millimetre Waves through the Atmosphere.—A. B. Crawford & D. C. Hogg. (*Bell Lab. Rec.*, Dec. 1957, Vol. 35, No. 12, pp. 494–497.) A general discussion of atmospheric absorption at mm wavelengths. A two-way transmission method is described for measuring absorption in the range 0·5–0·6 cm which is more accurate than the one-way method.

621.396.81.029.63 **1857**

Microwave Field Strength and Fading in the Presence of Intervening Ridges.—R. Vikramsingh, M. N. Rao & S. Uda. (*J. Instn Telecommun. Engrs, India*, Dec. 1957, Vol. 4, No. 1, pp. 18–24.) Propagation at 2 kMc/s over mountain ridges has been studied for two paths of length 14 and 54 km. Fresnel's diffraction theory gives values of path loss lower than those measured. The fading in the shadow region is reduced which may be of advantage in some microwave link applications. See also 1040 of April.

RECEPTION

621.396.621 : 621.314.7 **1858**

A Portable Transistor Receiver.—L. E. Jansson, J. B. Ruming & J. M. Tapley. (*Mullard tech. Commun.*, Dec. 1957, Vol. 3, No. 27, pp. 198–208.) A design, based on six transistors, giving maximum output power of 200 mW and a sensitivity of 200 μV across the aerial tuned circuit for 50 mW output.

621.396.621 : 621.376.33 **1859**

Frequency-Modulation Negative Feedback in F.M. Receivers.—L. Ya. Kantor. (*Radiotekhnika, Mosk.*, Jan. 1957, Vol. 12, No. 1, pp. 58–62.) The i.f. pass band required to ensure stability and to limit nonlinear distortion is determined.

621.396.621 : 621.376.33 : 621.373.421 **1860**

Choice of the Interstage Coupling Circuit in Frequency-Modulated Receivers.—L. Ya. Kantor. (*Elektrosvyaz'*, April 1957, No. 4, pp. 29–32.) The use of band-pass filters is suggested for low-sensitivity receivers; for high-sensitivity receivers single-tuned circuits are preferred. Expressions are derived for the increase in the receiver selectivity due to f.m. feedback. See also 1427 of 1952 (Hacks).

621.396.621.029.62 : 621.376.332 **1861**

Pulse-Counter F.M. Receiver.—M. G. Scroggie. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 181–183.) Supplementary notes on one year's operation of the prototype (2524 of 1956) are given, including particulars of conversion to crystal control.

621.396.621.029.64 **1862**

Designing Low-Noise Microwave Receivers.—C. T. McCoy. (*Electronic Ind. Tele-Tech*, Nov. & Dec. 1957, Vol. 16, Nos. 11 & 12, pp. 54–57 . . 154 & 64–65 . . 146.) The effects of local oscillator noise, mixer variations, and varying bandwidth are summarized. Experimental data on microwave crystal mixers are included.

621.396.621.54 : 621.396.96 : 621.396.822 : 621.317.6 **1863**

Pencil and Paper Calculation of Noise Level in Superheterodyne Radar Receivers.—D. W. Haney. (*Trans. Inst. Radio Engrs*, Dec. 1956, Vol. ANE-3, No. 4, pp. 157–160. Abstract, *Proc. Inst. Radio Engrs*, April 1957, Vol. 45, No. 4, p. 570.) See also 3828 of 1956.

621.396.662 **1864**
Miniature Ferrite Tuner Covers Broadcast Band.—E. A. Abbot & M. Lafer. (*Electronics*, 28th Feb. 1958, Vol. 31, No. 9, pp. 72–73.) A 'rotary-axial' r.f. tuning element using two pairs of ferrite cups is described; rotation of D-shaped centre sections combined with an axial movement separating the cores gives linear frequency variation from 500 to 1 600 kc/s for 270° shaft rotation.

621.396.812.5.029.51 : 523.75 **1865**
A New Effect of Chromospheric Eruptions.—M. Waldmeier. (*Naturwissenschaften*, Aug. 1957, Vol. 44, No. 16, p. 439.) Field strength measurements made near Berne of a 56·35-kc/s transmission from London during the solar eruption of 16th April 1957 show an unexpected decrease of signal strength by up to 14 dB, whereas on 101·65 kc/s an increase of up to 6 dB was measured.

621.396.821 **1866**
Atmospheric Noise Interference to Short-Wave Broadcasting.—S. V. C. Aiya. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 580–589.) An idealized cloud discharge containing four strokes, each with a stepped leader, is considered. The parameters required by a receiving system for assessing the noise are derived, taking into account the characteristics of the human ear. Each discharge may be considered as a single acoustic impulse, and the estimation of the noise level in terms of a modulated c.w. signal, as used in the noise meter described in 257 of 1955, is discussed. See also 3263 of 1955.

621.396.822 **1867**
Freedom from Interference in Different Systems of Radiotelegraphy.—Yu. S. Lezin. (*Elektrosvyaz'*, April 1957, No. 4, pp. 40–47.) A comparison of a.m., f.m. and ph.m. systems indicating the advantages of ph.m.

621.396.823 **1868**
Radio Interference by Corona Discharges on High-Voltage Lines.—W. Wechsung. (*Elektrotech. Z.*, *Edn B*, 21st Oct. 1957, Vol. 9, No. 10, pp. 385–388.) A method is proposed for assessing the amount of interference to radio reception caused by neighbouring high-voltage lines. Measurements are based on a frequency of 250 kc/s and a distance from the line of 20 m, and with a maximum permissible value of 1 mV/m for the interference field strength interference-free medium-wave reception should be possible at a distance of at least 100 m from the line where the useful signal strength is at least 1 mV/m.

STATIONS AND COMMUNICATION SYSTEMS

621.391 **1869**
On the Compressibility of the Spectrum of a Signal.—A. A. Kharkevich. (*Electrosvyaz'*, April 1957, No. 4, pp. 3–11.) General aspects of the problem of bandwidth compression are considered and some examples are given.

621.391 **1870**
Some Graphical Approaches to Coding Problems.—J. Dutka. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 466–474.) "The problem of constructing error-detecting and error-correcting codes for use in communicating information in binary coded form has received considerable attention in recent years. Some graphical methods for constructing such codes are presented, their geometrical interpretations are discussed, and some illustrative examples are worked out."

621.391 : 621.376 : 534.78 **1871**
On the Power Gained by Clipping Speech in the Audio Band.—W. Wathen-Dunn & D. W. Lipke. (*J. acoust. Soc. Amer.*, Jan. 1958, Vol. 30, No. 1, pp. 36–40.) Available data on speech amplitude distributions are briefly examined and the results of Davenport (3314 of 1952) are used to calculate the power increase obtainable by peak clipping and subsequent amplification.

621.391 : 621.376.56 **1872**
A Coder for Halving the Bandwidth of Signals.—A. R. Billings. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 182–184.) A possible practical system is described in which a continuous message of finite bandwidth is coded into a continuous signal of smaller bandwidth.

621.396.2 : 621.394.3 **1873**
Communication Technique for Multipath Channels.—R. Price & P. E. Green, Jr.—(*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 555–570.) A new system known as 'Rake' is described. The mark/space sequence of symbols is transmitted as a wide-band signal and those portions arriving at the receiver with different delays are isolated, using correlation detection techniques. The separated signals are weighted and appropriate delays applied to bring them back into time coincidence. Communication theory applicable to the system is reviewed and examples are given of experimental tests of the system. 53 references.

621.396.41 **1874**
Design Principles of Equipment for Simplified Systems of Multichannel Links in Cable and Radio Relay Systems.—K. P. Egorov & M. U. Polyk. (*Elektrosvyaz'*, April 1957, No. 4, pp. 48–54.) A coaxial-cable system is described with 32 groups of 30 channels operating in the frequency band 312–8 500 kc/s; repeater spacing is 6–6·5 km.

621.396.712.3 **1875**
Two New Large Vehicles for Sound-Broadcast Transmissions.—L. V. Türkheim. (*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 145–150.) Description of mobile control rooms used by the Bayerische Rundfunk for outside-broadcast transmissions via Post Office lines.

621.396.74.029.62 (43) **1876**
List of V.H.F. Transmitters in the German Federal Republic.—(*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 165–167.) See also 1894 below.

SUBSIDIARY APPARATUS

621.311.6 : 621.314.7 **1877**
Heat Transfer in Power Transistors.—I. G. Maloff. (*Electronic Ind. Tele-Tech*, Dec. 1957, Vol. 16, No. 12, pp. 54–55 .. 157.) A general discussion of thermal problems in operating power transistors between 25°C and 85°C.

