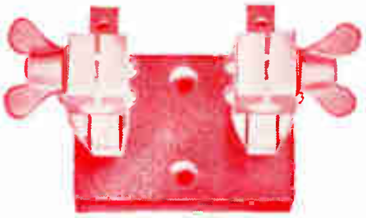
WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

AUGUST 1952

VOL. 29 No. 347 THREE SHILLINGS AND SIXPENCE



EDISWAN
SPECIAL PURPOSE
VALVES

Of particular interest

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The ES.833 is a high mu triode particularly suitable for use as an R.F. Power Amplifier, Oscillator or Class B Modulator. It is a direct plug-in replacement for the American type 833A.

The anode and grid connections are brought out at the top and are taken through metal-to-glass seals to heavy current terminals. As a result of this construction the valve is exceptionally efficient at higher radio frequencies, and may be operated under Class 'C' CW conditions at a maximum input of 2 kW at frequencies up to 30 Mcs. At a reduced input rating it is possible to operate the valve as high as 75 Mcs.

RATING		RADIATION COOLED	AIR COOLED
Filament Voltage (volts)	Vf	10.0	
Filament Current (amps)	If	10.0	
Maximum Anode Voltage (volts)	Va (max)	3000	4000
Maximum Anode Dissipation (watts) Radiation cooled	Wa (max)	300	
Maximum Anode Dissipation (watts) Forced Air cooled	Wa		400
Amplification Factor	μ	35	
Maximum Operating Frequency at full rating		30 Mcs.*	

** Operating frequency at reduced ratings up to 75 Mcs.*

Prices and technical data upon application.

EDISWAN

THE EDISON SWAN ELECTRIC CO. LTD., 155 CHARING CROSS RD., LONDON, W.C.2
Member of the A.E.I. Group of Companies
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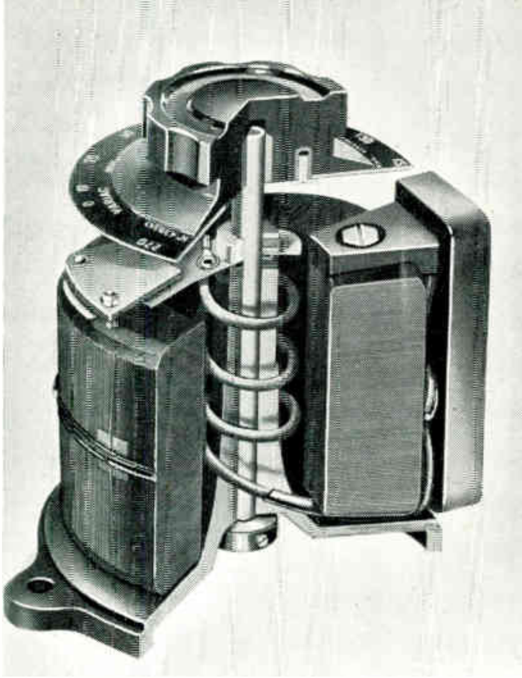

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SPECIFICATIONS

SERIES “50” Variacs

TYPE	LOAD RATING	INPUT VOLTAGE	CURRENT RATED	CURRENT MAXIMUM	OUTPUT VOLTAGE	NO-LOAD LOSS	NET PRICE £ s. d.*
50-A	5 kva.	115 v.	40 a.	45 a	0-135 v.	65 watts	44 18 6
50-B	7 kva.	230/115 v.	20 a.	31 a	0-270 v.	90 watts	44 18 6

SERIES “100” Variacs

TYPE	LOAD RATING	INPUT VOLTAGE	CURRENT RATED	CURRENT MAXIMUM	OUTPUT VOLTAGE	NO-LOAD LOSS	NET PRICE * £ s. d.
100-K	2000 va.	115	15 a.	17.5 a.	0-115	20 watts	17 17 0
100-KM	2000 va.	115	15 a.	17.5 a.	0-115	20 watts	18 12 0
100-L	2000 va.	230/115	8 a.	9 a.	0-230	25 watts	17 17 0
100-LM	2000 va.	230/115	8 a.	9 a.	0-230	25 watts	18 12 0
100-Q	2000 va.	115	15 a.	17.5 a.	0-135	20 watts	18 9 0
100-QM	2000 va.	115	15 a.	17.5 a.	0-135	20 watts	19 4 0
100-R	2000 va.	230/115	8 a.	9 a.	0-270	30 watts	18 9 0
100-RM	2000 va.	230/115	8 a.	9 a.	0-270	30 watts	19 4 0
100-LH	1200 va.	480/240	2 a.	2.5 a.	0-480	25 watts	21 15 0
500-L ⊙	1450 va.	180	8 a.	9 a.	0-180	25 watts	17 17 0
2000-K †	1400 va.	125	8 a.	9 a.	0-125	25 watts	17 17 0

⊙ For 500 cycles. † For 2,000 cycle service.

SERIES “200” Variacs

TYPE	LOAD RATING	INPUT VOLTAGE	CURRENT RATED	CURRENT MAXIMUM	OUTPUT VOLTAGE	NO-LOAD LOSS	NET PRICE £ s. d.*
200-CM 200-CU	860 va.	115 v.	5 a.	7.5 a.	0-135 v.	15 watts	7 17 6 6 15 0
200-CMH 200-CUH	580 va.	230 v. 115 v.	2 a. 0.5 a.	2.5 a. 2.5 a.	0-270 v. 0-270 v.	20 watts 20 watts	9 15 0 8 5 9

*All ‘VARIAC’ prices plus 20% as from 23rd Feb. 1952

Write for catalogue V549 which gives full details of ‘VARIAC’ transformers and suggestions for use, as well as data on other special patterns.

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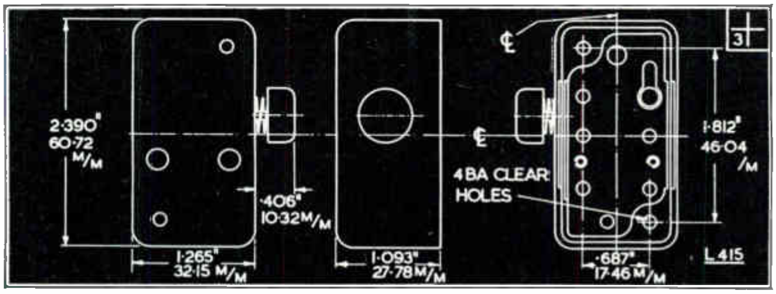
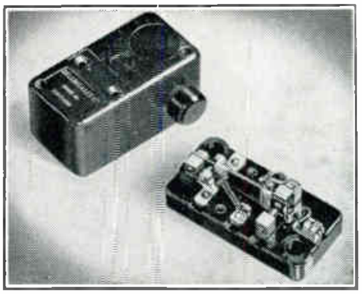
The "Belling-Lee" page for Engineers





THERMAL DELAY SWITCHES

LIST NUMBERS

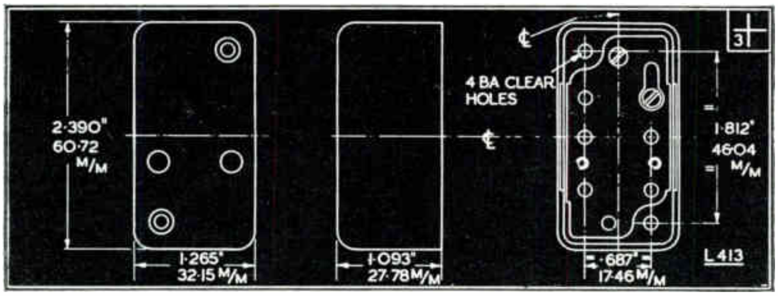
L.413 L.415

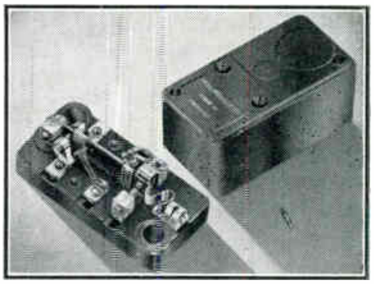
OTHER TYPES AVAILABLE

Thermal delay switches for use with motor drives, electrical tools, etc.; also for application to non-critical process timing. The type illustrated above (List Number L.415) is a manual resetting switch.

Where a self-resetting type is required, these are available under List Number L.413 (illustrated below).

Apart from the resetting mechanisms, the general arrangement and electrical characteristics of the two types are similar. Each is fitted with Standard instrument contacts which will break a maximum current of 4 amps. at 250 volts A.C., or 2 amps. at 50 volts D.C.





They can also be supplied to special orders, to break up to 20 amps. at 250 volts A.C. or 5 amps. at 100 volts D.C.

The maximum continuous rating is 10 amps.; normal heater loading for continuous operation, 3-4 watts at any voltage up to 50 volts A.C. or D.C.

For occasional short duration functions, a heater loading up to 25 watts is permissible. The heater may be in series with the main contacts, or independent, being insulated for 250 volts A.C. working potential to bi-metal.

FURTHER DETAILS ON APPLICATION

Please write for illustrated pamphlet Ref. P.377/W.E.

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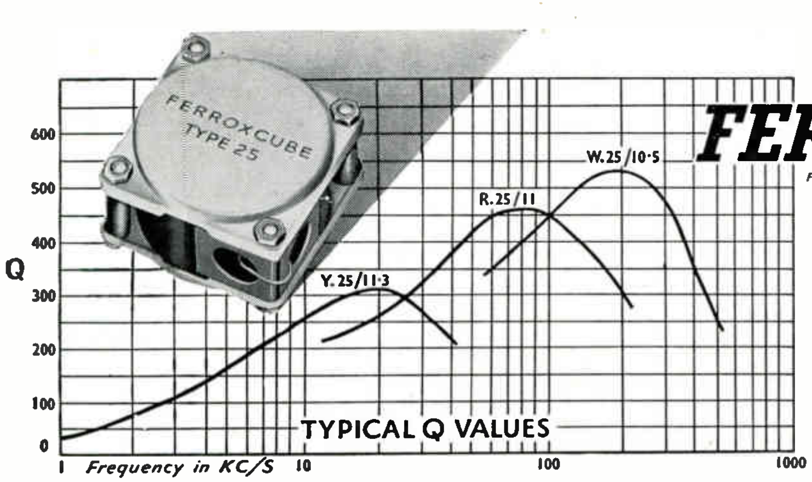

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FERROMAGNETIC FERRITE

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In the design of Mullard pot core assemblies types 25 and 36 full advantage is taken of the characteristics of Ferroxcube to produce inductances of remarkably high "Q" factors. This, combined with ease of winding, makes these cores very suitable for use in filter networks and wherever high quality inductances are required.

Fine adjustment of inductance is obtained by control of the air gap rather than by variation of the turns.

The good screening properties of the Ferroxcube and the convenient shape of the assemblies, which allows stacking or individual mounting, are other features which distinguish these Mullard cores.

OUTSTANDING FEATURES

★ Low hysteresis coefficient
★ High values of inductance
★ Low self capacitance
★ Controllable air gap facilitating inductance adjustment
★ Self screening
★ Controlled temperature coefficient
★ Operation over a wide frequency range
★ Ease of winding and tapping
★ Easily mounted

PLEASE WRITE FOR FULL DETAILS

Mullard

Mullard FERROXCUBE

FERROMAGNETIC FERRITE

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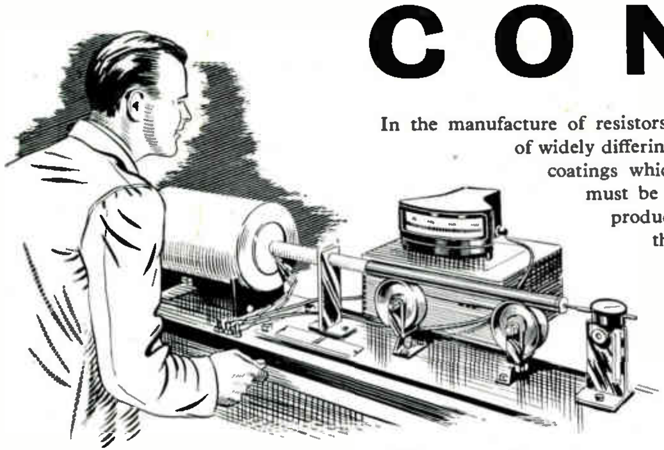
(MF372)

CONTROL



In the manufacture of resistors we consider thermal expansions at high temperatures—of widely differing materials such as ceramics, resistance wires and protective coatings which comprise the system. At high temperatures a balance must be achieved to eliminate any stress/strain relationships in the product at room and at working temperatures. To this end thermal expansions of ceramic cores and protective coatings are accurately measured by means of a Dilatometer. Precise knowledge of this fundamental character ensures maximum reliability and stability and is evidence of our insistence upon 100% efficiency throughout production.

Depending upon the duty of a resistor, special protective coatings are available :

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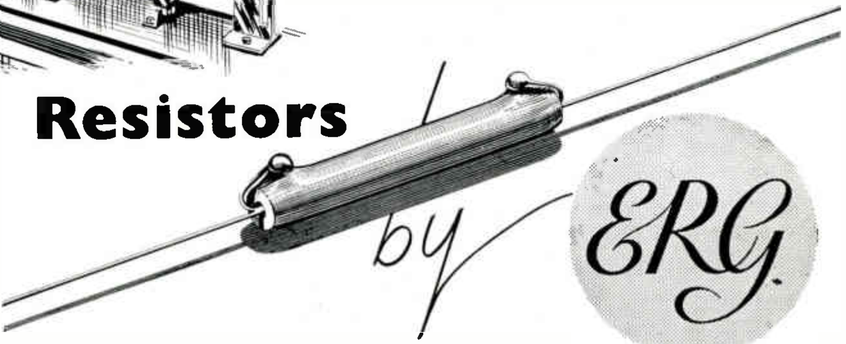
Resistors



(Type Approved for all Service Departments)

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World Radio History

G.E.C.

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FOR THE MEASUREMENT OF CRYSTAL PERFORMANCE

The G.E.C. Quartz Crystal Activity Test Set measures the equivalent parallel resistance of a quartz crystal when oscillating in a circuit having an input capacity of either 20 pF, 30 pF or 50 pF, the alternative capacities being selected by a switch.

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The accuracy of the loss dial calibration is ± 2%.

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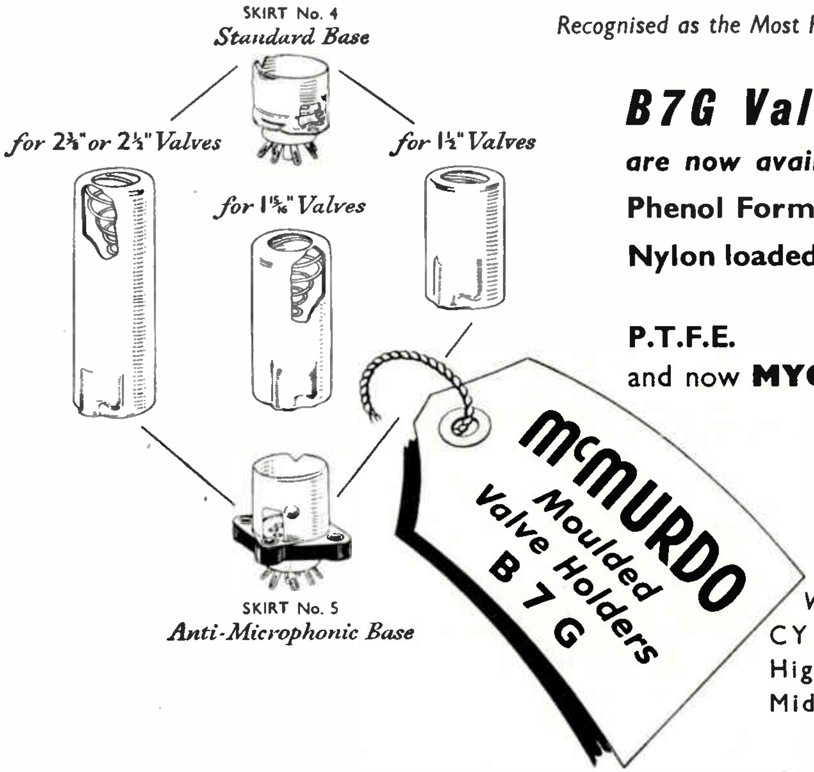

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Special units can be made to any specification, and although the demand for our products is continually increasing we can still offer reasonably prompt delivery.