621.311.6 : 621.396.931 **1878**
Power Supply and Suppression in Portable/Mobile Working.—D. T. Bradford. (*Short Wave Mag.*, Nov. 1957, Vol. 15, No. 9, pp. 465–469.) Practical details of the use of car batteries or petrol-electric sets for power supply.

621.311.6.027.3 : 621.385.032.22 **1879**
Stabilized E.H.T. Unit.—D. J. Collins & J. E. Smith. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 184–186.) Design of a compact equipment for anode supplies of 1 350–1 500 V.

621.316.722.078.3 **1880**
Improved Control Circuit for Regulated Power Supplies.—G. W. Jones. (*QST*, Nov. 1957, Vol. 41, No. 11, pp. 30–33.) A cathode follower is inserted between the control valve and the regulator valve in an electronic voltage regulator to increase the current range over which regulation may be maintained.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 **1881**
Economic Considerations in Closed-Circuit Television System Design.—D. Kirk, Jr. (*J. Soc. Mot. Pict. Telev. Engrs*, Nov. 1957, Vol. 66, No. 11, pp. 661–671.) During the last five years a number of systems for distribution of entertainment-type television programs via wire to homes of paying subscribers have been successfully installed and operated. The economics and design of such systems are considered, and possible improvements are outlined.

621.397.5 : 535.623 **1882**
Simulating Sharpness in Colour Television.—M. W. Baldwin, Jr. (*Bell Lab. Rec.*, Dec. 1957, Vol. 35, No. 12, pp. 481–484.) A 'defocusing' projector is used to simulate the blurring that results when a colour television picture is transmitted over a circuit of limited bandwidth.

621.397.6.001.4 **1883**
Pulse - Cross Modification of TV Receivers.—H. E. O'Kelley. (*Electronics*, 28th Feb. 1958, Vol. 31, No. 9, pp. 54–55.) "Phantastron circuits delay horizontal and vertical sync pulses when added to monitor or TV receiver to provide pulse-cross display. System gives simple means of checking operation of station sync generator."

621.397.61 : 535.623 : 778.5 **1884**

Advanced Performance and Stability in Colour TV Film Channel Amplifiers.—M. H. Diehl. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1957, Vol. 66, No. 12, pp. 750–754. Discussion, pp. 754–755.) "The use of three-channel a.g.c., precision gamma circuits, and high-level black clipper yields long-time stability of the critical parameters affecting colour balance. With large amounts of negative feedback in the monitoring section, drastic reduction in the number of controls, and built-in calibration features, set-up and adjustment can be rapidly accomplished."

621.397.61 : 778.5 : 621.396.665 **1885**

Automatic Gain Control in Television Automation.—M. H. Diehl, W. J. Hoffman & W. L. Shepard. (*J. Soc. Mot. Pict. Telev. Engrs*, Dec. 1957, Vol. 66, No. 12, pp. 755–757.) Describes a system providing a constant level output from a monochrome vidicon camera channel for light-level changes of 30 : 1. The application to colour film systems is also discussed.

621.397.61.001.4 **1886**

Video Testing Techniques in Television Broadcasting.—A. Ste-Marie. (*Elect. Engng, N.Y.*, Nov. 1957, Vol. 76, No. 11, pp. 968–973.) "A discussion is presented of the practical effects on signal degradation of the parameters in a video transmission system. An attempt is made to establish their true significance and correlation. New video testing techniques have been established for improved and simplified methods of specification and measurement."

621.397.611 : 535.8 **1887**

Resolution Chart aids TV Camera Focusing.—G. Southworth. (*Electronics*, 14th Feb. 1958, Vol. 31, No. 7, pp. 100–101.) The chart consists of a number of parallel lines of different thickness which are scanned and displayed on a waveform monitor.

621.397.611.2 **1888**

Test and Measurement Methods for Image Orthicon Camera Tubes.—F. Pilz. (*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 125–138.) Methods are outlined of measuring characteristics peculiar to the image orthicon, such as the secondary-emission factor, gain of signal multiplier, capacitance of target-mesh assembly, storage time, and modulation depth. Details of a specially designed test set for image orthicons are given.

621.397.62 : 621.314.7 **1889**

Transistor Television Circuits: Part 1—Synchronizing Separators and Timebase Oscillators.—J. N. Barry & G. W. Secker. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 154–158.) Practical details are given of a common-emitter sync separator, a line oscillator, and a frame oscillator, for a 17-inch television receiver.

621.397.62 : 621.373.444.1 : 621.317.7 **1890**

Television Timebase Measurements: Anode Dissipation of Line Output Valves.—Ciuciura. (See 1828.)

621.397.62 : 621.385.832.001.4 **1891**

Colour and Monochrome Cathode-Ray-Tube Performance Tests.—C. F. Otis. (*Elect. Engng, N.Y.*, Nov. 1957, Vol. 76, No. 11, pp. 990–995.) Acceptance tests for c.r. tubes should be simple and, if possible, quantitative. A description is given of a 6-position test rack for monochrome tubes, and the characteristics of 3-gun colour tubes most troublesome in set design are discussed.

621.397.62 : 621.396.665 **1892**

A.G.C. Circuits for Positive-Modulation Television Receivers.—P. L. Mothersole. (*Mullard tech. Commun.*, Dec. 1957, Vol. 3, No. 27, pp. 214–227.) A number of representative a.g.c. systems are examined and their main failings are discussed. The development of a simple gate circuit free from these faults and having other advantages is described.

621.397.621 **1893**

Investigation of Aperture Distortion by the Method of Split Image Reproduction.—E. L. Orlovskiĭ. (*Electrosvyaz'*, April 1957, No. 4, pp. 55–66.)

621.397.7 (43) **1894**

The Television Network of the German Federal Republic.—(*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 159–162.) Tabulated data on television transmitters as at 1st June 1957, with maps showing their location and that of television links.

621.397.7 (43) **1895**

List of Television Transmitters in the German Federal Republic.—(*Rundfunktech. Mitt.*, Aug. 1957, Vol. 1, No. 4, pp. 163–164.) See also 1894 above.

621.397.8 **1896**

Subjective Sharpness of Television Pictures.—W. N. Sproson. (*Electronic Radio Engr*, April 1958, Vol. 35, No. 4, pp. 124–132.) "The subjective sharpness of television pictures has been measured using a comparison technique and a multi-criterion scale for assessment. Two types of degrading network were used and the subjective sensitivity to changes in equivalent rectangular bandwidth has been evaluated for both static and moving pictures."

621.397.8 **1897**

Influence of Periodic-Type Interference on the Quality of a Television Image.—A. P. Efimov. (*Elektrosvyaz'*, April 1957, No. 4, pp. 22–28.) The effect of interference on the reproduction of static and moving images is examined for interference frequencies in the ranges 0·1–6·4 Mc/s and 0·1–26 Mc/s, respectively. The relation of interference level to image quality is determined experimentally.

621.397.8 **1898**

Band-V Signal Strength.—A. Hale. (*Wireless World*, April 1958, Vol. 64, No. 4, pp. 162–163.) Investigation of reception conditions along the A5 and A10 roads from London as far as Towcester and Cambridge.

621.397.822 **1899**

Measurement and Evaluation of Shot Noise in the Video Band.—E. Sennhenn. (*Elektronische Rundschau*, Sept. 1957, Vol. 11, No. 9, pp. 271–274.) Curves are derived relating the theoretical and observable peak noise voltages to the effective value as a function of frequency. Subjective tests were made with a noise signal 1 Mc/s wide, variable in the range 0–7 Mc/s, superimposed on a 625-line image. The peak noise voltage required to create a given impression of 'graininess' is plotted against noise frequency for two different viewing distances. The sensitivity of the observer decreases with increasing frequency. Photographs of images with interference of constant level but differing frequency are reproduced.

621.397.9 **1900**

New Developments in the Field of Industrial Television.—E. F. Spiegel. (*Elektronische Rundschau*, Sept. 1957, Vol. 11, No. 9, pp. 261–264.) Applications outlined include the testing of bore holes using a miniature camera with a vidicon-type tube of reduced size, temperature measurement by means of infrared-sensitive vidicons, and the telemetering of width, e.g. of sheet material in rolling mills, using two television cameras.