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Almost any pattern comprising pure line, halftones or a combination of both can be supplied on receipt of specific requirements, and two standard types are available.

Type J.101 — Test Chart "A"
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TYPICAL OPERATING DATA

Deflection	*electromagnetic*
Focus	*electrostatic*
Vh	6·3 *V*
Vg (cut-off)	−50 *V*
Va1	1200 *V*
Va2 (focus)	800/850 *V*
Va3 (wall)	1200 *V*
V target	1160/1200 *V*
I target	5μ*A*

Resolution better than 500 lines

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A cold advertisement with its printed claims is all very well as an introduction to a product, but how much better a practical test of a manufactured characteristic!

Over 3,000 people passed through our demonstration room at the recent Northern Radio Show held in Manchester. At the end of each demonstration period we received a round of spontaneous applause denoting the obvious pleasure that good quality reproduction can afford to the receptive ear.

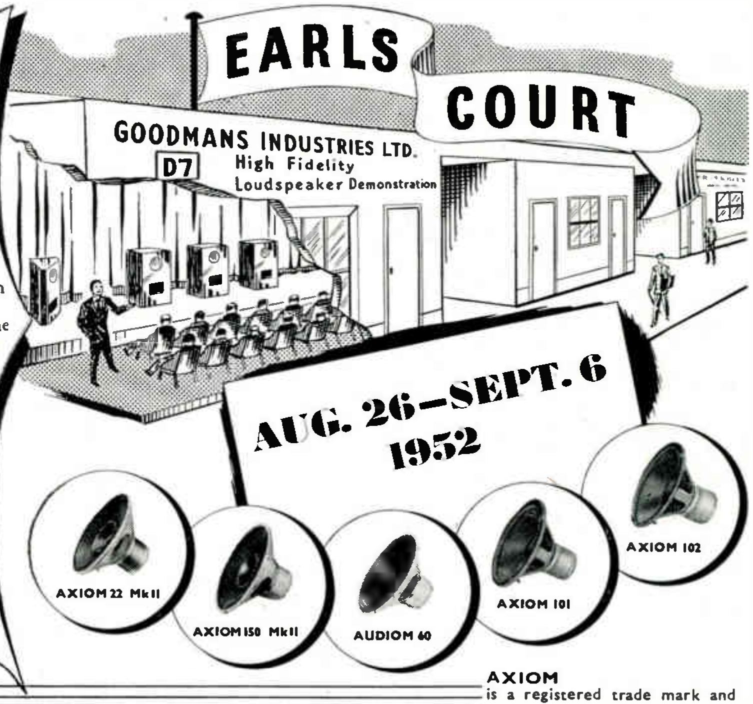
It is not often that the general public is given the opportunity of such a demonstration, and indeed seldom that a printed claim can be adequately proved to the entire satisfaction of all concerned.

At this year's "Radio Earls Court," we of the House of Goodmans once again have the pleasure of presenting for your enjoyment a practical demonstration of our high fidelity reproducers.

The reproducers demonstrated will include the well-known Axiom series in its modified form, and, of recent design, two special high fidelity 8 in. units—the Axiom 101 and 102.

We look forward to the pleasure of your company.

Remember—Demonstration Room No. D.7 (also Stand 41), Earls Court, August 26th to September 6th, 1952.



A TRADITION OF EXCELLENCE · FIDELITY · EFFICIENCY

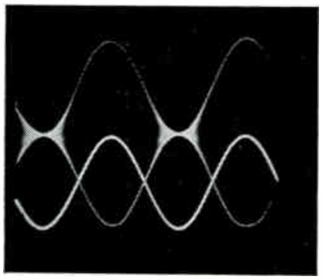
GOODMANS

GOODMANS INDUSTRIES LIMITED

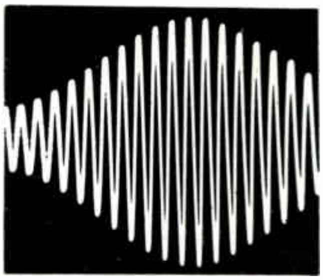
Axiom Works, Wembley, Middlesex

Telephone: WEMbley 1200 (8 lines)

AXIOM is a registered trade mark and denotes our High Fidelity Range of reproducers.
AUDIOM is the registered trade mark that is applied to our good quality commercial reproducers.
GOODMANS is the registered trade mark of all the Company's products.



320kc/s modulated 400c/s; audio on second beam.



24kc/s modulated 60% 1kc/s.

100% MODULATION . . . . A.M. WITHOUT F.M.

Excellent amplitude modulation is a feature of the Standard Signal Generator TF 867—a.m. accompanied by minimum spurious f.m.—less than 100 c/s below 5 Mc/s, 1000 c/s above. Other features are:

Wide range—15 kc/s (or less) to 30 Mc/s in 11 bands on full vision scale.
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Easy tuning—discrimination 1 part in 10,000 on total 15 ft. scale length.
High output—4 volts down to 0.4 microvolts.
Flexible modulation—internal 400 and 1,000 c/s, 0-100% external 50-10,000 c/s ± 2db.

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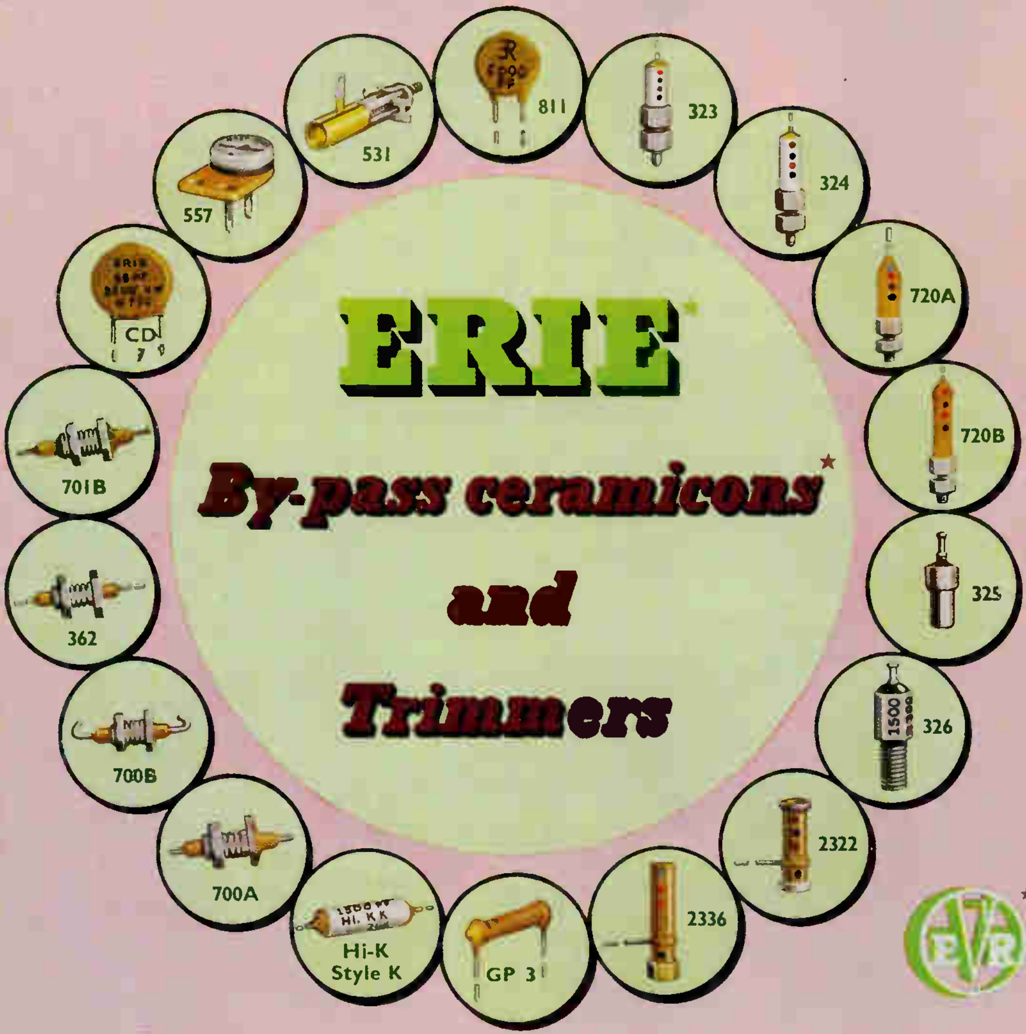
World Radio History

for VHF *applications*



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By-pass ceramicons★ and Trimmers*

ERIE CAPACITORS fulfil all requirements for efficient by-passing at very high frequency—compact design, low inductance and conservative voltage rating.

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Write for complete information to meet your specific requirements.

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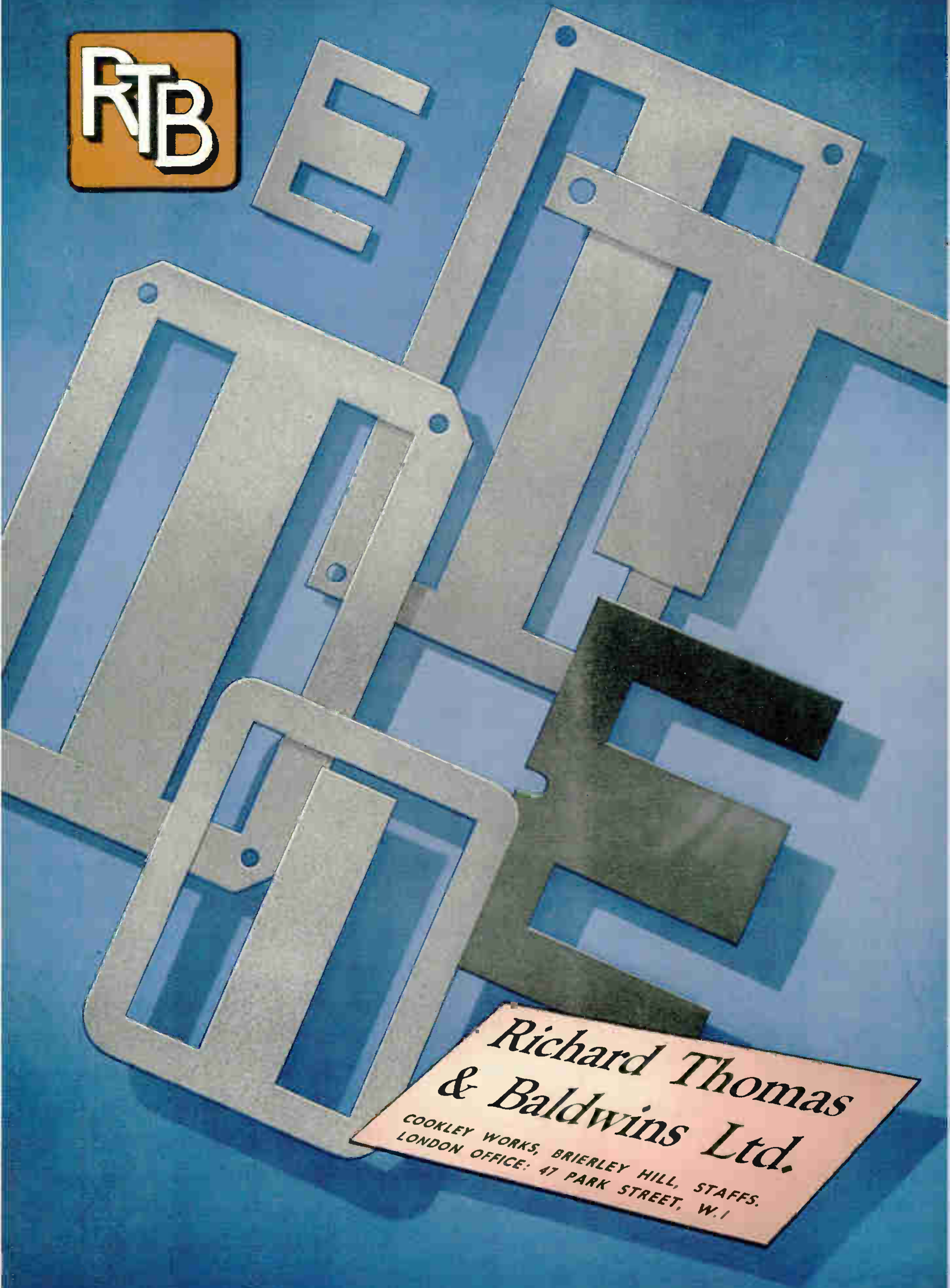
CARLISLE ROAD, THE HYDE, LONDON, N.W.9. Tel: COLindale 8011. *Factories : London and Great Yarmouth ; Toronto, Canada ; Erie, Pa., U.S.A.*

A*



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A NEW VALVE VOLTMETER

Zero and Calibration Stability under all operating conditions

Rivlin Instruments present their new valve voltmeter, Model VV.1. This instrument is characterised by its accuracy and exceptionally high degree of zero and calibration stability, without which accurate voltage indication is of no value.

Model VV.1 permits measurement of A.C. and D.C. potentials in the following ranges: 1, 2.5, 10, 25, 100 and 250 volts f.s.d. Accuracy is better than ± 2% A.C. and D.C. and calibration remains unaffected by normal mains variation.

For high-frequency measurements a low-capacitance probe is incorporated, which remains plugged into the front panel when required for low frequency measurement. Separate terminals are provided for D.C. measurements, either positive or negative, the input impedance being of the high value of 40 megohms.

Construction is to the finest instrument standards. All components are conservatively rated, and only condensers of paper or mica dielectric are used.

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Production has been organised to enable early delivery of the Rivlin Valve Voltmeter, Model VV.1, to be offered. Descriptive leaflet available on request to Dept. 7.

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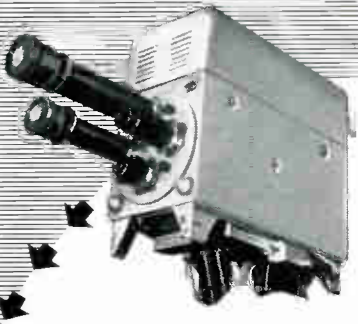
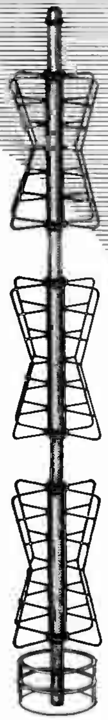


Marconi Image Orthicon Television Camera, Type BD.624.



Marconi lower band super turnstile television aerial.

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transmitting equipment

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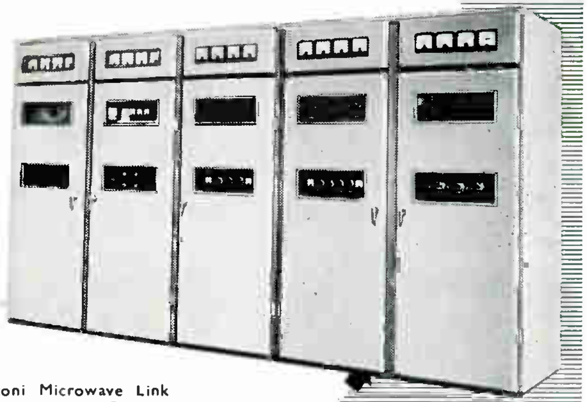
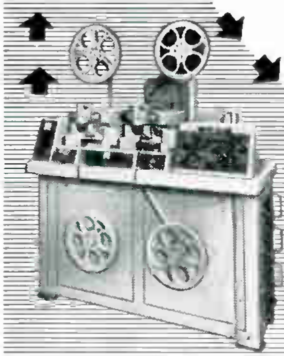
Marconi Picture and Waveform Monitor, Type BD.627 with Camera Control Unit, Type BD.626.

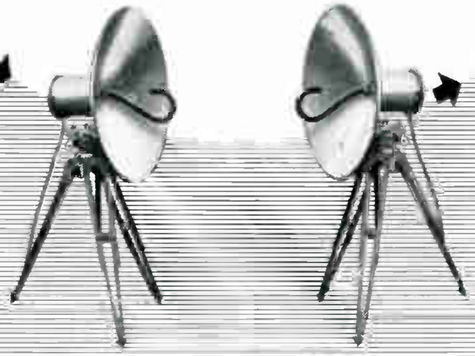
Marconi Four Channel Vision Mixer Type BD.633 with Picture and Waveform Monitor.



Marconi 5kW television transmitter Type, BD.352.



Marconi Image Orthicon Telecine Equipment, Type BD.680.



Marconi Microwave Link Equipment, Type BD.401.

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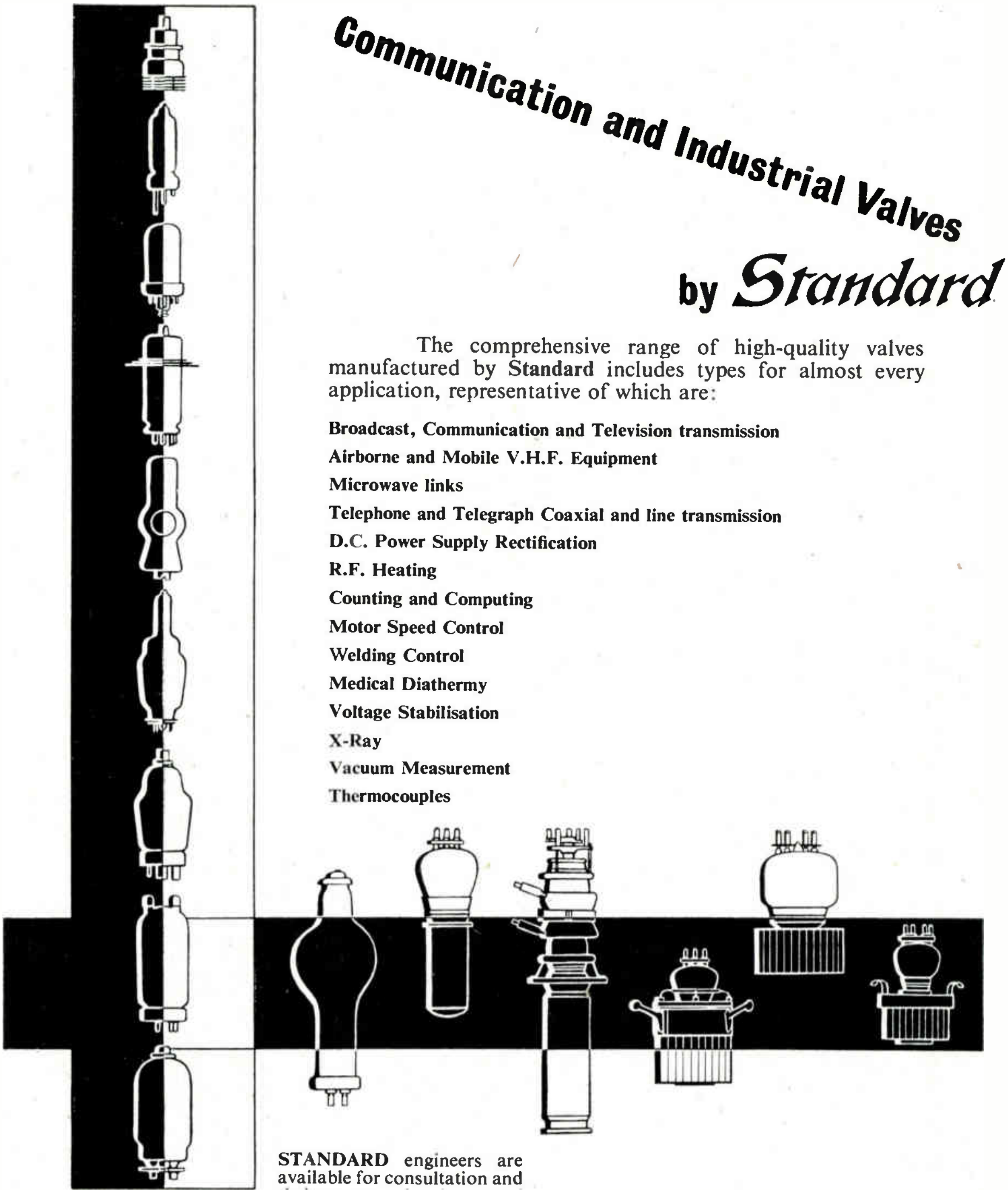

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WIRELESS ENGINEER

The Journal of Radio Research and Progress

Managing Editor: HUGH S. POCOCK, M.I.E.E.
Editor: W. T. COCKING, M.I.E.E. Technical Editor: Professor G. W. O. HOWE, D.Sc., LL.D., M.I.E.E.

Volume 29 · Number 347

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AUGUST 1952

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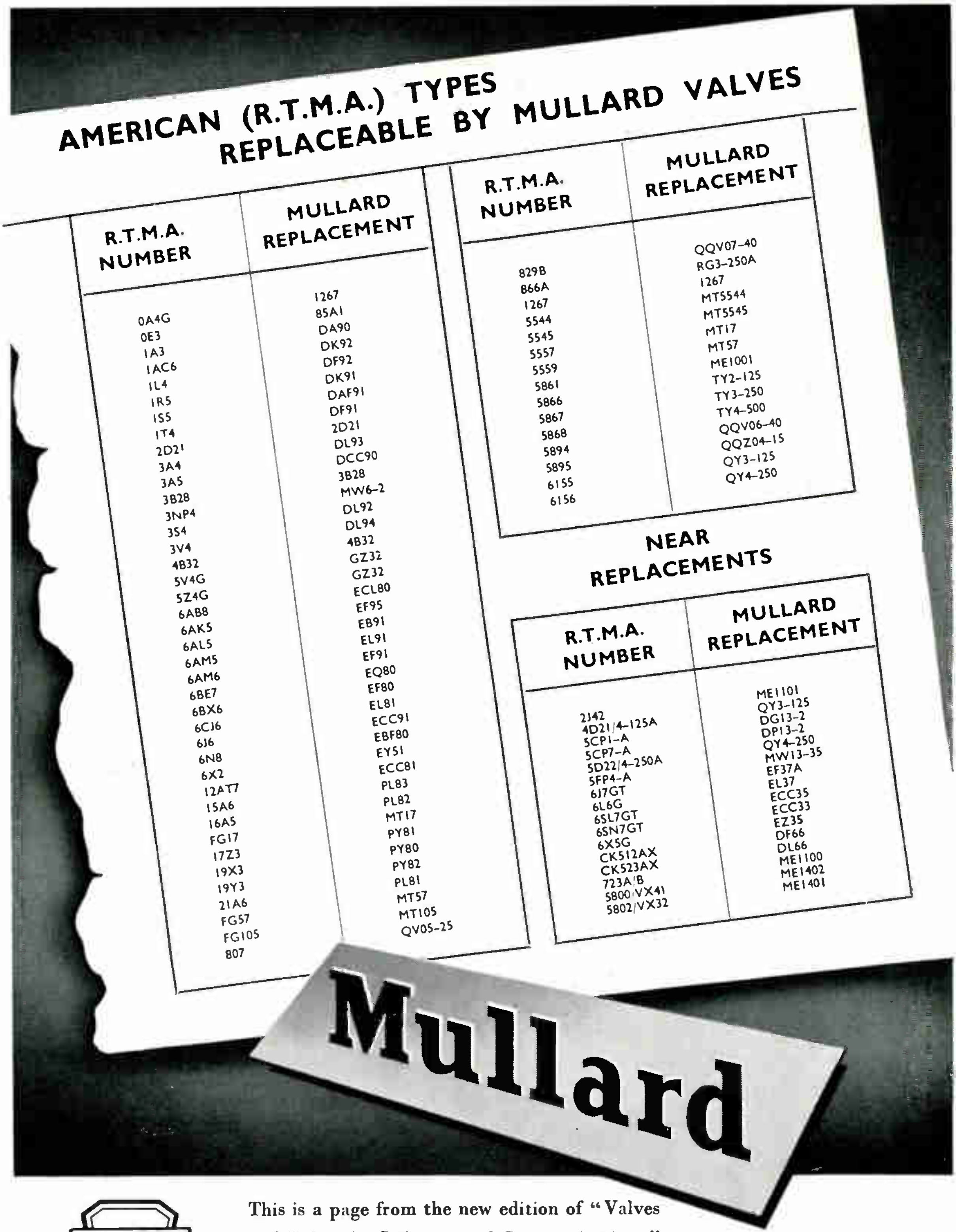

AMERICAN (R.T.M.A.) TYPES REPLACEABLE BY MULLARD VALVES

R.T.M.A. NUMBER	MULLARD REPLACEMENT
0A4G	1267
0E3	85A1
1A3	DA90
1AC6	DK92
1L4	DF92
1R5	DK91
1S5	DAF91
1T4	DF91
2D21	2D21
3A4	DL93
3A5	DCC90
3B28	3B28
3NP4	MW6-2
3S4	DL92
3V4	DL94
4B32	4B32
5V4G	GZ32
5Z4G	GZ32
6AB8	ECL80
6AK5	EF95
6AL5	EB91
6AM5	EL91
6AM6	EF91
6BE7	EQ80
6BX6	EF80
6CJ6	EL81
6J6	ECC91
6N8	EBF80
6X2	EY51
12AT7	ECC81
15A6	PL83
16A5	PL82
FG17	MT17
17Z3	PY81
19X3	PY80
19Y3	PY82
21A6	PL81
FG57	MT57
FG105	MT105
807	QV05-25

R.T.M.A. NUMBER	MULLARD REPLACEMENT
829B	QQV07-40
866A	RG3-250A
1267	1267
5544	MT5544
5545	MT5545
5557	MT17
5559	MT57
5861	ME1001
5866	TY2-125
5867	TY3-250
5868	TY4-500
5894	QQV06-40
5895	QQZ04-15
6155	QY3-125
6156	QY4-250

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WIRELESS ENGINEER

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Ionic Bombardment of Silicon

IN a recent number of the *Bell System Technical Journal** some very interesting experiments were described, which showed how greatly the nature of the surface of silicon can be changed by bombardment with gaseous ions. The technique has been so developed that it is possible to prepare silicon surfaces having a wide range of properties by suitably varying the bombarding voltage, the target temperature, and the time of exposure. The effect of impurities in the silicon is well known but it is surprising to learn that silicon, so contaminated with boron that it shows little rectification, can be modified by bombardment until it is better than most unbombarded materials. As these bombarded surfaces are chemically stable and show no aging effects, the process is undoubtedly of great practical importance in the field of signal rectifiers, transistors, etc.

The material to be bombarded is fused in silica crucibles, and then, when solidified, is ground to a cylinder about 1·5 inches diameter from which thin wafers are cut and polished on one side. The bombarding chamber is a vertical cylinder, near the bottom of which is a graphite circular plate on which the specimen is placed with its polished side upwards. Under the graphite holder there is an induction heater and a thermo-couple, so that the specimen can be raised to any desired temperature. There are arrangements for evacuating the chamber and admitting the desired gas.

Near the top of the chamber is the tungsten cathode and the grid. The grid is maintained positive with respect to the cathode, and many of the emitted electrons pass through the grid but then slow up and return to the grid. Due to the impact between electrons and the gas molecules positive ions are produced and these are accelerated downwards to the silicon specimen and its graphite support which are maintained at a negative voltage which can be varied from 100 to 30,000 V. The temperature of the specimen has been varied between 20°C and 400°C; results show that it should be above 250°C but otherwise it is not important. A suitable current is about 100 μA which corresponds to about 5 μA per square centimetre of target area. The bombardment should be maintained for one to two minutes.

After removal from the chamber the rough back of the wafer was covered with a thin layer of evaporated rhodium, and it was then cut into $\frac{1}{8}$-in squares, a convenient size for testing. Four different gases have been tried, viz., hydrogen, helium, nitrogen, and argon, the atomic weights of which are 1, 4, 14 and 40 respectively. They all worked well but helium was the easiest to handle.

Optical examination of the surface after bombardment shows that the selenium has undergone considerable change of structure, and the results observed with different gases are strikingly different. A strip of the surface was screened by a mask of nichrome ribbon 5 mils wide and 1 mil thick. After prolonged bombardment by helium the adjacent unscreened surface may be elevated by as much as $2{\cdot}25 \times 10^{-6}$ cm, whereas with argon the adjacent unscreened surface may be lowered by as much as $1{\cdot}3 \times 10^{-6}$ cm. These results indicate some complex interaction between the bombarding ions and the lattice structure.

When testing the specimens for their rectifying characteristics an important point is the contact

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* "The Properties of Ionic Bombarded Silicon", by Russell S. Ohl, January 1952, p. 104.

pressure. The best result is usually obtained with a light pressure but for good stability higher pressures are essential. The effect of pressure was found to be less with bombarded material than with unbombarded. A contact force of 10 grams was adopted as standard.

Fig. 1 shows the general form of the voltage-current characteristic. As an approximation one may assume that it is made up of three straight lines, CD and BA, the slopes of which give the resistances in the forward and backward directions, and the nearly horizontal BC corresponding to a very high resistance. The voltage $e = (E_B - E_F)/2$ is called the self-biasing voltage; it is the direct voltage across a large capacitance in series with the rectifier across an alternating supply voltage.

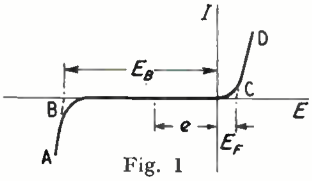

Fig. 1

Fig. 2 is typical of the results obtained. Fig. 2(a) shows the characteristic curve of the silicon before bombardment and Fig 2(b) after bombardment with helium ions under a voltage of 30 kV. The characteristics were measured under a contact force of 10 grams. The resistance corresponding to the portion BC of Fig. 1 is about 10 000 times as big in (b) as in (a), and for an applied p.d. of ± 1 V the ratio of the forward to the backward current is increased from about 20 to more than 10 000. The actual curves differ somewhat depending on the purity of the material, the gas employed, the voltage and temperature of bombardment and the contact pressure, but the differences are relatively small.

Experiments showed that similar results could be obtained by exposing the selenium surface to alpha-particle bombardment from a radio-active surface. A nickel surface was plated with a thin layer of polonium followed by a layer of gold. The selenium surface was placed against the layer of gold. This is, however, a very slow process, taking several days or even weeks to obtain the desired result.