TRANSMISSION

621.396.61 : 621.376.2 **1901**

The Effect of Electron Inertia on the Shape of the Modulation Characteristic of A.M. Transmitters.—L. N. Kolesov. (*Radiotekhnika, Mosk.*, Dec. 1956, Vol. 11, No. 12, pp. 28–36.) Recommendations are made for ensuring the linearity of the modulation characteristic, and the requisite design formulae are derived.

VALVES AND THERMIONICS

621.314.63+621.314.7] : 539.169 **1902**

The Effects of Short-Duration Neutron Radiation on Semiconductor Devices.—W. V. Behrens & J. M. Shaull. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 601–605.) Transistors suffered a decrease in forward current gain and an increase in backward collector current, the effect being much greater with a.f.- than with h.f.-transistors, the former being virtually useless after irradiation at 10^{13} neutrons/cm^2. Diodes exhibited an increase in forward resistance and a decrease in back resistance. The results suggest that the integrated neutron dosage is of more importance to semiconductor devices than the rate of exposure.

621.314.63+621.314.7] : 621.396.822 **1903**

Theory of Junction-Diode and Junction-Transistor Noise.—A. van der Ziel & A. G. T. Becking. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 589–594.)

A rigorous yet general proof of the equations governing shot noise, in which the only restriction to the model is that individual holes may be considered independent. See also 600 of 1956.

621.314.7 **1904**

Transistor Technology.—(*Nature, Lond.*, 14th Dec. 1957, Vol. 180, No. 4598, pp. 1329–1330.) Summaries of papers read at a conference on transistors organized by the Institute of Physics and held at Acton, Middlesex, 27th-28th September 1957.

621.314.7 **1905**

Transistors, Reliability and Surfaces.—C. G. B. Garrett. (*Bell Lab. Rec.*, Nov. 1957, Vol. 35, No. 11, pp. 466–470.) A descriptive account of the physical effects of the surface oxide film on semiconducting materials.

621.314.7 **1906**

Influence of Hydration-Dehydration of the Germanium Oxide Layer on the Characteristics of *P-N-P* Transistors.—J. T. Wallmark & R. R. Johnson. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 512–524.) When Ge *p-n-p* transistors are subjected to a change in temperature, the zero-frequency common-emitter current gain shows a corresponding change, approaching an asymptotic value in approximately 48 h. The effect is interpreted in terms of a hydrated oxide layer on the Ge surface.

621.314.7 **1907**

Variation of Junction-Transistor Current Amplification Factor with Emitter Current.—A. W. Matz. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 616–617.) A short mathematical note on the volume recombination and emitter efficiency terms which helps to resolve the difference between analyses of Webster (2798 of 1954) and Rittner (3390 of 1954).

621.314.7 **1908**

Experimental Determination of the Base and Emitter Lead Resistances of Alloy-Junction Transistors by means of Low-Frequency Measurements.—W. Guggenbühl & W. Wunderlin. (*Arch. elekt. Übertragung*, Sept. 1957, Vol. 11, No. 9, pp. 355–358.) The method described is based on the low-frequency *h*-parameters; results obtained compare satisfactorily with those given by more elaborate methods thereby proving the validity of one-dimensional equations in determining the low-frequency characteristics of alloy-junction transistors.

621.314.7 : 621.317.7 **1909**

A 1-kc/s Junction-Transistor *T*-Parameter Measurement Set.—Hall. (See 1827.)

621.314.7 : 621.376.54 **1910**

A Conductivity-Storage Transistor Pulse-Width Modulator.—Price. (See 1695.)

621.314.7 (083.57) **1911**

Designing Stability into Transistor Circuits.—S. Schenkerman. (*Electronics*, 14th Feb. 1958, Vol. 31, No. 7, pp. 122, 124.) Chart and nomographs simplify calculation of circuit and cooling-facility parameters necessary for stable operation of Ge and Si transistors at elevated junction temperatures.

621.314.7.001.1 (091) **1912**

Research Leading to Point-Contact Transistor.—J. Bardeen. (*Science*, 19th July 1957, Vol. 126, No. 3264, pp. 105–112.)

621.383.2 **1913**

On the Phototubes Sensitive to the Wide Spectral Region.—M. Sugawara. (*J. phys. Soc. Japan*, Nov. 1957, Vol. 12, No. 11, pp. 1282–1290.) The light transmission and photoelectric yields of thin photo cathode films were investigated. A combination of Sb-Cs and Ag-CsO surfaces gave a photocell suitable for spectral photometry from ultraviolet to infrared.

621.383.4 **1914**

Maximum Performance of High-Resistivity Photoconductors.—R. W. Redington. (*J. appl. Phys.*, Feb. 1958, Vol. 29, No. 2, pp. 189–193.) It is shown that the transit time in a material showing space-charge-limited current cannot be less than the charge relaxation time. In consequence the photoconductor cannot simultaneously act as a detector, an amplifier and a storage element and still have a response time as short as the storage time. This puts a restriction on the performance of high-resistivity photoconductive devices.

621.383.4 : 535.371.07 **1915**

Solid-State Light Amplifiers.—B. Kazan & F. H. Nicoll. (*J. opt. Soc. Amer.*, Oct. 1957, Vol. 47, No. 10, pp. 887–894.) The characteristics of photoconductive and electroluminescent materials are summarized with particular emphasis on the problems of their optimum combination in the double-layer type of intensifier with or without optical feedback. The use of the overall device for radar display and X-ray intensification is discussed. See also 1897 of 1956 (Diemer et al.).

621.383.42 **1916**

Internal Resistance and Capacitance of a Selenium Photocell at Low Temperatures.—G. Blet. (*J. Phys. Radium*, Oct. 1957, Vol. 18, No. 10, pp. 572–578.) Observed variations of capacitance appear to be related to the resistance variations reported earlier (960 of March), and are such that the product *CR* tends in general to a constant value.

621.385.029.6 **1917**

Understanding the Travelling-Wave Amplifier.—D. A. Dunn. (*Electronic Ind. Tele-Tech*, Nov. 1957, Vol. 16, No. 11, pp. 67–69, 142.) A fundamental discussion of the physical processes occurring in a helical slow-wave structure.

621.385.029.6 **1918**

Low-Noise Tunable Preamplifiers for Microwave Receivers.—M. R. Currie & D. C. Forster. (*Proc. Inst. Radio Engrs*, March 1958, Vol. 46, No. 3, pp. 570–579.) A description of a new gun design for S-band backward-wave amplifier valves. The operation of the valve as a receiver component is discussed and detailed experimental performance data are given. Noise figures less than 6 dB and 4·5 dB for 25% and 10% of the tuning range respectively have been measured. Still lower figures appear possible for backward-wave and other microwave valves using the new gun design.

621.385.029.6 : 537.533 **1919**

On the Adiabatic Self-Constriction of an Accelerated Electron Beam Neutralized by Positive Ions.—J. D. Lawson. (*J. Electronics Control*, Dec. 1957, Vol. 3, No. 6, pp. 587–594.) The build-up of the self magnetic field causes the transverse oscillation of the electrons to be damped, but this damping is partly counteracted by an outward diffusion due to multiple scattering of the electrons on the ions. Inductive effects appear as an apparent increase in the mass of the electrons.

621.385.029.6 : 537.533 : 538.691 **1920**

Structure in Magnetically Confined Electron Beams.—H. F. Webster. (*J. appl. Phys.*, Dec. 1957, Vol. 28, No. 12, pp. 1388–1397.) "A number of observations have been made of structure changes that occur in hollow and solid electron beams which are confined by a magnetic field. These structure changes occur in both the density of the beam and the transverse velocity components of the beam electrons." See also 4060 of 1957 (Kyhl & Webster).

621.385.029.6 : 621.317.337 **1921**

'Cold' Methods of Measuring Magnetron Quality.—Schmidt. (See 1821.)

621.385.029.65 **1922**

Travelling-Wave-Tube Experiments at Six Millimetres Wavelength: Part 1—Phase-Velocity Measurements.—K. Kamiryo, H. Hozumi, Y. Shibata & Y. Fukushima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, June 1956, Vol. 8, No. 1, pp. 35–47.) A report of phase-velocity measurements carried out on first and second spatial harmonic waves both forward and backward in a travelling-wave valve of the type described by Millman (547 of 1952 and 1187 of 1953).

621.385.032.2 : 537.533 **1923**

Aperture Lens Formula Corrected for Space Charge in the Electron Stream.—C. K. Birdsall. (*Trans. Inst. Radio Engrs*, April 1957, Vol. ED-4, No. 2, pp. 132–134. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1957, Vol. 45, No. 8, p. 1163.)

621.385.032.21 : 537.533 **1924**

Valve Instability with Cathode Standing Waves of Cylindrical Symmetry.—W. W. H. Clarke. (*Brit. J. appl. Phys.*, Dec. 1957, Vol. 8, No. 12, pp. 486–490.) Experimental results obtained with a circular cathode of very small dimensions show a significant repeatability not attainable in previously reported experiments with a larger cathode. This confirms the existence of preferred standing-wave patterns, satisfying the boundary conditions, and associated preferred emission current characteristics in the cathodes of thermionic valves. See also 1913 of 1956.

621.385.032.21 **1925**

The Physics of the Cathode.—L. S. Nergaard. (*RCA Rev.*, Dec. 1957, Vol. 18, No. 4, pp. 486–511.) Recent work on thermionic emitters suggests some generalizations bearing on all electron emitters. Four propositions are advanced for consideration and discussion. The propositions are: (*a*) every cathode is a reducing agent; (*b*) every cathode lives in equilibrium with its environment; (*c*) every cathode is a dispenser cathode; (*d*) monolayer film emitters do not exist. Evidence to support these propositions is adduced. The evidence for the first three propositions is regarded as conclusive. 56 references.

621.385.032.213.13 **1926**

On the Mechanism of Operation of the Barium Aluminate Impregnated Cathode.—E. S. Rittner, W. C. Rutledge & R. H. Ahlert. (*J. appl. Phys.*, Dec. 1957, Vol. 28, No. 12, pp. 1468–1473.) Emission and evaporation characteristics of a porous tungsten cathode impregnated with the composition $5BaO.2Al_2O_3$ are presented and are interpreted in terms of the cathode mechanism. Emission is less than that of an L cathode, presumably because of release of a poisoning agent accompanying the activator. See also 2978 and 2979 of 1957.

621.385.032.213.13 **1927**

Cavity-Type Barium-Tungsten Cathode.—T. Hashimoto. (*Rep. elect. Commun. Lab., Japan*, Oct. 1957, Vol. 5, No. 10, pp. 1–8.) A report on development of a procedure for manufacturing satisfactorily dispenser-type L cathodes.

621.385.032.213.13 : 537.311.33 **1928**

Analysis of the D.C. and Pulsed Thermionic Emission from BaO.—G. A. Haas. (*J. appl. Phys.*, Dec. 1957, Vol. 28, No. 12, pp. 1486–1492.) "The effects of field penetration and donor mobility on the chemical potential of BaO have been computed by using a non-degenerate single donor level semiconductor model. Calculations which neglect the effects of surface states and porosity predict that the pulsed emission starts lower, but increases with field more rapidly than given by simple Schottky theory, actually being capable of exceeding the theoretical Schottky emission. The d.c. emission level is always lower than the pulsed emission, the difference being more pronounced at higher fields and for less active cathodes."

621.385.032.269.1 **1929**

The Theory of the Pierce-Type Electron Gun.—D. E. Radley. (*J. Electronics Control*, Feb. 1958, Vol. 4, No. 2, pp. 125–148.) "The problem of determining the Pierce electrodes in a gun reduces to a Cauchy problem on Laplace's equation. The questions of existence, uniqueness and instability of solutions to general problems of this type are considered, together with their relevance to the design of electron guns. A general procedure for solving a Cauchy problem is developed, and simplified for two-dimensional cases. This method is then used to determine the electrodes which will maintain strip, wedge, cylindrical and conical-shaped beams. The Cauchy conditions for these problems are given by the space-charge-limited potential from the appropriate complete diode solution."

621.385.1 : 621.376.32.029.3 : 621.3.011.3 **1930**

The Reactance Valve at Audio Frequencies.—Alcock. (See 1646.)

621.385.2 **1931**

Transit Time and Space Charge for the Cylindrical Diode.—L. Gold. (*J. Electronics Control*, Dec. 1957, Vol. 3, No. 6, pp. 567–572.) "Introduction of time-dependent Poisson equation admits a form of solution which leads to interesting relations for transit time and the current-voltage dependence for both zero and finite initial electron velocities. The analysis is expedited by employment of a reduced spatial variable and the development of inverse power series. The general correspondence with Langmuir's classic solution is demonstrated and, in particular, in the limit of vanishing cathode radius, a simple, non-series description corresponds to the case $\beta = 1$."

621.385.2 : 537.525.92 **1932**

Parametric Solution for the Diode Space Charge at Relativistic Energies.—L. Gold. (*J. Electronics Control*, Dec. 1957, Vol. 3, No. 6, pp. 564–566.) "Transformation of the basic equations that describe the behaviour of the diode space charge (including Poisson's relation) into a time-dependent form allows straightforward solution in the relativistic domain. The parametric solution lends itself to securing various approximate explicit results and readily yields the extreme relativistic limit of a current linearly dependent upon anode voltage." See also 4076 of 1957 (Acton).

621.385.2 : 621.396.822 **1933**

Space Charge as a Source of Flicker Effect.—C. S. Bull. (*Proc. Instn elect. Engrs*, Part B, March 1958, Vol. 105, No. 20, pp. 190–194.) Three types of fluctuation are predicted by the analysis: (*a*) shot noise, (*b*) an enhanced shot noise, and (*c*) a flicker effect dependent on the magnitude of the electronic capacitance. The results are discussed in relation to previous work (see e.g. 3080 of 1954 and 303 of 1955).

621.385.3 **1934**

Thermionic Emission from the Grid in Metal/Ceramic Valves.—E. P. Korchagine & G. M. Utkin. (*Elektrosvyaz'*, April 1957, No. 4, pp. 12–21.) Emission characteristics of Russian modulator valves Types GI-6B, GI-7B and GS-9B are discussed.

621.385.5(083.57) **1935**

Peak Anode Current Nomograms for Line Output Valves.—A. Ciuciura. (*Mullard tech. Commun.*, Oct. 1957, Vol. 3, No. 26, pp. 162–168.) The nomograms take into account production spread in valve characteristics, fall in emission with time, and feedback across the screen-grid resistor.

621.385.83 **1936**

Electronic Devices Employing Cathode-Ray Beams.—L. S. Allard. (*Brit. Commun. Electronics*, Oct. 1957, Vol. 4, No. 10, pp. 620–625.) The operating principles and applications of c.r. devices such as flying-spot scanners, image converters, storage, switching and multiplier tubes, and function generators are outlined.

621.385.832 **1937**

Level Displacement Diagrams for the Description of Charging Processes in Storage Tubes.—M. Knoll & W. Harth. (*Elektrotech. Z., Edn A*, 1st Aug. 1957, Vol. 78, No. 15, pp. 543–548.) Potential diagrams for a typical signal-storage tube (radechon) and a typical viewing storage tube are considered. 25 references.

621.385.832.001.4 : 621.397.62 **1938**

Colour and Monochrome Cathode-Ray-Tube Performance Tests.—Otis. (See 1891.)

MISCELLANEOUS

001.891 : 621.39 : 061.4 **1939**

The Post Office Research Station.—(*Nature, Lond.*, 7th Dec. 1957, Vol. 180, No. 4597, pp. 1240–1242.) A review of some of the research work of the station as reflected in exhibits at an open day at Dollis Hill, London, 27th September 1957. These included 4-kMc/s equipment for radio-communication, an echo waveform corrector for long-distance television links, an automatic error counter and a fading simulator for tests on frequency-shift telegraph systems, an a.f. analyser, and items relating to the technology of Ge components.

025.4 **1940**

A Statistical Approach to Mechanized Encoding and Searching of Literary Information.—H. P. Luhn. (*IBM J. Res. Developm.*, Oct. 1957, Vol. 1, No. 4, pp. 309–317.)

621.39 (047.1) **1941**

Communication Engineering and Radiolocation.—(*VDI Z.*, 11th Feb. 1958, Vol. 100, No. 5, pp. 193–205.) Progress report covering recent developments with references mainly to German literature. The following sections are surveyed:—

(*a*) Telecommunications.—W. Althans (pp. 193–195).

(*b*) Sound Broadcasting and Television.—E. Schwartz (pp. 195–199). 76 references.

(*c*) Electroacoustics.—H. Harz (pp. 199–202). 60 references.

(*d*) Radar and Radio Navigation.—W. Stanner (pp. 202–203). 26 references.