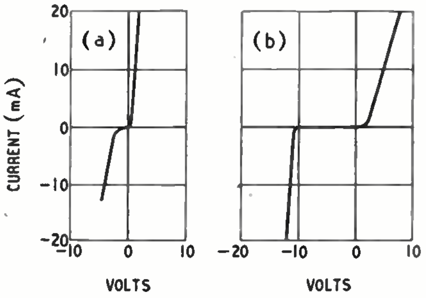

Fig. 2. (*a*) *Unbombarded silicon,* (*b*) *silicon bombarded with 30-kV helium ions.*

We have only given a general outline of this very interesting subject. A great many detailed curves are given in the original paper.

G. W. O. H.

VELOCITY-MODULATED DETECTOR

By F. N. H. Robinson, M.A.

(*Royal Naval Scientific Service**)

MOST microwave receivers employ a superheterodyne circuit using a crystal as a mixer, but in certain applications where its simplicity outweighs its lack of sensitivity, a crystal-video receiver may be used. In this type of receiver the r.f. signal is detected by a crystal immediately at the input from the aerial and the resulting signal is amplified by a high-gain video amplifier. Apart from not requiring a local oscillator or i.f. amplifier such a receiver has the added advantage of a greater r.f. bandwidth, this being limited only by the selectivity of the input circuit. On the other hand the sensitivity of crystal-video receivers is currently some 40 db less than that of a good superheterodyne. This, of course, is due to the fact that, in the absence of a local oscillator signal, the crystal operates as a square-law device at low signal levels.

Attempts in the past to make a velocity-modulated mixer to replace the crystal in superheterodyne receivers have met with little success, as this type of tube has been found to be unexpectedly noisy, see, for instance, reference (1). A suggestion, by R. Kompfner, that the source of this excess noise might be connected with the mechanism of mixing, led the author to investigate whether a similar type of tube used as a 'straight' square-law detector might not be more successful.

In the v.m. detector shown diagrammatically in Fig. 1, the r.f. signal is fed on to a helix through which passes an electron beam. If the velocity of the electrons in the beam is approximately equal to the velocity of wave propagation in the helix, the beam emerges from the helix velocity-modulated. That is to say, electrons in the beam

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* Now at the Clarendon Laboratory, Oxford.

MS accepted by the Editor, October 1951

have superimposed on their d.c. energy V_0 (expressed in electron volts) an a.c. component $V_1 \sin \omega t$. V_1 is related to the input r.f. signal power P_1 watts by

$$V_1^2 = P_1 Z_0 (2\pi N)^2 \quad .. \quad .. \quad .. \quad (1)$$

where Z_0 is the effective helix impedance and N is the length of the helix expressed in wavelengths at the beam velocity.[2] After leaving the helix the beam impinges on a collector at approximately cathode potential.

In the absence of any signal all the electrons will be collected when this electrode is at cathode potential, provided that they all approach the collector in trajectories normal to its surface. In the presence of a signal, however, some electrons will have lost energy to the r.f. field, (i.e., those electrons which entered the helix at a phase such that $V_1 \sin \omega t$ is negative) and will then have insufficient energy to reach the collector. A signal on the helix thus results in a mean reduction in collector current and will appear as a signal in any circuit connected to it.

In practice, two factors modify this ideal behaviour. First, the electrons, due to their origin at a thermionic cathode, possess a Maxwellian distribution of velocities, and hence some electrons may arrive at the collector even when it is slightly negative. Secondly, the electrons may not approach the collector in trajectories normal to its surface, so that they will only have sufficient energy to land when the potential of the collector is sufficiently positive to compensate for that part of their energy which is 'wasted' in components of motion tangential to the collector surface. As a result of both these factors the collector current, instead of changing abruptly as the collector potential is varied relative to the cathode, changes more gradually.

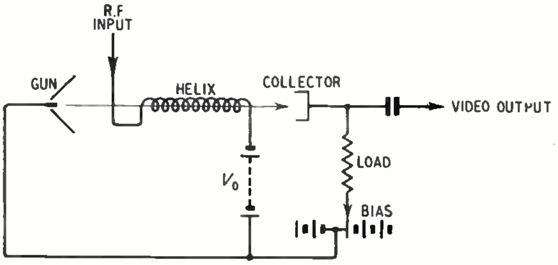


Fig. 1. *Diagram of v.m. tube as a signal detector.*

An actual collector characteristic curve is shown in Fig. 2. We may also interpret this curve as showing the dependence of collector current I on the energy lost or gained by electrons in the helix when a signal is present. That is, we may regard Fig. 2 as a curve of I against $V_1 \sin \omega t$.

In order to understand this, it is, perhaps, advantageous to think of the characteristic as being taken by keeping the collector potential fixed while varying the cathode potential with respect to it. It is then apparent that electrons emerging from the helix and approaching the collector may have had their kinetic energy changed either by a change in cathode potential or by the effect of the r.f. field, but in either case the effect on the collector current will be the same. We may therefore write the collector current as a function $I(V)$ of either the extra beam energy or the collector potential, and use Maclaurin's theorem to express the change in collector current due to a small change δV in the beam energy.

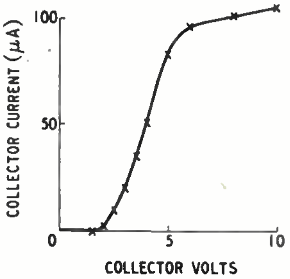


Fig. 2. *Static characteristic of v.m. tube at about 950 V; the total current is 120 μA.*

$$\delta I = I' \delta V + \tfrac{1}{2} I'' (\delta V)^2 + \text{etc.} \quad .. \quad (2)$$

where a dash denotes differentiation with respect to V. If we now identify δV with the r.f. modulation $V_1 \sin \omega t$ and average over one r.f. cycle we obtain

$$\delta I = \tfrac{1}{4} V_1^2 I'' \quad .. \quad .. \quad .. \quad (3)$$

and this is the rectified signal current.

If the input impedance of the succeeding video amplifier is R, this results in a video signal $V_s = R\delta I$ and using equations (1) and (3) we find

$$V_s = (\pi N)^2 Z_0 R I'' P_1 \quad .. \quad .. \quad .. \quad (4)$$

In order to calculate the sensitivity of this system we need an expression for the mean-square noise voltage V^2_N at the input to the video amplifier. We may express the noise originating within the amplifier in terms of the thermal noise of a fictitious equivalent noise resistance R_n; to this will be added the thermal noise in the load resistance R and the noise voltage developed across R by shot fluctuations in the beam current I reaching the collector. In addition, we should also consider any r.f. noise entering with the signal and any amplified shot noise generated within the helix. However, these last two sources of noise will, in nearly all practical cases, be quite negligible and we may write

$$V^2_n = (R + R_n) 4kTB_v + 2eIR^2B_v \quad .. \quad (5)$$

where k is in Boltzmann's constant, T is conventionally taken to be 288° Kelvin and B_v is the *video* bandwidth.

Equating V_s^2 and V_n^2 and using the approximate relation 1 electron volt ≈ 40 kT we find that the r.f. signal power which causes a video signal equal to noise is

$$P_m = \frac{\{4(R + R_n) + 80IR^2\}^{\frac{1}{2}}}{(\pi N)^2 R Z_0 I''} (kTB_v)^{\frac{1}{2}} \qquad (6)$$

We may use equation (6) to estimate the sensitivity we should obtain with a tube having an ideal collector whose characteristic is determined solely by the Maxwellian velocity distribution of the electrons. In this case the function $I(V)$ is given by

$$I = I_0 \exp\left(\frac{eV}{kT_c}\right) \text{ when } V < 0 \quad .. \qquad (7)$$

$$I = I_0 \text{ when } V \geqslant 0$$

$$\text{and } I'' = \left(\frac{e}{kT_c}\right)^2 I \quad .. \quad .. \quad .. \qquad (8)$$

If we take T_c, the cathode temperature, to be 1100°K we see that $I'' = 100\ I$. Substituting this result in equation (6) and taking $I = 10\,\mu\text{A}$, $N = 10$, $Z_0 = 200\ \Omega$ and $R_n = 1000\ \Omega$, $R = 10{,}000\ \Omega$ and $B_v = 1$ Mc/s which are all figures easily obtained in practice, we find $P_m = 10^{-11}$ watts which is some hundred times better than the sensitivity obtained using a video crystal.

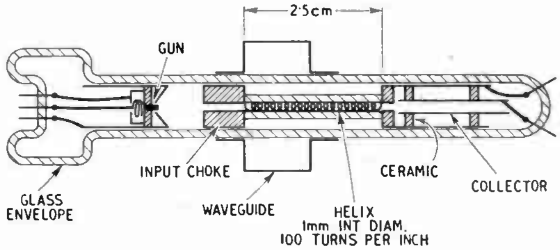


Fig. 3. *Form of v.m. tube used in the experiments at 3-cm wavelength.*

Experiments have been made at wavelengths of 10 and 3 cm. No particular attempt was made to design an efficient collector, and the simple structure shown in Fig. 3 was adopted. Nevertheless despite the fact that this collector is extremely inefficient, having a value of I'' some thousand times less than the ideal characteristic, sensitivities of 10^{-9} watts at 10 cm and 10^{-8} watts at 3 cm with a video bandwidth of 1 Mc/s were obtained, and at 10 cm using a higher value of load resistance R and a video bandwidth of 1 kc/s a sensitivity of 10^{-11} watt was achieved.

In Fig. 4 is shown a curve of video output for constant signal input as a function of collector bias. Reference to Fig. 2, which also refers to the tube shown in Fig. 3, shows that the peaks correspond to the points of maximum curvature of the characteristic where I'' is greatest. This particular tube (3 cm) had a helix of 100 turns per inch 2·5 cm long and 1-mm diameter supported in a quartz tube brazed at each end to metal collars, one of which served as the input choke. The tube operated at 950 volts and when the value of I'' taken from the measured characteristic is inserted in equation (6), together with the relevant values of the other parameters, the predicted sensitivity is in fair agreement with the measured value.

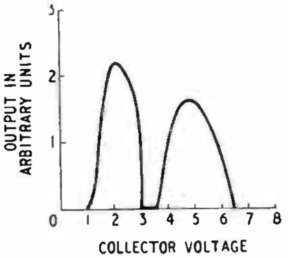


Fig. 4. *Detector video output for* 10^{-7} *watts r.f. input. Total beam current* 120 μA*; beam voltage* 950.

Clearly the sensitivity could be greatly improved by a better design of collector. It has been found possible in other contexts to design collectors having characteristics approaching within a factor 10 of the ideal and if this is attained with the v.m. detector, then sensitivities of 10^{-10} watts at 3 cm and 10^{-11} watts at 10 cm with a video bandwidth of 1 Mc/s would be available. This would compare very favourably with the sensitivity of video-crystal receivers.

Apart from the promise of greater sensitivity the v.m. detector has several other advantages over a crystal. Among these may be noted its freedom from burning out by large signals and its great r.f. bandwidth. This is determined principally by the selectivity of the input circuit and the match between it and the helix. A match over a range of half an octave (1500 Mc/s at 10 cm) is quite feasible.

Acknowledgment

I am indebted to the Chief of the Royal Naval Scientific Service for permission to publish this work.

REFERENCES

[1] A. H. W. Beck (1947), "Velocity Modulated Thermionic Tubes, Cambridge University Press.

[2] R. Kompfner, "The Travelling-Wave Tube," *Wireless Engineer*, 1947, Vol. 24, p. 255.

B.B.C. APPOINTMENTS

Consequent upon the retirement of Sir Noel Ashbridge, the Director of Technical Services is to be H. Bishop, C.B.E., B.Sc. (Eng.), F.C.G.I., M.I.E.E., M.I.Mech.E., while R. T. B. Wynn, C.B.E., M.A., M.I.E.E., will become Chief Engineer. The Deputy Chief Engineer will be F. C. McLean, M.B.E., B.Sc., M.I.E.E. The appointments are effective from 1st August 1952.

Sir Noel retires at the age of 62, having been with the B.B.C. since 1926. He has been Director of Technical Services since 1947.

NEGATIVE-FEEDBACK AMPLIFIERS

Overloading under Pulse Conditions

By J. E. Flood, Ph.D., A.M.I.E.E.

SUMMARY.—The overloading of simple resistance-coupled negative-feedback amplifiers is investigated for an input signal which rises from zero to its final value at a uniform rate. The amplifiers are assumed to be linear unless the applied signal exceeds the permitted value. For the single-stage amplifier, the permissible input voltage decreases as the rise-time of the signal is reduced. Two-stage and three-stage critically-damped amplifiers can be made to handle quickly-changing signals which are as large as the maximum permissible slowly-changing signal provided that the time-constant of the first stage of the amplifier is sufficiently large compared with that of the second stage. Overloading of the amplifier by step voltages and by sinusoidal signals is studied in appendices.

1. Introduction

NEGATIVE-FEEDBACK amplifiers often distort a signal, such as a pulse, which changes rapidly with time, although the amplitude of the signal is less than that required to overload the amplifier when the rate of change of the signal is small.[1] This is because the feedback voltage changes more slowly than the input voltage, with the result that the voltage applied to the grid of one of the valves becomes large enough to drive it into grid current or beyond cut-off.

The inability to handle quickly-changing input voltages has often been noticed in the case of a cathode follower with a capacitive load, especially for negative-going voltages which cut the valve off so that the cathode follows the input voltage with a time-constant equal to the load capacitance multiplied by the load resistance, instead of with the much smaller time-constant equal to the load capacitance multiplied by the output resistance of the stage. A. J. Shimmins[2] has studied the overloading of the cathode follower by a unit-step input voltage (i.e., a voltage which is zero for $t<0$ and unity for $t>0$). B. Y. Mills[3] has considered an input signal of the form shown in Fig. 1 and has determined how the permissible signal magnitude decreases as the rise-time (T) is reduced. R. H. Baer[4] has published a nomogram giving the ratio of the input signal (V_i) permissible for a quick change (of the form shown in Fig. 1) to that permissible for a slow change ($T \rightarrow \infty$).

The object of the present paper is to investigate the overloading of simple two-stage and three-stage negative-feedback amplifiers by a signal of the form shown in Fig. 1. The maximum permissible input signal is that which causes the voltage applied to the grid of any valve to equal the maximum permitted for the valve and it is assumed that the amplifier is linear unless the signal exceeds the permitted value. The amplifiers considered have simple resistance-coupled stages of the form shown in Fig. 2; R is the anode load resistance and C is the total anode-earth stray capacitance.

MS accepted by the Editor, September 1951.