(*e*) High Frequency Measurements.—H. Schneider (pp. 204–205).

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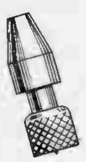

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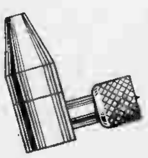

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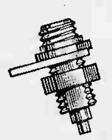

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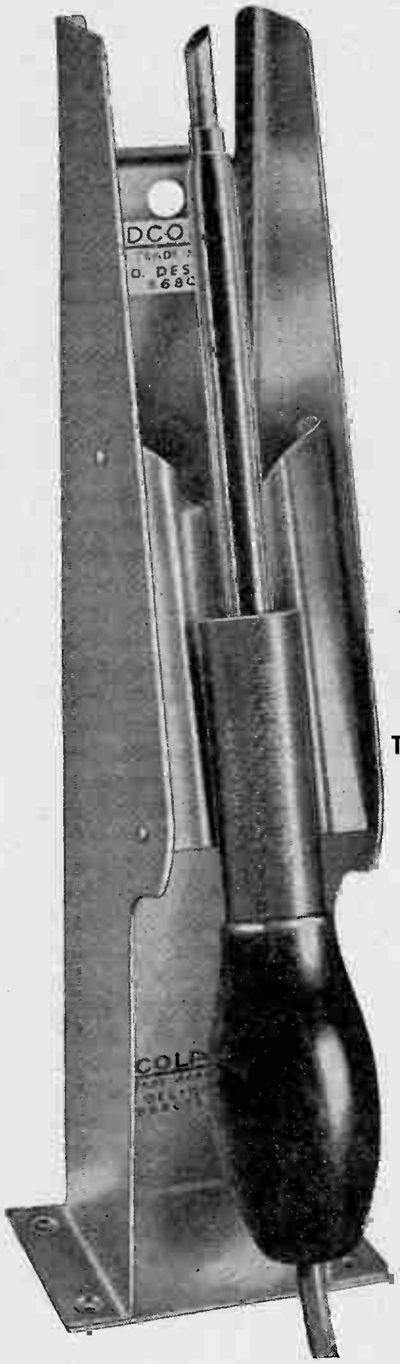

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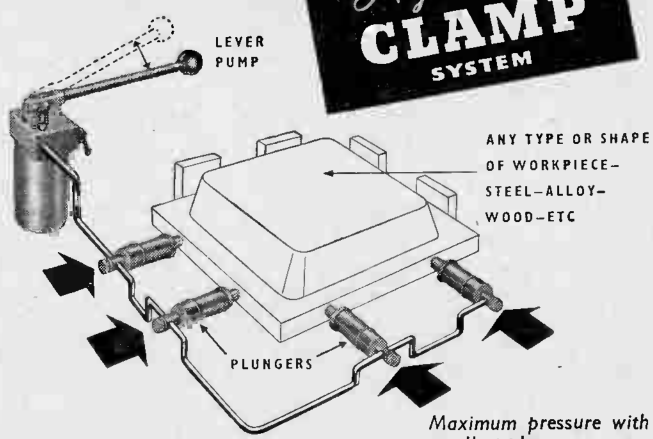

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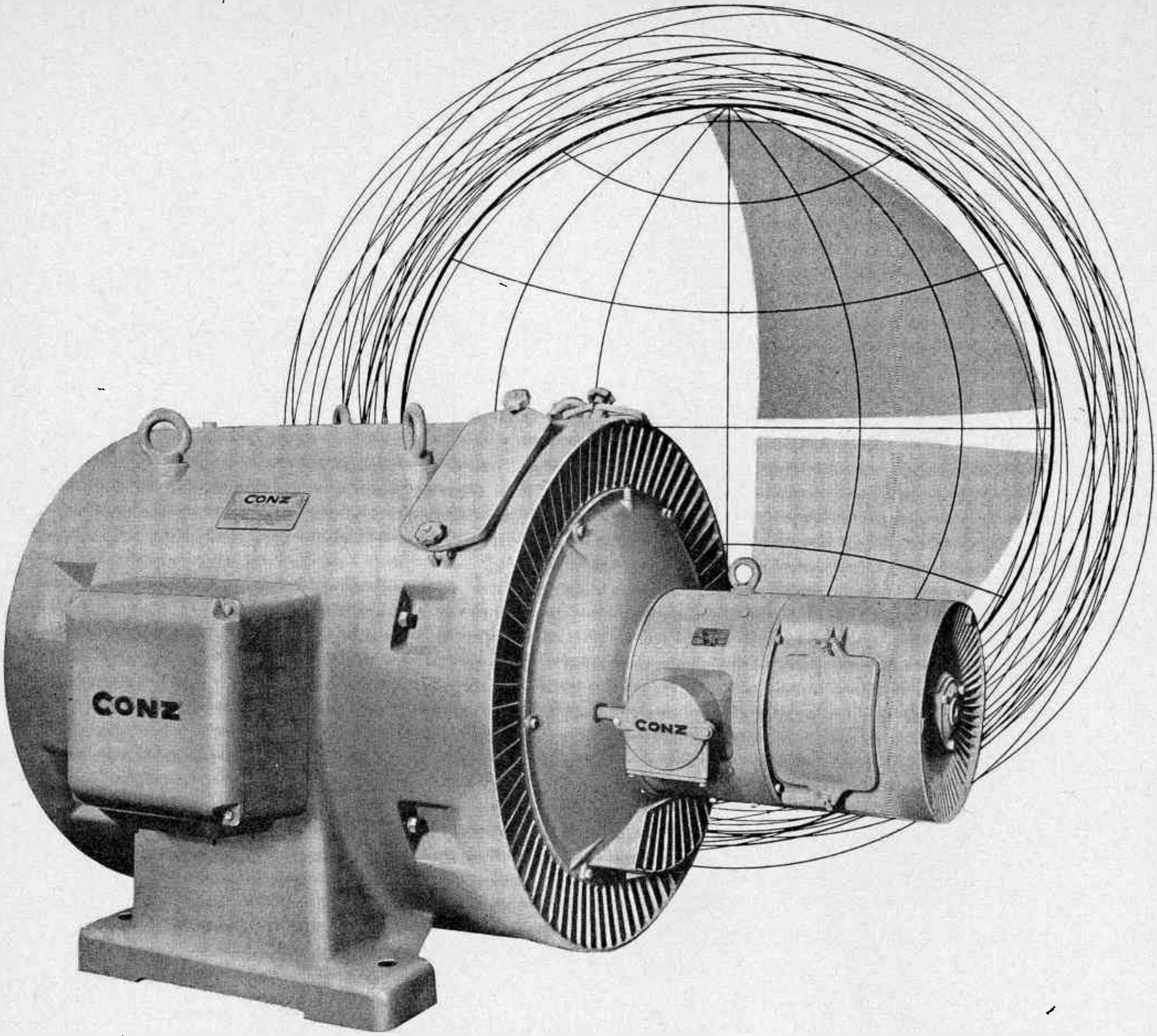
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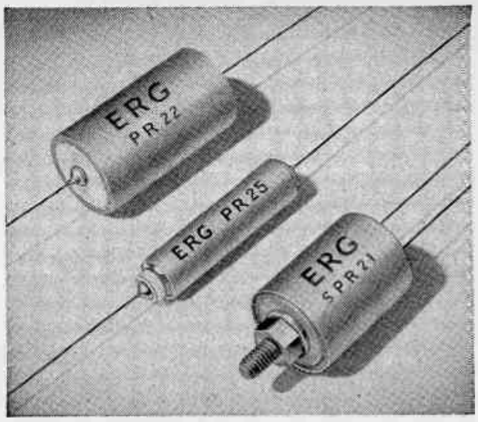

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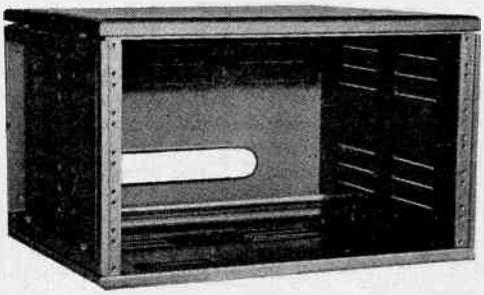

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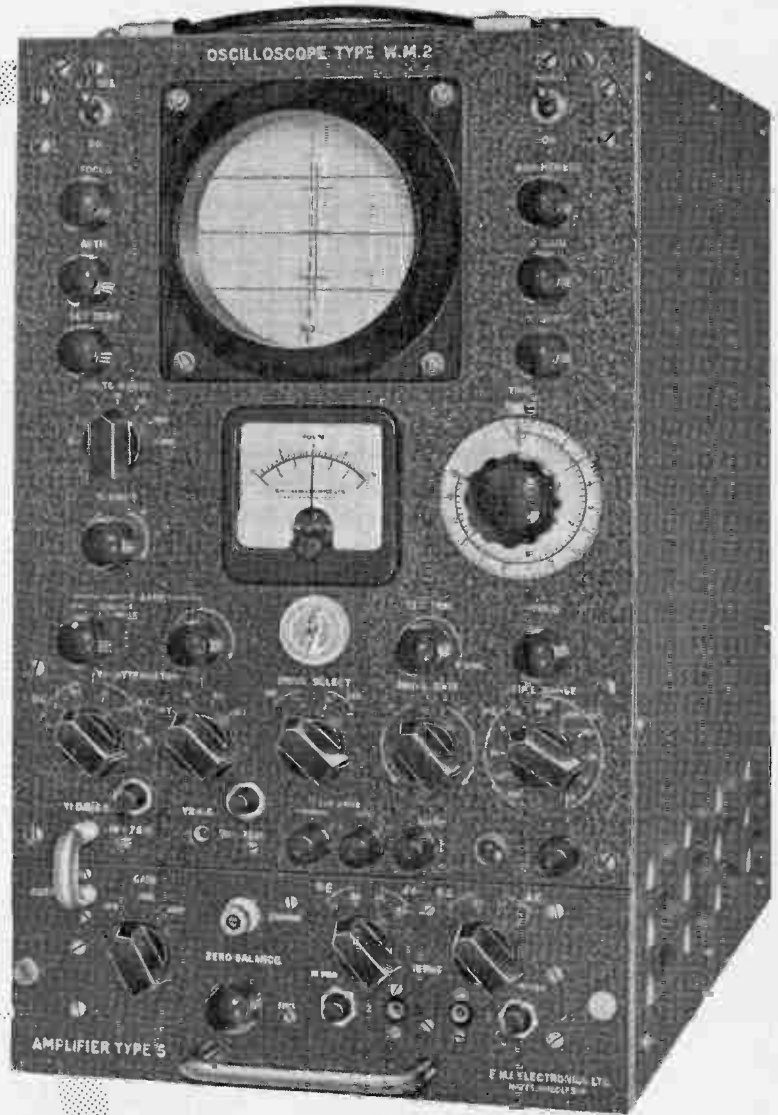

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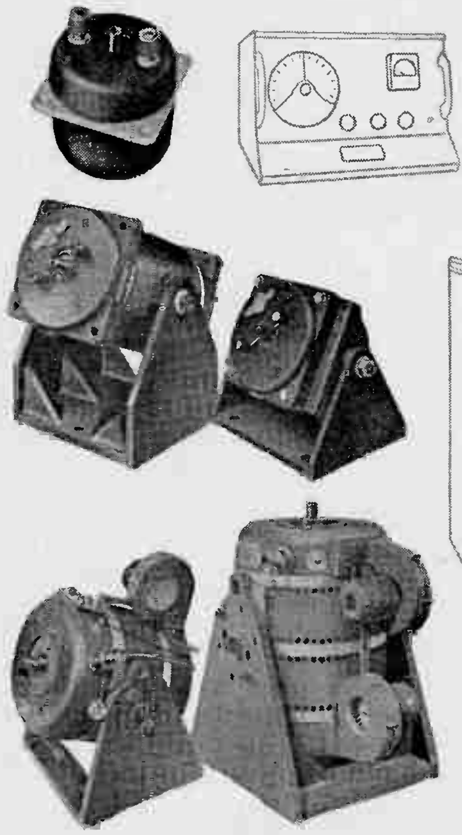

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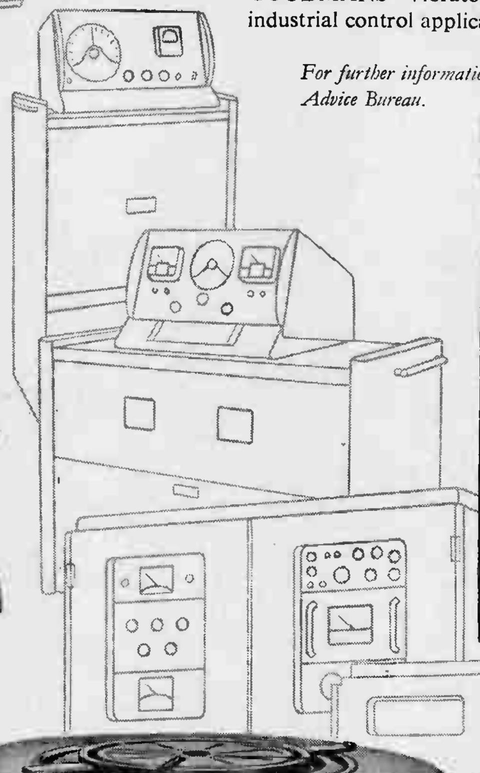

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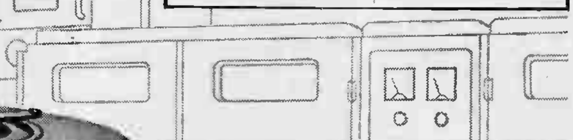

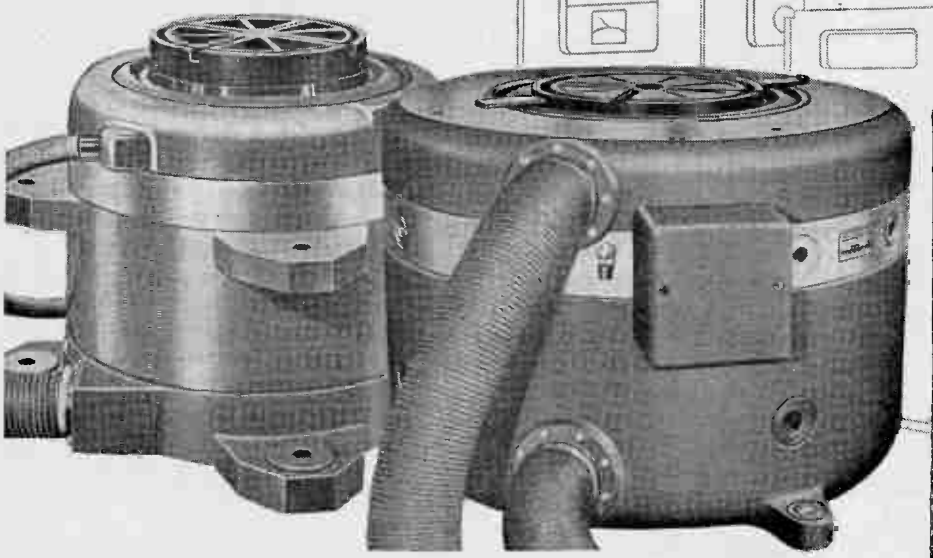

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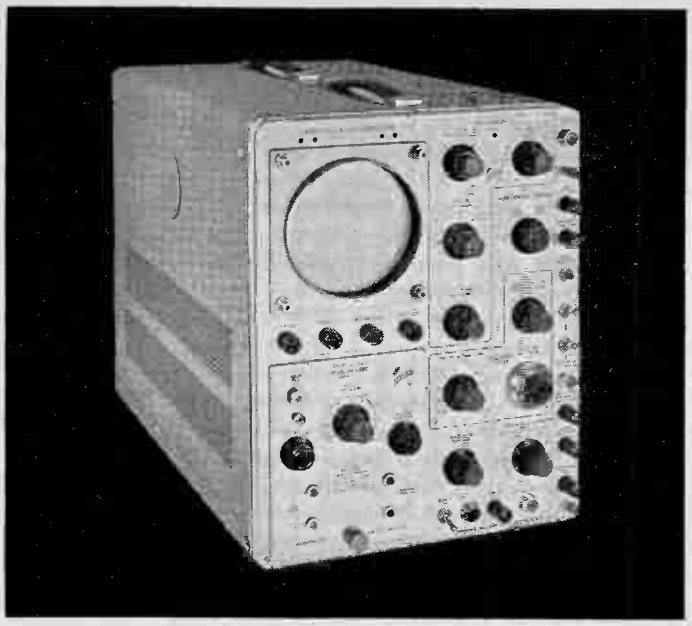

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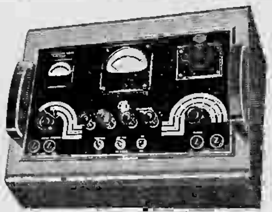
The Y amplifier is direct coupled for sensitivities from 150mV to 300V/cm peak to peak and has a rise time of 0.05μs.