2. Single-Stage Amplifier

If the Heaviside operational expression for the indicial response (i.e., the response to a unit step) of the stage is $\mu(p)\ 1$ and the voltage applied to the grid of the valve is $v_g(t)\ 1$, the output voltage is

$$v_a(t)\ 1 = \mu(p).\ v_g(t)\ 1$$

If the transfer coefficient, β, of the feedback path is independent of frequency the voltage returned to the input circuit to produce feedback is

$$\beta v_a(t)\ 1 = \mu(p).\beta\ v_g(t)\ 1$$

If the input voltage to the feedback amplifier is $v_i(t)\ 1$ then

$$v_g(t)\ 1 = v_i(t)\ 1 - \mu(p).\beta v_g(t)\ 1$$

$$\therefore \quad v_g(t)\ 1 = \frac{1}{1+\mu(p).\beta} \cdot v_i(t)\ 1$$

$$v_a(t)\ 1 = \frac{\mu(p)}{1+\mu(p).\beta} \cdot v_i(t)\ 1$$

The sign convention is chosen so that μ and β are both positive when the feedback is negative. For a stage of the form shown in Fig. 2,

$$\mu(p) = \mu_1 \frac{\alpha_1}{p+\alpha_1}$$

where $\mu_1 = g_m R$ and $\alpha_1 = 1/RC$

$$\therefore v_g(t)\ 1 = \frac{1}{1+\mu_1\beta\dfrac{\alpha_1}{p+\alpha_1}} \ .\ v_i(t)\ 1$$

$$= \frac{p+\alpha_1}{p+\alpha} \ .\ v_i(t)\ 1 \quad \ldots \quad \ldots \quad (1)$$

where $\alpha = \alpha_1(1+\mu_1\beta)$

For the input voltage shown in Fig. 1:

$$v_i = kt \text{ (for } 0 \leqslant t \leqslant T) \quad \ldots \quad \ldots \quad \ldots \quad (2a)$$

$$= \frac{k}{p}\ 1 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2b)$$

where

$$k = V_1/T \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2c)$$

$$\therefore\ v_g = \frac{k}{p}\cdot\frac{p+\alpha_1}{p+\alpha}\,1 \text{ [from equations (1) and (2b)]}$$

$$= \frac{k}{\alpha}\left\{\frac{\alpha_1}{p} + \frac{\alpha}{p+\alpha}\left[1-\frac{\alpha_1}{\alpha}\right]\right\}1$$

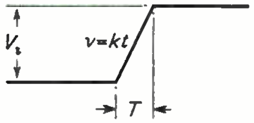
$$= \frac{k}{\alpha}\left\{\alpha_1 t + \left[1-\frac{\alpha_1}{\alpha}\right]\left[1-e^{-\alpha t}\right]\right\} \quad \text{(for } 0 \leqslant t \leqslant T\text{)}$$

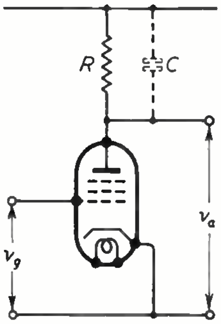
But $\alpha = \alpha_1(1+\mu_1\beta)$

$$\therefore\ v_g = \frac{kt}{1+\mu_1\beta}\left\{1 + \frac{\mu_1\beta}{\alpha t}\left[1-e^{-\alpha t}\right]\right\}$$

Fig. 1 (below). *Input signal with uniform slope.*

Fig. 2 (right). *Simple resistance-coupled stage.*





The grid voltage has its largest value, $V_g(T)$, when $t = T$,

$$\therefore\ V_g(T) = \frac{V_i}{1+\mu_1\beta}\left\{1 + \frac{\mu_1\beta}{\alpha T}\left[1-e^{-\alpha T}\right]\right\} \quad .. \quad (3)$$

If the signal is to be negligibly distorted it is required that $e^{-\alpha T} \ll 1$, so we may write

$$V_g(T) = \frac{V_i}{1+\mu_1\beta}\left\{1 + \frac{A}{\alpha T}\right\} \quad .. \quad .. \quad (4)$$

where $A = \mu_1\beta$.

The largest voltage applied to the grid of the valve when the transition is very slow $(T \rightarrow \infty)$ is

$$V_g(\infty) = \frac{V_i}{1+\mu_1\beta}$$

$$\therefore\ \frac{V_g(T)}{V_g(\infty)} = 1 + \frac{A}{\alpha T} \quad .. \quad .. \quad .. \quad (5)$$

If this ratio is plotted again αT for various values of A a family of rectangular hyperbolae is obtained, as shown in Fig. 3. Equation (5) is inaccurate for small values of αT because the exponential term is neglected; according to this equation $V_g(T)$ tends to infinity when $T \rightarrow 0$, but, from equation (3), $\lim\limits_{T\rightarrow 0} V_g(T) = V_i$. This is because, when the input signal is a step voltage, the whole of it is applied to the grid of the valve before the feedback voltage can reach any appreciable magnitude.

$$\therefore\ V_g(0)/V_g(\infty) = 1 + \mu_1\beta.$$

An approximate curve for small values of αT can therefore be obtained by drawing on Fig. 3 the tangent from the point $(0,\ 1+\mu_1\beta)$ to the curve for the appropriate value of A.

For a given maximum permissible grid swing, the ratio of the permissible input voltages for rapid and slow transitions is $[1 + A/\alpha T]^{-1}$. This expression is of the same form as that obtained by Mills for the cathode follower.[3]

3. Two-Stage Amplifier

Fig. 4 shows a typical two-stage negative-feedback amplifier. If $Z_f \gg R_2$, the indicial response of the amplifying path is

$$\mu(p)\,1 = \mu_0\,\frac{\alpha_1}{p+\alpha_1}\cdot\frac{\alpha_2}{p+\alpha_2}\,1$$

where $\alpha_1 = 1/R_1C_1$, $\alpha_2 = 1/R_2C_2$, $\mu_0 = \mu_1\mu_2$ and $\mu_1 = g_mR_1$, $\mu_2 = g_mR_2$.

The analysis is simplified if the feedback amplifier is critically damped. Critical damping can be obtained by suitable adjustment of the time constants of the two stages and of the feedback path, as described in a previous paper.[5] In practice, an indicial response which has an overshoot may sometimes be preferred, but the results obtained here will still hold approximately provided the overshoot is small.

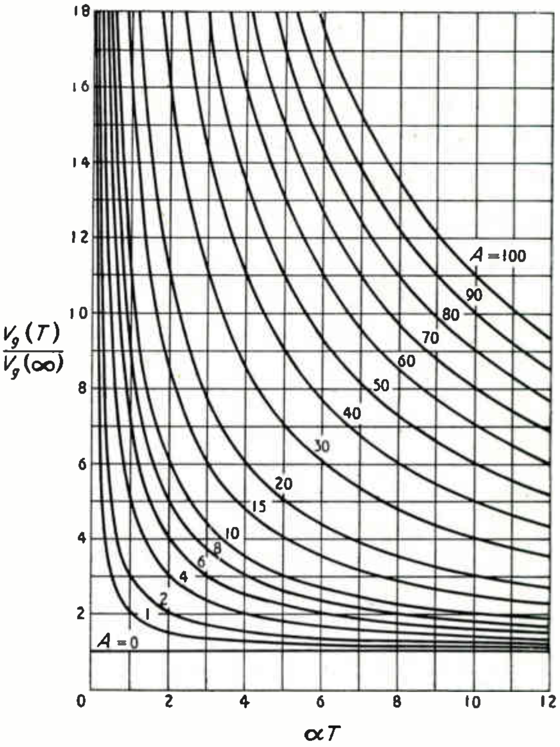


Fig. 3. *Maximum voltage applied to valve when signal has rise-time T.*

It can be shown[5] that the indicial response of the critically-damped amplifier is

$$h(p)\,1 = \frac{\mu_o}{1+\mu_o\beta_o}\cdot\frac{\alpha^2}{(p+\alpha)^2}\,1 \quad .. \quad (6a)$$

$$\therefore\ h(t) = \frac{\mu_0}{1+\mu_0\beta_0}\left[1-e^{-\alpha t}\,(1+\alpha t)\right]$$

$$(\text{for } t \geqslant 0) \quad .. \quad (6b)$$

$$\text{where} \quad \alpha^2 = \alpha_1\alpha_2(1+\mu_0\beta_0) \quad .. \quad .. \quad (7a)$$

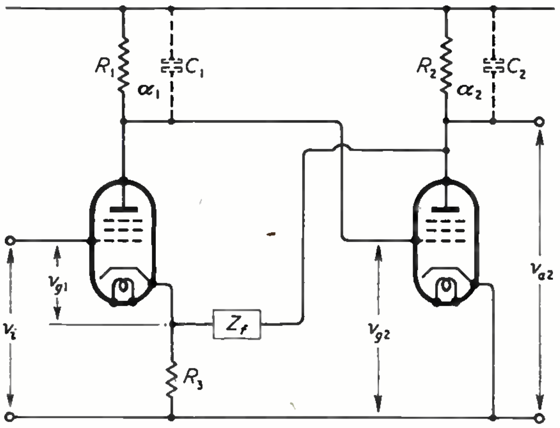


Fig. 4. *Typical two-stage amplifier.*

and it is required[5] that the feedback factor, $1+\mu_0\beta_0$, satisfies the condition

$$1+\mu_0\beta_0 \geqslant \frac{(\alpha_1+\alpha_2)^2}{4\alpha_1\alpha_2} \quad .. \quad .. \quad .. \quad (7b)$$

$$\therefore \quad \alpha \geqslant \tfrac{1}{2}(\alpha_1+\alpha_2) \quad .. \quad .. \quad .. \quad (7c)$$

From equation (6a), the output voltage, v_{a2}, is

$$v_{a_2} = \frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{\alpha^2}{(p+\alpha)^2}\cdot v_i\,1 \quad .. \quad (8)$$

For the input voltage, shown in Fig. 1, equations (8) and (2b) give

$$v_{a_2} = \frac{\mu_0}{1+\mu_0\beta_0}\cdot\frac{\alpha^2}{(p+\alpha)^2}\cdot\frac{k}{p}\,1 \quad .. \quad .. \quad (9a)$$

$$= \frac{\mu_0 k}{1+\mu_0\beta_0}\left\{\frac{1}{p}-\frac{2}{p+\alpha}+\frac{p}{(p+\alpha)^2}\right\}1$$

$$= \frac{\mu_0 k}{1+\mu_0\beta_0}\left\{t-\frac{2}{\alpha}\left(1-e^{-\alpha t}\right)+te^{-\alpha t}\right\}1$$

$$= \frac{\mu_0 kt}{1+\mu_0\beta_0}\left\{(1+e^{-\alpha t})-\frac{2}{\alpha t}(1-e^{-\alpha t})\right\}$$

$$(\text{for } 0 \leqslant t \leqslant T) \quad .. \quad .. \quad (9b)$$

$$\text{Now} \quad v_{a_2} = \mu_2\frac{\alpha_2}{p+\alpha_2}\cdot v_{g_2}\,1$$

$$\therefore \quad v_{g_2} = \frac{p+\alpha_2}{\mu_2\alpha_2}\cdot v_{a_2}\,1$$

and substituting from equation (8) for v_{a_2}:

$$v_{g_2} = \frac{\mu_1}{1+\mu_0\beta_0}\cdot\frac{\alpha^2(p+\alpha_2)}{\alpha_2(p+\alpha)^2}\cdot v_i\,1 \quad .. \quad (10)$$

$$(\text{since } \mu_0 = \mu_1\mu_2)$$

For the input voltage shown in Fig. 1, equations (10) and (2b) give

$$v_{g_2} = \frac{\mu_1}{1+\mu_0\beta_0}\cdot\frac{k}{p}\cdot\frac{p+\alpha_2}{\alpha_2}\cdot\frac{\alpha^2}{(p+\alpha)^2}\,1$$

$$.. \quad .. \quad .. \quad .. \quad .. \quad (11a)$$

$$= \frac{\mu_1}{1+\mu_0\beta_0}\cdot\frac{k}{\alpha_2}\left\{\frac{\alpha_2}{p}+\frac{\alpha-2\alpha_2}{p+\alpha}-\frac{(\alpha-\alpha_2)p}{(p+\alpha)^2}\right\}1$$

$$= \frac{\mu_1}{1+\mu_0\beta_0}\left\{kt+k\left(\frac{\alpha-2\alpha_2}{\alpha\,\alpha_2}\right)\cdot\left[1-e^{-\alpha t}\right]-kt\left(\frac{\alpha-\alpha_2}{\alpha_2}\right)e^{-\alpha t}\right\}1$$

$$= \frac{\mu_1}{1+\mu_0\beta_0}kt\left\{\left[1-\frac{\alpha-\alpha_2}{\alpha_2}e^{-\alpha t}\right]+\frac{\alpha-2\alpha_2}{\alpha\alpha_2 t}\left[1-e^{-\alpha t}\right]\right\} \quad .. \quad (11b)$$

$$(\text{for } 0 \leqslant t \leqslant T)$$

If $e^{-\alpha t} \ll 1$, the largest value of v_{g2} occurs at time T and is

$$V_{g2}(T) = V_i\frac{\mu_1}{1+\mu_0\beta_0}\left[1+\frac{\alpha-2\alpha_2}{\alpha\,\alpha_2 T}\right] \quad .. \quad (12)$$

For a very slow transition ($T \to \infty$):

$$V_{g2}(\infty) = V_i\frac{\mu_1}{1+\mu_0\beta_0}$$

$$\therefore \frac{V_{g2}(T)}{V_{g2}(\infty)} = 1+\frac{\alpha-2\alpha_2}{\alpha\alpha_2 T} \quad .. \quad .. \quad (13)$$

$$\therefore \frac{V_{g2}(T)}{V_{g2}(\infty)} = 1+\frac{A}{\alpha T}$$

$$\text{where} \quad A = \frac{\alpha}{\alpha_2}-2$$

$$= [(1+\mu_0\beta_0)\alpha_1/\alpha_2]^{\frac{1}{2}}-2$$

(from equation 7a)

The voltage applied to the grid of the second valve during a rapid change does not exceed that for a slow change if $A \leqslant 0$

$$\text{i.e., if } \alpha \leqslant 2\alpha_2 \quad .. \quad .. \quad .. \quad .. \quad (14a)$$

$$\text{whence } 1+\mu_0\beta_0 \leqslant 4\alpha_2/\alpha_1 \quad .. \quad .. \quad (14b)$$

When $A > 0$, however, the voltage applied to the grid of the second valve is greater for a rapid than for a slow change. The maximum permissible input signal can be determined from Fig. 3, using the value of A given above. Fig. 5 shows the variation of the parameter A with αT for several feedback factors. Equation (13) is inaccurate for small values of αT because the exponential terms are neglected. According to equation (13), V_{g2} becomes infinite when the input is a step voltage ($T \rightarrow 0$), whereas it is shown in Appendix 2 that

$$\frac{V_{g2}(0)}{V_{g2}(\infty)} = 1 + \left(\frac{\alpha}{\alpha_2} - 1\right) e^{-\left(\frac{\alpha}{\alpha - \alpha_2}\right)} \quad .. \quad (15)$$

Fig. 6 shows the variation of this ratio with α_1/α_2 for several values of the feedback factor. An approximate curve for $V_{g2}(T)$ for small values of αT can, therefore, be obtained by drawing on Fig. 3 the tangent from the point $[0, V_{g2}(0)/V_{g2}(\infty)]$ to the curve for the appropriate value of A.

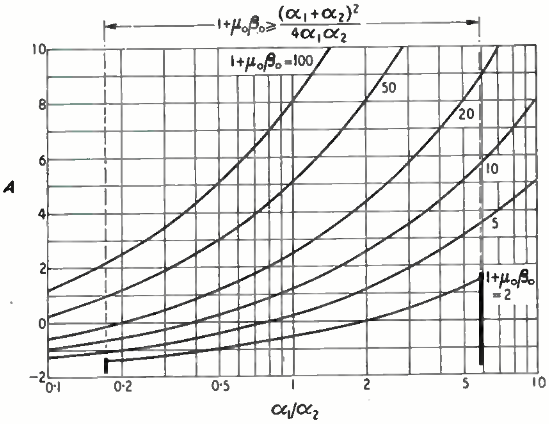


Fig. 5. *Overloading parameter for second stage.*

In order to ensure that the overload voltage of the amplifier does not decrease when the rise-time of the input signal is reduced, the time-constant of the second stage must usually be small compared with that of the first stage [inequality (14b)]. However, in order to obtain the largest possible output voltage, for a given maximum permissible grid voltage, with slowly-changing signals, the anode load resistance of the second stage is commonly made as large as possible. This results in the time-constant of the second stage being too large to satisfy inequality (14) with consequent reduction of the permissible input signal when the rise-time is reduced. The second stage has the largest anode load resistance, for which there is no reduction of permissible input for quickly-changing signals when $\alpha_2 = \frac{1}{2}\alpha$. If we now multiply the anode load resistance by m, while keeping constant the stray capacitances and the overall gain and time-constant, the resultant time-constant of the second stage ($1/\alpha'_2$) is then given by $\alpha'_2 = \alpha/2m$ and the permissible output voltage for slowly changing signals is multiplied by m. If the applied signal has rise-time T, the permissible output voltage is divided by $1 + A/\alpha T$ where

$$A = \frac{\alpha}{\alpha'_2} - 2 = 2(m - 1)$$

If the permissible output voltage is to be greater than that when $\alpha_2 = \frac{1}{2}\alpha$ we require

$$\frac{m}{1 + A/\alpha T} > 1$$

$$\therefore \quad m > 1 + 2(m - 1)/\alpha T$$

and $\quad \alpha T > 2$

This result is not very accurate because the exponential terms have been neglected, but, if the rise-times of the input signals will always be substantially greater than twice the amplifier time-constant, the overload signal voltage can be increased by making the anode load resistance of the second stage greater than that required to satisfy inequality (14). The largest permissible input signal can then be determined from the permissible slowly changing signal by means of Fig. 3. If, however, the rise-times of the signals are likely to be of the order of twice the time-constant, or less, the largest overload signal is obtained when $\alpha_2 = \frac{1}{2}\alpha$ so that the overload signal does not decrease as the rise-time is reduced.

If the time-constant of the first stage is made sufficiently large compared with the time-constant of the second stage ($4\alpha_2/\alpha_1 \geqslant 1 + \mu_0\beta_0$), the voltage received by the grid of the second stage when a rapid change is applied to the amplifier does not exceed that received when the change is slow. The permissible input voltage of the amplifier is therefore not reduced when the change is rapid unless the voltage applied to the grid of the first stage can exceed that permissible.

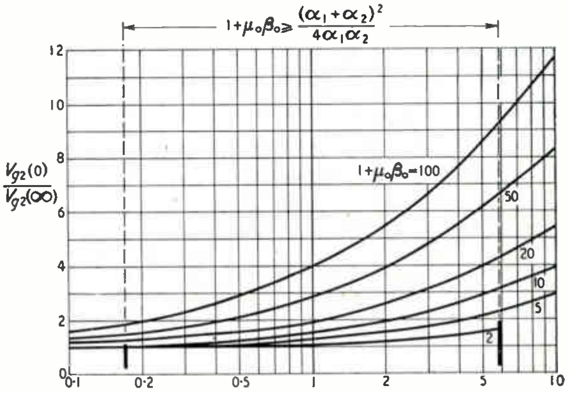


Fig. 6. *Maximum voltage applied to second stage when input signal is a step voltage.*

$$\text{Now} \quad v_{g2} = \mu_1 \frac{\alpha_1}{p + \alpha_1} \, . \, v_{g1} \, 1$$

$$\therefore \quad v_{g1} = \frac{p + \alpha_1}{\mu_1 \alpha_1} \, . \, v_{g2} \, 1$$

Substituting from equation (10) for v_{g2}:

$$v_{g1} = \frac{1}{1+\mu_o\beta_o} \cdot \frac{p+\alpha_1}{\alpha_1} \cdot \frac{p+\alpha_2}{\alpha_2} \cdot \frac{\alpha^2}{(p+\alpha)^2} \cdot v_i\, 1 \qquad \ldots \quad (16)$$

For the input voltage shown in Fig. 1, equations (16) and (2b) give

$$v_{g1} = \frac{1}{1+\mu_0\beta_0} \cdot \frac{k}{p} \cdot \frac{p+\alpha_1}{\alpha_1} \cdot \frac{p+\alpha_2}{\alpha_2} \cdot \frac{\alpha^2}{(p+\alpha)^2}\, 1 \qquad \ldots\ \ldots\ \ldots\ \ldots\ \ldots\ (17a)$$

$$\therefore\ v_{g1} = \frac{1}{1+\mu_o\beta_o} \cdot \frac{k}{\alpha_1\alpha_2}\left\{\frac{\alpha_1\alpha_2}{p} + \frac{\alpha(\alpha_1+\alpha_2) - 2\alpha_1\alpha_2}{p+\alpha} + \frac{[\alpha^2+\alpha_1\alpha_2 - \alpha(\alpha_1+\alpha_2)]p}{(p+\alpha)^2}\right\} 1$$

$$= \frac{k}{1+\mu_0\beta_0}\left\{t + \left[\frac{1}{\alpha_1}+\frac{1}{\alpha_2}-\frac{2}{\alpha}\right]\left[1-e^{-\alpha t}\right] + \left[1+\frac{\alpha^2}{\alpha_1\alpha_2}-\frac{\alpha}{\alpha_1}-\frac{\alpha}{\alpha_2}\right]te^{-\alpha t}\right\}$$

$$(\text{for } 0 \leqslant t \leqslant T) \quad \ldots\ \ldots\ (17b)$$

If $e^{-\alpha T} \ll 1$, the largest value of v_{g1} occurs at time T and is

$$V_{g1}(T) = \frac{V_i}{1+\mu_0\beta_0}\left\{1+\frac{1}{\alpha T}\left[\frac{\alpha}{\alpha_1}+\frac{\alpha}{\alpha_2}-2\right]\right\} \qquad \ldots\ \ldots\ \ldots\ \ldots\ \ldots\ (18)$$

$$\therefore \frac{V_{g1}(T)}{V_{g1}(\infty)} = 1 + \frac{A}{\alpha T}$$

where $A = \frac{\alpha}{\alpha_1} + \frac{\alpha}{\alpha_2} - 2$

$$= \left[\frac{1+\mu_0\beta_0}{\alpha_1\alpha_2}\right]^{\frac{1}{2}}(\alpha_1+\alpha_2) - 2$$

[from equation (7a)]

$V_{g1}(T)$ is greater than $V_{g1}(\infty)$ if $A > 0$.
But $4\alpha_1\alpha_2/(\alpha_1+\alpha_2)^2 \leqslant 1$ and $\mu_0\beta_0 > 0$.

$$\therefore\ 1 + \mu_0\beta_0 > \frac{4\alpha_1\alpha_2}{(\alpha_1+\alpha_2)^2}$$

so that $A > 0$.

The voltage applied to the grid of the first valve is therefore always greater for a rapid change than for a slow change. The voltage applied to the grid of the first valve can be determined from Fig. 3 using the value of A given above. The results are inaccurate for small values of αT because the exponential terms are neglected in equation (18). When the input signal is a step voltage ($T \rightarrow 0$) the whole of the input voltage is immediately applied to the first valve, as shown in Appendix 2.

$$\therefore \frac{V_{g1}(0)}{V_{g2}(\infty)} = 1 + \mu_0\beta_0$$

An approximate curve for $V_{g1}(T)$ for small values of αT can therefore be obtained by drawing on Fig. 3, the tangent from the point $(0, 1+\mu_0\beta_0)$ to the curve for the appropriate value of A.

It is, however, impossible to overload the first valve, even with a step voltage input, without also overloading the second valve unless the overall gain of the amplifier is very low. Consider, as an example, an amplifier whose two valves are identical and have equal stray capacitances across their anode loads, then

$$\frac{\alpha_1}{\alpha_2} = \frac{\mu_2}{\mu_1} \qquad \ldots\ \ldots\ \ldots\ \ldots\ (19)$$

The maximum voltage applied to the grid of the first valve, when the input is a step voltage is equal to the input voltage. The second valve can be made not to receive a voltage greater than the steady-state value by complying with inequality (14a). Therefore, if an applied voltage V_i overloads the first valve without overloading the second:

$$V_i > V_{g2}(\infty) = \frac{\mu_1}{1+\mu_0\beta_0} V_i$$

$$\therefore \mu_1 < 1 + \mu_0\beta_0$$

and $\mu_2 < \frac{\alpha_2}{\alpha_1}(1+\mu_0\beta_0)$ [from equation (19)]

$$\therefore \mu_0 = \mu_1\mu_2 < \frac{\alpha_2}{\alpha_1}(1+\mu_0\beta_0)^2$$

But $1 + \mu_0\beta_0 \leqslant 4\alpha_2/\alpha_1$ [inequality (14b)]

$$\therefore \frac{\mu_0}{1+\mu_0\beta_0} < 4$$

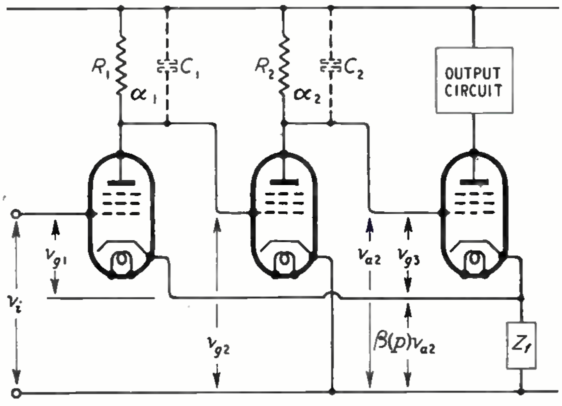


Fig. 7. *Typical three-stage amplifier.*

Thus, if the overall gain of the amplifier is greater than 4, any signal which overloads the first stage of the amplifier will also overload the second. Most practical amplifier circuits have a much larger gain than this, so by ensuring that the conditions are satisfied which prevent the second stage from being overloaded by a smaller input when the signal changes rapidly than when it changes slowly, it is possible to ensure that the magnitude of the input signal required to overload the amplifier does not decrease when the rise-time of the signal is reduced.

When the time-constant of the amplifier is sufficiently small for the output signal to be negligibly distorted ($e^{-\alpha T} \ll 1$) the second stage does not receive a voltage larger than that for a slow change provided that $\alpha \leqslant 2\alpha_2$ [inequality (14a)]. It is shown in Appendix 2 that, when the input signal is a step voltage, the second valve receives a voltage larger than that for a slow change unless $\alpha \leqslant \alpha_2$. However, when $\alpha_2 < \alpha < 2\alpha_2$, the reduction in the permissible input voltage is small; when $\alpha = 2\alpha_2$, for example, the permissible step voltage input is 12% smaller than the maximum permissible slowly changing voltage. It is shown in Appendix 3 that, when sinusoidal signals are applied to the amplifier, the permissible input voltage is smaller over some range of frequencies than it is at low frequencies unless $\alpha \leqslant \sqrt{2}\alpha_2$; when $\alpha = 2\alpha_2$, however, the maximum reduction in the permissible input voltage is only 13%.

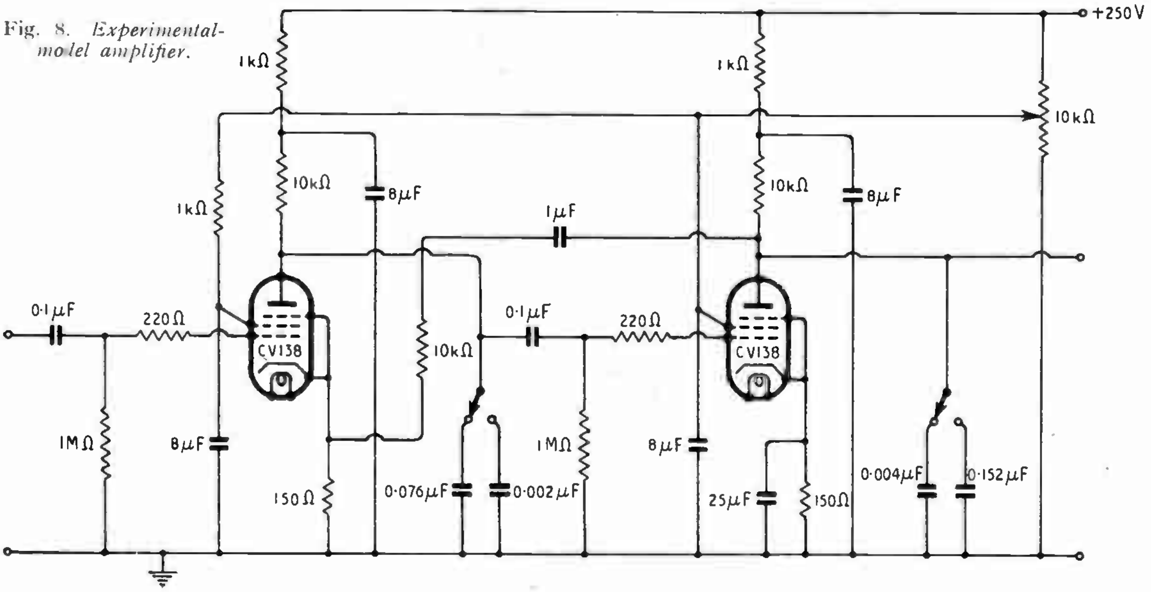


Fig. 8. *Experimental-model amplifier.*

4. Three-Stage Amplifier

Fig. 7 shows a typical three-stage negative-feedback amplifier. It will again be assumed that the amplifier is critically damped.

Now, $v_{g3}\,\mathbf{1} = (1 - \beta(p)).\,v_{a2}\,\mathbf{1}$ .. .. (20)

But $v_i\,\mathbf{1} = v_{g1}\,\mathbf{1} + \beta(p)\,.\,v_{a2}\,\mathbf{1}$

$\therefore \quad v_{g3} = v_{a2} + v_{g1} - v_i$ .. .. (21)

Now, for the input voltage shown in Fig. 1, v_i is given by equation (2b), v_{a2} is given by equation (9b) and v_{g1} is given by equation (17b).

∴ Substituting in equation (21) and neglecting the exponential terms:

$$v_{g3} = \frac{kt}{1+\mu_0\beta_0}\left\{\mu_0(1-\beta_0) + \frac{1}{\alpha t}\left[\frac{\alpha(\alpha_1+\alpha_2)}{\alpha_1\alpha_2} - 2(1+\mu_0)\right]\right\}$$

$$(\text{for } 0 \leqslant t \leqslant T)$$

v_{g3} does not exceed that for a slow change if

$$(1+\mu_0) \geqslant \frac{\alpha(\alpha_1+\alpha_2)}{2\alpha_1\alpha_2}$$

But $\dfrac{\alpha(\alpha_1+\alpha_2)}{2\alpha_1\alpha_2} \leqslant \dfrac{\alpha^2}{\alpha_1\alpha_2} = 1+\mu_0\beta_0$ [from (7c) and (7a)]

and $1 + \mu_0\beta_0 \leqslant 1 + \mu_0$ (since $\beta_0 \leqslant 1$)

$$\therefore (1+\mu_0) \geqslant \frac{\alpha(\alpha_1+\alpha_2)}{2\alpha_1\alpha_2}$$

Thus the voltage applied to the grid of the third stage does not exceed that for a slow change. The amplifier can, therefore, handle as large an input with a small rise-time as with a large unless the voltage applied to the grid of either the second or first valve increases sufficiently to overload it. It is shown in Section 3 that the first valve cannot be overloaded without also overloading the second, unless the overall gain is very low. Moreover, the second valve cannot be overloaded without also overloading the third, unless the gain of the second stage is very low.