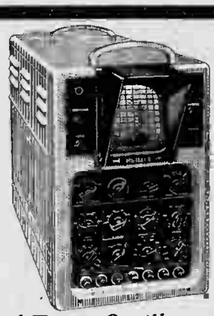
For sensitivities between 4.5mV and 150mV/cm the amplifier is a.c. coupled (useful frequency range 5c/s to 10Mc/s) and has a rise time of 0.06μs.

Signals are applied to a high impedance probe with variable backing-off facilities so that selected portions of the input waveform can be expanded for examination.



Transistor Tester
L264

This compact Transistor Tester has been developed to provide a rapid means of checking general characteristics and ascertaining if a transistor has been damaged in circuit. Facilities for measuring the important parameters, α', base-collector short-circuit current gain, as well as $I'c(o)$ and the collector turnover voltage are provided. Although designed as a simple testing unit, it has an accuracy of ±5%.



Dual Trace Oscilloscope
L101/2

This high grade, general purpose instrument incorporates two identical amplifiers with sensitivities of 20mV/cm over a bandwidth of 4Mc/s. Voltage measurements (±5%) and time measurements (±10%) are by null methods. The time base may be free running or triggered, and sweep speeds are from 0.1μs/cm to 10ms/cm.

Precision Pulse Generator
L141

The pulse generator type L.141 provides two main pulses with accurately controllable amplitude, duration, p.r.f., and separation, together with additional pulses for triggering or synchronising.

The amplitude of the two main pulses is continuously variable between 300μV and 100V, their duration from 0.5 to 100μs, and the separation between them from 0 to 100ms.

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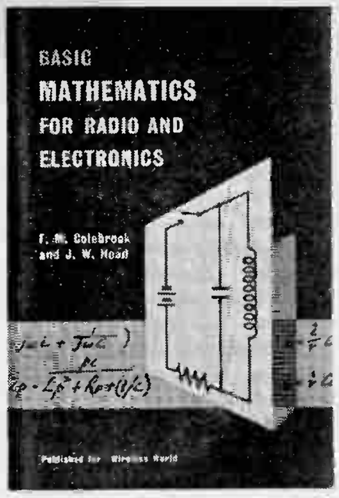

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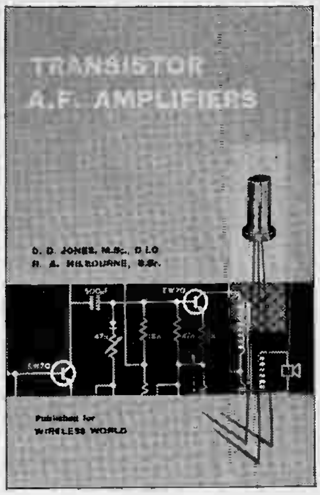

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ELECTRONIC & RADIO ENGINEER

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MINISTRY OF SUPPLY RESEARCH AND DEVELOPMENT ESTABLISHMENTS mainly in southern half of England, require (*a*) Senior Scientific Officers (minimum age 26) and (*b*) Scientific Officers for work in physics, electronics, electrical or mechanical engineering, applied mathematics, chemistry or metallurgy. First or second class hons. degree or equivalent required and for S.S.O. at least three years' post-graduate experience. Starting salary in range (*a*) £1,130–£1,330, (*b*) £595–£1,050 (male, in provinces). Rates for women somewhat lower but reaching equality in 1961. Superannuable under F.S.S.U. Opportunities may occur for those under 32 to compete for established posts. Candidates should indicate fields of work in which interested. Houses available for letting to married staff, and opportunities for new graduates to have workshop training at National Gas Turbine Establishment, Pyestock, Hants. Forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1. (Quoting A.182/8A). [1209

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WITH first-rate technical background in radio communications and electronic development work to take charge of Engineering Dept. of established progressive company in West Middlesex area. Applications are invited from qualified electrical engineers able to control and direct the work of a medium size development engineering group including drawing office and laboratory staff. Previous experience providing a good fundamental knowledge of small transmitter-receiver practice would be an asset; also some development experience with precision electro-mechanical devices. The over-riding requirement however is a sound theoretical knowledge of electronics coupled with organizing ability for control of design and engineering aspects enabling successful manufacture of products.

GOOD salary commensurate with qualifications and experience will be offered. Pension and Life Assurance Scheme.

ADDRESS applications in confidence to Managing Director giving full details, age, qualifications, salary required, education and experience, c/o Box 5D Q6967, A.K. Advertising, 212A Shaftesbury Avenue, London, W.C.2. [1211

ELECTRONIC Engineers required to fill one Senior and one Junior position in Test Engineering Department, concerned with design and maintenance of Production Test Equipment. New projects involving VHF, Transistors and Printed Circuits demand expansion of this Department, with engineers preferably experienced in one or more of these fields.

POSITIONS are permanent and pensionable. Applications in writing in first instance to Employment Officer, The Ever Ready Co. (GB), Ltd., Radio Division, Park Lane, Wolverhampton. [1202

SENIOR and Junior Engineering Inspectors are required by an Establishment in North Bucks, for specialised work relating to Telephone, Electronic, Radio and Audio Frequency Engineering at home and abroad.

EXTENSIVE overseas travelling, for short duration tours, is involved.

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SALARY: Senior £974 to £1,180. Junior £820 to £947. Plus £50 p.a. special allowance and appropriate subsistence rates when travelling. Apply Box No. 6067. [1203

SENIOR Scientific and Scientific Officers, Experimental and Assistant Experimental Officers required at Government Communications Headquarters, Cheltenham, for telecommunications and electronic research in following fields: VHF receiver design and microwave techniques; propagation trials; applications of semi-conducting devices; high speed electronic switching; data storage and handling; computer programming, and operational research. Candidates must normally be natural born British subjects of natural born British parents. For SSO/SO 1st or 2nd class hons. degree in physics, mathematics or engineering or equivalent, at least three years' post-graduate experience for SSO. For EO/AEO Pass Degree, H.N.C. or near equivalent. Salaries for men (provincial rates) S.S.O. £1,130–£1,330; S.O. £595–£1,050 (according to age and qualifications); E.O. £920–£1,130; A.E.O. £370–£800 (according to age). Appointments unestablished but opportunities to compete for established posts. Forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting A147/8A. [1200

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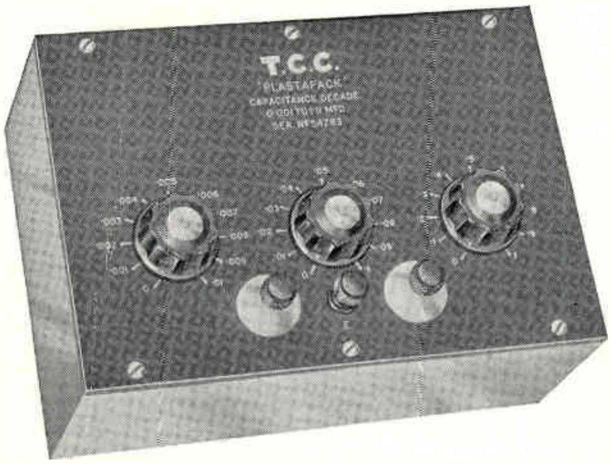
TCC **TEST INSTRUMENTS**

PLASTIC FILM DECADE CONDENSER



Requirements for more accurate measurements and the demand for standard condensers with very high insulation resistance and excellent power factor have created a need for this new product. It is particularly useful in research and development work on computors and integrator circuitry and low level A.C. amplifiers. Stability of capacity and constancy of power factor as a function of frequency also make it extremely useful in measuring circuits, and as a component in filters and tuned circuits.

Condenser elements are non-inductively wound and heat cycled and so stabilized. The plastic film is specially purified high-molecular weight polystyrene, having high IR and freedom from polarization.

New, very low capacity switches, together with high-grade moulded polythene terminations ensure high performance under the most stringent conditions.