Consider, as an example, an amplifier whose second and third valves are identical. The maximum voltage applied to the grid of the second valve occurs when the input is a step voltage ($T \rightarrow 0$) and, from equation (15), it is

$$V_{g2}(0) = \frac{\mu_1 V_i}{1 + \mu_0 \beta_0} \left\{ 1 + \left(\frac{\alpha}{\alpha^2} - 1 \right) e^{\left(\frac{\alpha}{\alpha - \alpha_2} \right)} \right\}$$

$$< \frac{\mu_1 V_i}{1 + \mu_0 \beta_0} \cdot \frac{\alpha}{\alpha_2}$$

The maximum voltage applied to the grid of the third valve is the steady-state voltage,

$$V_{g3}(\infty) = \frac{\mu_0 V_i}{1 + \mu_0 \beta_0} \cdot (1 - \beta_0)$$

The maximum voltage applied to the second valve is therefore less than that applied to the third valve when

$$\mu_1 \frac{\alpha}{\alpha_2} < \mu_0 \, (1 - \beta_0)$$

$$\therefore \mu_2 > \alpha/\alpha_2(1 - \beta_0) \qquad \text{(since } \mu_0 = \mu_1 \mu_2\text{)}$$

and $1 + \mu_0 \beta_0 < \mu_2{}^2(1 - \beta_0)^2 \alpha_2/\alpha_1$.

The ratio between the time-constants of the first two stages can therefore be reduced by the factor $4/\mu_2{}^2(1 - \beta_0)^2$ compared with the ratio required to satisfy inequality (14b).

5. Experimental Results

The amplifier shown in Fig. 8 was built with capacitors connected between the valve anodes and earth, so that the valve capacitances and wiring stray capacitances could be neglected and tests made using relatively long pulses. The screen voltage of the valves was adjusted so that the voltage gain without feedback was 56 db at low frequencies. The low-frequency gain with feedback was then 36 db. The anode-earth capacitances had the values calculated to give a critically-damped indicial response ($\alpha_1 = 38\alpha_2$ or $\alpha_2 = 38\alpha_1$). The amplifier had the same frequency response and time response, for small signals, either when $\alpha_1 = 38\alpha_2$ ($C_2 = 0{\cdot}002\,\mu$F, $C_2 = 0{\cdot}152\,\mu$F) or when $\alpha_2 = 38\alpha_1$ ($C_1 = 0{\cdot}076\,\mu$F, $C_2 = 0{\cdot}004\,\mu$F).

Positive-going rectangular pulses of 1-msec duration were applied to the amplifier and oscillograms were taken of the input and output voltages both when $\alpha_1 = 38\alpha_2$ and when $\alpha_2 = 38\alpha_1$. Fig. 9 shows some of the results obtained.

When the first stage had a short time-constant and the second stage had a long time-constant ($\alpha_1 = 38\alpha_2$), the rise of the output signal became slower when the input pulse exceeded about 0·1 V and further increases in input signal made the rise still slower. When the input pulse exceeded about 0·2 V, the output voltage substantially failed to reach the steady-state value during the pulse and further increases in the input pulse produced little change in the size or shape of the output pulse. This was because the negative-going pulse received by the grid of the second valve was of sufficient size to cut the valve off so that its anode potential rose with time-constant $1/\alpha_2$ instead of with the much smaller time-constant $1/\alpha$.

When the first stage had a long time-constant and the second stage had a short time-constant ($\alpha_2 = 38\alpha_1$) no deterioration of the rise-time was noticed. The output voltage increased linearly with the input voltage until the latter exceeded 1 V, as shown in Fig. 10.

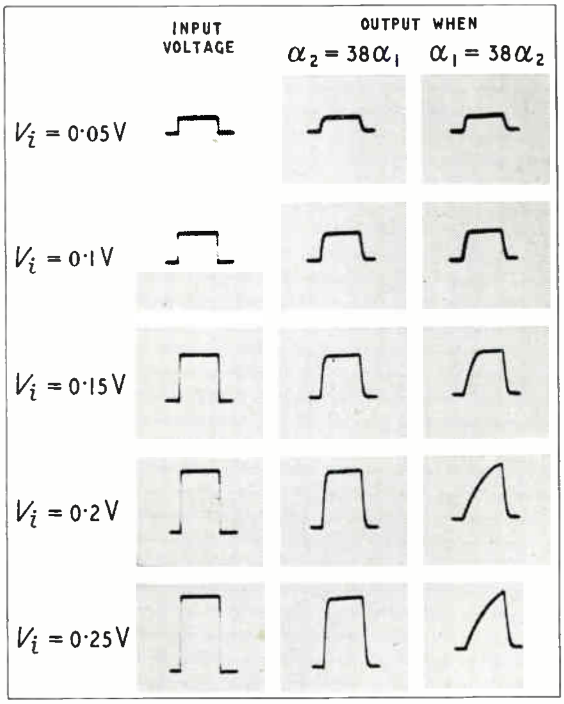


Fig. 9. *Response of model amplifier to 1-msec pulse.*

6. Conclusions

When an input signal of the form shown in Fig. 1 is applied to the single-stage negative-feedback amplifier the magnitude of signal required to overload the valve decreases as the rise-time of the signal is reduced. For the two-stage amplifier, the voltage applied to the first valve increases as the rise-time of the signal is reduced, but only if the gain of the amplifier is very small is it possible for the first valve to be overloaded by a signal which does not also overload the second valve. The second valve is, therefore, normally the first to overload. If the time-constant of the first stage is sufficiently large compared with that of the second stage, the input signal required to overload the second valve does not decrease as the rise-time of the signal is reduced. For the three-stage amplifier, the voltage applied to the third stage is never greater for a quick change than for a slow. Therefore, if the first two stages are designed not to overload, the signal required to overload the amplifier is as large when it changes quickly as when it changes slowly.

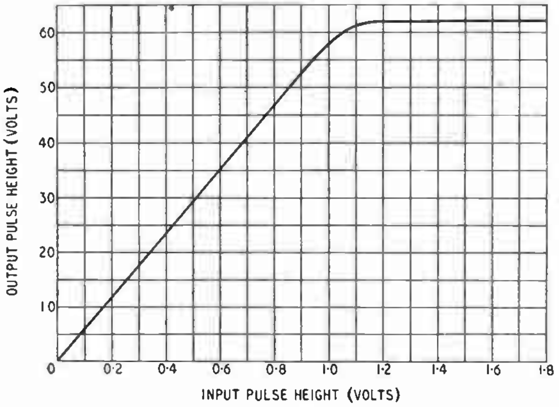


Fig. 10. *Response to 1 msec pulse of model amplifier* ($\alpha_2 = 38\alpha_1$).

7. Acknowledgment

Acknowledgment is made to the Engineer-in-Chief of the G.P.O. for permission to make use of the information included in this paper.

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APPENDIX 1

Operational Formulae Used

$(t > 0)$

$$\frac{1}{p}1 = t$$

$$\frac{p}{p+\alpha}1 = e^{-\alpha t}$$

$$\frac{\alpha}{p+\alpha}1 = 1 - \frac{p}{p+\alpha}1$$
$$= 1 - e^{-\alpha t}$$

$$\frac{p}{(p+\alpha)^2}1 = te^{-\alpha t}$$

$$\frac{\alpha^2}{(p+\alpha)^2}1 = \frac{\alpha}{p+\alpha}1 - \frac{\alpha p}{(p+\alpha)^2}1$$
$$= 1 - e^{-\alpha t}(1+\alpha t)$$

$$\frac{p^2}{(p+\alpha)^2}1 = \frac{p}{(p+\alpha)}1 - \frac{\alpha p}{(p+\alpha)^2}1$$
$$= e^{-\alpha t}(1-\alpha t)$$

APPENDIX 2

Overloading with Step-Voltage Input

(*a*) *Single-Stage Amplifier*

When the input voltage is a unit step, the grid voltage of the valve, from equation (1) is

$$v_g = \frac{p+\alpha_1}{p+\alpha}1 \qquad \text{[from equation (1)]}$$

where $\alpha = \alpha_1(1 + \mu_1\beta)$

$$\therefore \quad v_g = e^{-\alpha t} + \frac{\alpha_1}{\alpha}(1 - e^{-\alpha t}) \quad \text{for } t > 0$$

$$= e^{-\alpha t} + \frac{1}{1+\mu_1\beta}(1 - e^{-\alpha t}) \quad \ldots \quad \ldots \quad (22)$$

The grid voltage rises immediately to the input voltage and then decays to $1/(1 + \mu_1\beta)$ of the input voltage. To avoid overloading, therefore, the input voltage must not exceed the permissible grid swing of the valve.

(*b*) *Two-Stage Amplifier*

When the input voltage is a unit step, the grid voltage of the second stage, from equation (10), is

$$v_{g2} = \frac{\mu_1}{1+\mu_0\beta_0} \cdot \frac{p+\alpha_2}{\alpha_2} \cdot \frac{\alpha^2}{(p+\alpha)^2}1$$
$$= \frac{\mu_1}{1+\mu_0\beta_0}\left\{\frac{\alpha^2}{(p+\alpha)^2} + \frac{\alpha^2 p}{\alpha_2(p+\alpha)^2}\right\}1$$
$$= \frac{\mu_1}{1+\mu_0\beta_0}\left\{1 - e^{-\alpha t}(1+\alpha t) + \frac{\alpha^2}{\alpha_2}te^{-\alpha t}\right\} \quad \text{for } t > 0$$
$$= \frac{\mu_1}{1+\mu_0\beta_0}\left\{1 - e^{-\alpha t}\left[1 - \alpha t\left(\frac{\alpha}{\alpha_2} - 1\right)\right]\right\} \quad \ldots \quad \ldots \quad (23)$$

If $\alpha \leqslant \alpha_2$, v_{g2} increases monotonically to $\mu_1/(1 + \mu_0\beta_0)$, but if $\alpha > \alpha_2$, v_{g2} has a maximum at $t = 1/(\alpha - \alpha_2)$; the maximum value of v_{g2} is then

$$V_{g2}(0) = \frac{\mu_1}{1+\mu_0\beta_0}\left\{1 + \left(\frac{\alpha}{\alpha_2} - 1\right)e^{-\left(\frac{\alpha}{\alpha-\alpha_2}\right)}\right\}$$

$$\therefore \frac{V_{g2}(0)}{V_{g2}(\infty)} = 1 + \left(\frac{\alpha}{\alpha_2} - 1\right)e^{-\left(\frac{\alpha}{\alpha-\alpha_2}\right)} \quad \ldots \quad (15)$$

Fig. 6 shows the variation of this ratio with α_1/α_2 for several values of feedback factor.

The grid voltage of the first stage, from equation (16), is

$$v_{g1} = \frac{1}{1+\mu_0\beta_0} \cdot \frac{p+\alpha_1}{\alpha_1} \cdot \frac{p+\alpha_2}{\alpha_2} \cdot \frac{\alpha^2}{(p+\alpha)^2}1$$

But $\alpha^2 = \alpha_1\alpha_2(1 + \mu_0\beta_0)$ [equation (7a)]

$$\therefore v_{g1} = \left\{\frac{p^2}{(p+\alpha)^2} + \frac{(\alpha_1+\alpha_2)p}{(p+\alpha)^2} + \frac{\alpha_1\alpha_2}{(p+\alpha)^2}\right\}1$$
$$= (1-\alpha t)e^{-\alpha t} + (\alpha_1+\alpha_2)te^{-\alpha t} + \frac{\alpha_1\alpha_2}{\alpha}\left[1 - e^{-\alpha t}(1+\alpha t)\right] \quad \text{for } t > 0$$
$$= \frac{\alpha_1\alpha_2}{\alpha^2} + e^{-\alpha t}\left(1 - \frac{\alpha_1\alpha_2}{\alpha^2}\right) - \left(1 - \frac{\alpha_1}{\alpha}\right)\left(1 - \frac{\alpha_2}{\alpha}\right)\alpha te^{-\alpha t} \quad \ldots \quad \ldots \quad (24)$$

Therefore v_{g1} instantly rises to unity and then decays to $1/(1+\mu_0\beta_0)$. If $\alpha < \alpha_1$ or $\alpha < \alpha_2$, this decay is monotonic, but if $\alpha > \alpha_1$ and $\alpha > \alpha_2$ there is a single overshoot.

(c) Three-Stage Amplifier

$$v_{g3} = v_{a2} + v_{g1} - v_i \quad \text{[equation (21)]}$$

When the input voltage is a unit step, v_{a2} is given by equation (6b) and v_{g1} by equation (24).

$$\therefore \quad v_{g3}(t) = \frac{\mu_0}{1+\mu_0\beta_0} - \frac{\mu_0}{1+\mu_0\beta_0}e^{-\alpha t}(1+\alpha t) + \frac{\alpha_1\alpha_2}{\alpha^2} + e^{-\alpha t}\left(1-\frac{\alpha_1\alpha_2}{\alpha^2}\right) - \left(1-\frac{\alpha_1}{\alpha}\right)\left(1-\frac{\alpha_2}{\alpha}\right)\alpha t e^{-\alpha t} - 1$$

$$\text{for } t > 0 \quad \ldots \quad \ldots \quad \ldots \quad (25)$$

$$\text{Now } v_{g3}(\infty) = \frac{\mu_0}{1+\mu_0\beta_0} + \frac{\alpha_1\alpha_2}{\alpha^2} - 1 \quad \text{[from equation (25)]}$$

$$\therefore \quad v_{g3}(t) \leqslant v_{g3}(\infty) \text{ if}$$

$$\frac{\mu_0}{1+\mu_0\beta_0}e^{-\alpha t}(1+\alpha t) - e^{-\alpha t}\left(1-\frac{\alpha_1\alpha_2}{\alpha^2}\right) + \left(1-\frac{\alpha_1}{\alpha}\right)\left(1-\frac{\alpha_2}{\alpha}\right)\alpha t e^{-\alpha t} \geqslant 0$$

$$\therefore \left[\frac{\mu_0}{1+\mu_0\beta_0} + \frac{\alpha_1\alpha_2}{\alpha^2} - 1\right] + \alpha t\left[\frac{\mu_0}{1+\mu_0\beta_0} + \left(1-\frac{\alpha_1}{\alpha}\right)\left(1-\frac{\alpha_2}{\alpha}\right)\right] \geqslant 0$$

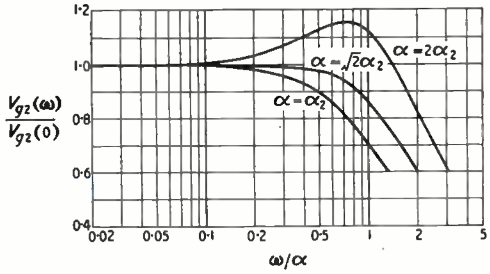

Fig. 11. *Voltage applied to second stage when input signal is sinusoidal.*

This inequality is satisfied for all values of $t (> 0)$ because

$$\left[\frac{\mu_0}{1+\mu_0\beta_0} + \frac{\alpha_1\alpha_2}{\alpha^2} - 1\right] = \frac{\mu_0(1-\beta_0)}{1+\mu_0\beta_0} \geqslant 0 \text{ for } \beta_0 \leqslant 1$$

and

$$\left[\frac{\mu_0}{1+\mu_0\beta_0} + \left(1-\frac{\alpha_1}{\alpha}\right)\left(1-\frac{\alpha_2}{\alpha}\right)\right] = 1 + \frac{1+\mu_0}{1+\mu_0\beta_0} - \frac{\alpha_1+\alpha_2}{\alpha} \geqslant 0 \text{ for } \beta_0 \leqslant 1$$

[because $(\alpha_1 + \alpha_2)/\alpha \leqslant 2$ from inequality (7c)].