The 0.001 μF decade employs ten separate condenser units, while the two larger decades are made up of four units whose capacities are in the ratio of 1:2:3:4 being connected in parallel combinations; **PRICE £52** *(strictly nett)*

Write for leaflet

Features: High insulation resistance ; low dielectric absorption ; capacity and power factor vary only slightly over wide frequency range ; excellent long-time stability.

Capacity Range: 0.001 μF to 1.11 μF in steps of 0.001 μF. The three decades are in steps of 0.001, 0.01 and 0.1 μF.

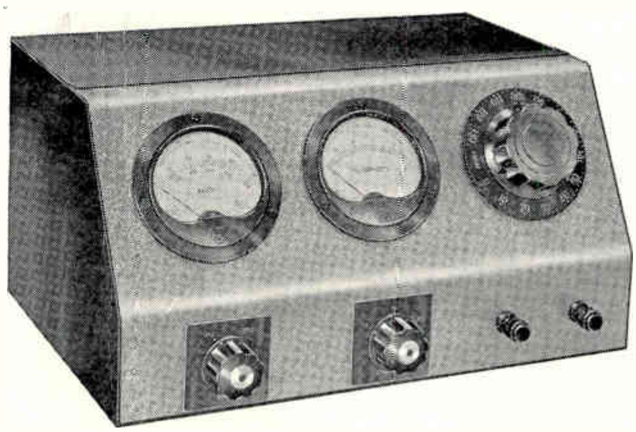
Working Voltage: 350v. D.C. **Test Voltage:** 700v. D.C.

Accuracy: Better than ± 1% at 20°C at 1 k/Cs. Decades with closer tolerance can be produced if required.

Power Factor: Better than 0.0005 in the audio-frequency range.

Insulation Resistance: Better than 500,000 MΩ, at 350 volts at 20°C.

REFORMING & LEAKAGE TESTER



An electrolytic condenser which has been in store (or out of use) for six months or more will initially pass a high leakage current when its working voltage is applied. This Test Instrument has been produced to make possible the reforming of such condensers to their normal value. It can also be used for leakage measurements.

Specification

Mains input: 110, 220, 230, 240V at 50/60 cycles.
Output: 2 Voltage Ranges, 0 - 100V, 0 - 700V.
Current rating: 100 mA. D.C.
The voltages are continuously variable over both ranges, with a voltmeter calibrated 0 - 100V D.C. and a second range 0 - 1 kV. D.C.
Current meter: with 3 ranges: 0-1 mA D.C., 0-10 mA D.C., 0-100 mA D.C.

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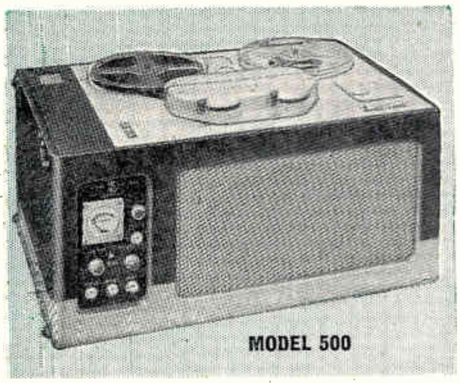
Reflectograph

TAPE RECORDERS AND CONTINUOUS PLAYERS FOR INDUSTRY



MODEL 500

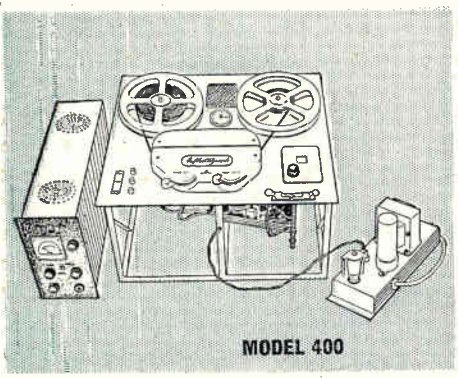
THE REFLECTOGRAPH professional tape recorder previously supplied in limited quantities to Broadcasting authorities, Recording Studios and Laboratories, is now being manufactured by Multimusic Ltd.

The Reflectograph Model 500 is self-contained. The portable duo-tone case is finished in luxan hide and pigskin colours and is complete with an output amplifier and two matched loudspeakers.

Model 501 has the same technical specification but is supplied in a metal case for industrial use.

Reflectograph 400 is supplied in three units for incorporation in laboratory equipment. It comprises the Reflectograph deck mounted on a stand, combined record amplifier and playback pre-amplifier and power pack. The units complete with inter-connecting leads have been designed for operation in conjunction with leading makes of high fidelity amplifiers.

Reflectograph Model 550 has the same technical specification as Model 500 with an additional replay amplifier to provide stereophonic reproduction from tape and from records if a suitable pick-up is connected. The fine wood case does not incorporate the loudspeakers.

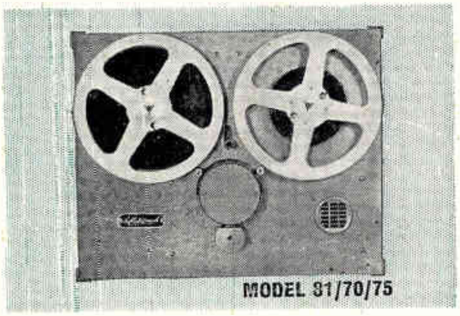


MODEL 400

THE REFLECTOGRAPH IS THE ONLY RECORDER IN THE WORLD POSSESSING ALL THESE FEATURES

- Fitted with 3 heads, separate record and replay amplifiers, enabling instant comparison to be made between signal recorded on tape and the input.
- Variable speed between 8 and $3\frac{1}{2}$ i.p.s. Stroboscope, lit by neon lamp, shows precise speeds of $7\frac{1}{2}$ and $3\frac{3}{4}$ i.p.s.
- Easy tape threading into a straight slot. Provision for conversion to stereo. Lever deck controls, providing variable speed wind forward and back from extra fast to inching for editing; sound available for editing if required; instant stop and start.
- Peak level recording meter; Push-button record-playback controls with record safety latch; Clock-type tape position indicator; 3 Garrard motors; 2 matched loudspeakers; Accommodates up to $8\frac{1}{4}''$ reels.
- 3 watts undistorted output; overall response strictly to C.C.I.R. recommended specifications; 2 input and 2 output sockets. Fitted with Bib tape splicer on deck, complete with reel of tape, spare reel, 2 screened jack plugs.

Model 400 specification is similar to above, excluding 3 watts output, 2 loudspeakers and 1 output socket. Additional facilities include sockets on chassis for radio and pick-up; socket for microphone on instrument panel, where an additional switch provides instant selection of 3 inputs.



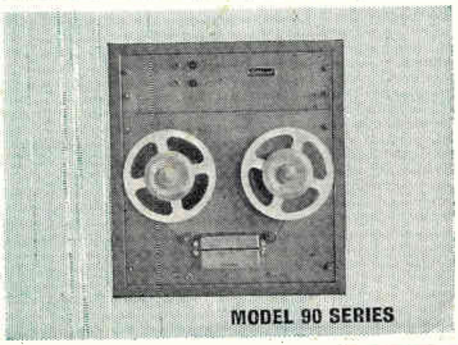
MODEL 81/70/75

CONTINUOUS TAPE REPRODUCERS FOR BACKGROUND MUSIC AND AUTOMATION

The Reflectograph Continuous Players are probably the first British Made heavy duty machines in quantity production, specially designed to play recorded tapes continuously. Tapes are easily threaded and the machine may be started and stopped manually, remotely or by a clock. The tape is played down one track, automatically reverses and continues to play on alternate tracks until switched off.

Model 81/70/75 operates at $3\frac{3}{4}$ i.p.s. and plays for up to 2 hrs. 8 mins. on each of two tracks. By means of a 20 cycle note, recorded at the end of each track, the machine automatically reverses. A 3 watt amplifier is incorporated but a high level output is available. Start and stop is electrically controlled so that remote or clock operation is available.

An alternative model incorporating a recording head and amplifier is available with a selective amplifier for the recording and reproduction of tones for instrumentation and automation.



MODEL 90 SERIES

Model 90 series, made to special order, provides all the above facilities with reversing by note or light with capacity of up to 12 hours playing time before repeating. These machines are of the standard size for rack mounting and can be supplied to operate at other speeds than $3\frac{3}{4}$ i.p.s.

For full information please write to:

REFLECTOGRAPH DIVISION, MULTIMUSIC LTD.,
MAYLANDS AVENUE, HEMEL HEMPSTEAD, HERTS.