The voltage applied to the third valve when the input to the amplifier is a unit step therefore never exceeds its steady-state value. This valve cannot, therefore, be overloaded by applying to the amplifier a step voltage whose magnitude is less than that of the slowly-changing input signal required to overload the valve.

APPENDIX 3

Overloading with a Sinusoidal Input Voltage

(a) Single-Stage Amplifier

$$V_g(\omega) = \frac{\alpha_1 + j\omega}{\alpha + j\omega}V_i \quad \text{[from equation (1)]}$$

$$\therefore \quad \left|\frac{V_g(\omega)}{V_i}\right| = \frac{\alpha_1}{\alpha}\sqrt{\frac{1+(\omega/\alpha_1)^2}{1+(\omega/\alpha)^2}}$$

$$\therefore \quad \left|\frac{V_g(\omega)}{V_g(0)}\right| = \sqrt{\frac{1+(\omega/\alpha_1)^2}{1+(\omega/\alpha)^2}} \quad \ldots \quad \ldots \quad \ldots \quad (26)$$

$$> 1 \text{ for } \omega > 0 \text{ because } \alpha > \alpha_1.$$

As the frequency increases, the proportion of the input voltage which is applied to the valve increases; consequently the input voltage required to overload the amplifier decreases as the frequency increases.

(b) Two-Stage Amplifier

$$V_{g2}(\omega) = \frac{\mu_1}{1+\mu_0\beta_0}\cdot\frac{\alpha^2(\alpha_2+j\omega)}{\alpha_2(\alpha+j\omega)^2}V_i \quad \text{[from equation (10)]}$$

$$\therefore \quad \left|\frac{V_{g2}(\omega)}{V_i}\right| = \frac{\mu_1}{1+\mu_0\beta_0}\frac{\sqrt{1+(\omega/\alpha_2)^2}}{1+(\omega/\alpha)^2}$$

$$\therefore \quad \left|\frac{V_{g2}(\omega)}{V_{g2}(0)}\right|^2 = \frac{1+(\omega/\alpha_2)^2}{1+2(\omega/\alpha)^2+(\omega/\alpha)^4} \quad \ldots \quad \ldots \quad (27)$$

$$\therefore \quad |V_{g2}(\omega)| \leqslant |V_{g2}(0)| \quad \text{for all frequencies.}$$

$$\text{if} \quad \alpha^2 \leqslant 2\,\alpha_2^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (28a)$$

$$\therefore \quad 1 + \mu_0\beta_0 \leqslant 2\,\alpha_2/\alpha_1 \quad \text{[from equation (7a)]} \quad \ldots \quad (28b)$$

Fig. 11 shows the variation of $|V_{g2}|$ with frequency [from equation (27)] for several values of α/α_2.

$$V_{g1}(\omega) = \frac{1}{1+\mu_0\beta_0}\cdot\frac{(\alpha_1+j\omega)(\alpha_2+j\omega)}{\alpha_1\alpha_2}\cdot\frac{\alpha^2}{(\alpha+j\omega)^2}\cdot V_i \quad \text{[from equation (16)]}$$

$$\therefore \quad \left|\frac{(V_{g1}(\omega)}{V_i}\right| = \frac{1}{1+\mu_0\beta_0}\cdot\frac{[1+(\omega/\alpha_1)^2]^{\frac{1}{2}}[1+(\omega/\alpha_2)^2]^{\frac{1}{2}}}{1+(\omega/\alpha)^2}$$

$$\therefore \quad \left|\frac{V_{g1}(\omega)}{V_{g1}(0)}\right|^2 = \frac{1+\omega^2[1/\alpha_1^2+1/\alpha_2^2]+\omega^4/\alpha_1^2\alpha_2^2}{1+2\omega^2/\alpha^2+\omega^4/\alpha^4}$$

Now $1/\alpha_1^2\alpha_2^2 > 1/\alpha^4$ for $\mu_0\beta_0 > 0$ [from equation (7a)] and $1/\alpha_1^2 + 1/\alpha_2^2 \geqslant 2/\alpha_1\alpha_2 > 2/\alpha^2$

$\therefore |V_{g1}(\omega)| \geqslant |V_{g1}(0)|$ for all frequencies.

As the frequency of the applied signal is increased from zero to infinity, the voltage applied to the first valve increases from $V_i/(1+\mu_0\beta_0)$ to V_i.

(c) Three-Stage Amplifier

$$V_{g3}(\omega) = V_{a2}(\omega) + V_{g1}(\omega) - V_i(\omega) \quad \text{[from equation (21)]}$$

$\therefore$ From equations (8) and (16):

$$V_{g3}(\omega) = V_i(\omega)\left\{\frac{\mu_0\alpha^2}{(1+\mu_0\beta_0)(\alpha+j\omega)^2} + \frac{\alpha^2(\alpha_1+j\omega)(\alpha_2+j\omega)}{\alpha_1\alpha_2(1+\mu_0\beta_0)(\alpha+j\omega)^2} - 1\right\}$$

$$= V_i(\omega)\frac{\mu_0 + (1+j\omega/\alpha_1)(1+j\omega/\alpha_2) - (1+\mu_0\beta_0)(1+j\omega/\alpha)^2}{(1+\mu_0\beta_0)(1+j\omega/\alpha)^2}$$

$$\therefore \left|\frac{V_{g3}(\omega)}{V_i(\omega)}\right| = \frac{\sqrt{\mu_0^2(1-\beta_0)^2 + \omega^2\left[\left(\frac{\alpha_1+\alpha_2}{\alpha_1\alpha_2}\right) - 2\left(\frac{1+\mu_0\beta_0}{\alpha}\right)\right]^2}}{(1+\mu_0\beta_0)(1+\omega^2/\alpha^2)^2}$$

$$\therefore \left|\frac{V_{g3}(\omega)}{V_{g3}(0)}\right|^2 = \frac{1 + \frac{\omega^2}{\mu_0^2(1-\beta_0)^2}\left[\left(\frac{\alpha_1+\alpha_2}{\alpha_1\alpha_2}\right) - 2\left(\frac{1+\mu_0\beta_0}{\alpha}\right)\right]^2}{1 + 2\frac{\omega^2}{\alpha^2} + \frac{\omega^4}{\alpha^4}} \quad (28)$$

Now $|V_{g3}(\omega)| \leqslant |V_{g3}(0)|$ at all frequencies if

$$\frac{2}{\alpha^2} \geqslant \frac{1}{\mu_0^2(1-\beta_0)^2}\left[\left(\frac{\alpha_1+\alpha_2}{\alpha_1\alpha_2}\right) - 2\left(\frac{1+\mu_0\beta_0}{\alpha}\right)\right]^2$$

$$\therefore \frac{2\mu_0^2(1-\beta_0)^2}{(1+\mu_0\beta_0)^2} \geqslant \left[\frac{\alpha_1+\alpha_2}{\alpha} - 2\right]^2 \quad \text{[from equation (7a)]}$$

$$\text{But } \frac{\alpha_1+\alpha_2}{\alpha} \leqslant 2 \quad \text{[from inequality (7c)]}$$

$$\therefore \frac{\sqrt{2}\mu_0(1-\beta_0)}{1+\mu_0\beta_0} \geqslant 2 - \frac{\alpha_1+\alpha_2}{\alpha} < 2$$

$$\text{But } \frac{\mu_0(1-\beta_0)}{1+\mu_0\beta_0} = \frac{\mu_0+1}{1+\mu_0\beta_0} - 1 > \frac{\mu_0}{1+\mu_0\beta_0} - 1$$

Therefore, when $\frac{\mu_0}{1+\mu_0\beta_0} \geqslant 1+\sqrt{2}$

then $|V_{g3}(\omega)| \leqslant |V_{g3}(0)|$ for all frequencies.

The voltage applied to the third valve at high frequencies is therefore less than that at low frequencies when the gain of the first two stages (with feedback) is greater than 2.4.

HARMONIC DISTORTION OF MODULATION

Production by Echoes in a Radio System

By E. G. Hamer, B.Sc. (Eng.) Hons.

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IN most types of communication networks where radio links are used distortion may be caused by echo signals. This distortion is additional to that caused by non-linearity in valves, tuned circuits, and the aerial array itself. These echo signals may be produced by multipath propagation, by stations in a network using the same radio frequency, or a mismatched aerial-feeder cable. The former cases, especially when frequency modulation is used, have been dealt with extensively in previous papers[1,2,3,4] and in the latter case where frequency modulation is used by L. Lewin.[5] Distortion caused by feeder-cable echoes can to some extent be controlled by careful design although the results obtained in previous papers are either for echo signals of the same order of magnitude as the main radio-frequency signal, or alternatively, are not in a very convenient form to use.

In most instances distortion due to feeder echo is only of importance where multichannel systems are used, and the results derived in the paper are expressed in the form of nomograms so that the amount of distortion may be readily estimated over a wide range of modulating frequencies, and depths of modulation, or frequency deviation. As the time delay is varied there is a slow change due to the modulation phase-angle difference, and a more rapid change due to the carrier phase-angle difference. Small changes of feeder length or reflection coefficient angle cause large changes of carrier phase angle, and in all cases the maximum value of distortion for the worst carrier phase angle has been evaluated.

Systems Using Amplitude Modulation

The analysis of the effects of an echo signal in a system using amplitude modulation is dealt with in Appendix 1. The resultant signal is the combination of two carrier waves of the same frequency, carrying the same modulating frequency but with different phase angles, and depths of modulation; and this gives a final wave which is modulated both in frequency and in amplitude. If a perfectly linear detector is assumed the effects of frequency modulation may be neglected[4] and if

$\omega_c = 2\pi \times$ carrier frequency
$\omega_a = 2\pi \times$ modulating frequency (f_a)
$m =$ degree of modulation
$r =$ voltage ratio of $\frac{\text{echo signal}}{\text{main signal}}$
$t_0 =$ time delay of echo signal.

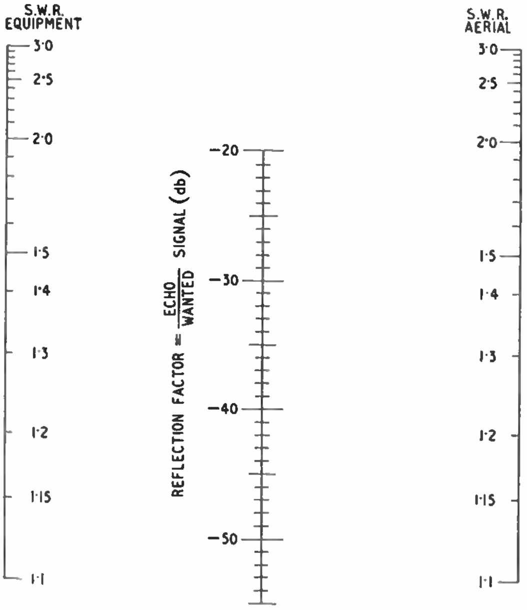

Fig. 1. *Nomogram for reflection factor of aerial feeder.*

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then the maximum values of the distortion components are:—

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = \frac{mr^2}{4} \sin^2 \omega_a t_0 \quad \ldots \quad (1)$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = \frac{m^2 r^2}{8} \sin^2 \omega_a t_0 \quad \ldots \quad \ldots \quad (2)$$

The actual value of the distortion varies cyclically with changes of carrier phase angle between the echo and main signals; so that small changes of feeder length, or reflection coefficient will cause the distortion to vary between zero and the maximum value stated. There is also a slow cyclic change as t_0 is increased, but in nearly all practical cases the first maximum is never reached.

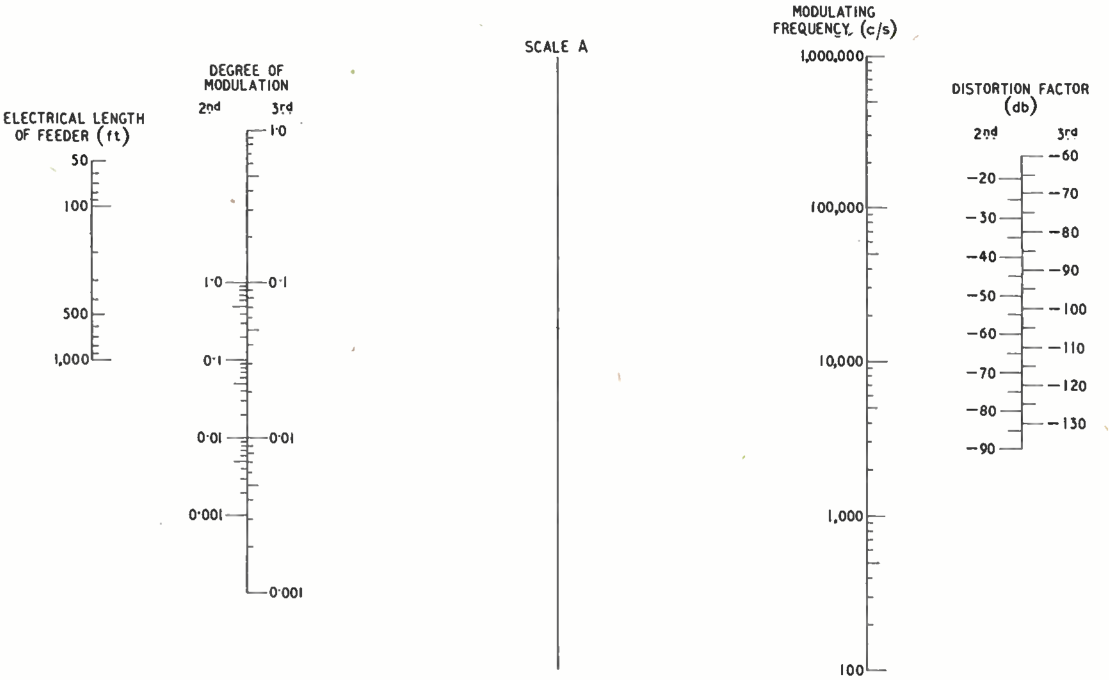

Fig. 2. *Diagrams showing distortion in a.m. system due to single-feeder cable. To find distortion factor—Draw line through appropriate degree of modulation and modulating frequency and find intersection on scale A. Join intersection on scale A to electrical length and produce to obtain intersection on appropriate distortion-factor scale.*

Fig. 1 is a nomogram for the determination of r when the standing-wave ratio is known at each end of the feeder. It is necessary to take into account the mismatch due to the generator as in this case, due to the dynamic conditions involved, it is one of the factors determining the magnitude of the echo signal. The value of r is expressed in decibels as a 'Reflection Factor', and twice the value must be added to the 'Distortion Factor' obtained from Fig. 2 to determine the total harmonic distortion.

Fig. 2 is a nomogram for equations (1) and (2) where t_0 has been expressed in the form of the electrical length of feeder.

$t_0 = 2l/v$

if l = electrical length of feeder

v = velocity of propagation

any small effects due to the reflection coefficient angles are neglected, and only the maximum value of the 'Distortion Factor' is shown. In a particular case the actual value will depend upon the exact length of feeder, angle of reflection coefficient, and carrier frequency.

Systems Using Frequency Modulation

The analysis for systems using frequency modulation is dealt with in Appendix 2, and the resultant wave is seen to contain both amplitude- and frequency-modulated components. Assuming ideal limiters the maximum values of the distortion components are:—

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = \frac{4rf_a}{F_D} J_2\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right) \quad \ldots \quad (3)$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = \frac{6rf_a}{F_D} J_3\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right) \quad \ldots \quad (4)$$

where F_D = frequency deviation and J_n is the Bessel Function of the 1st kind (nth order).

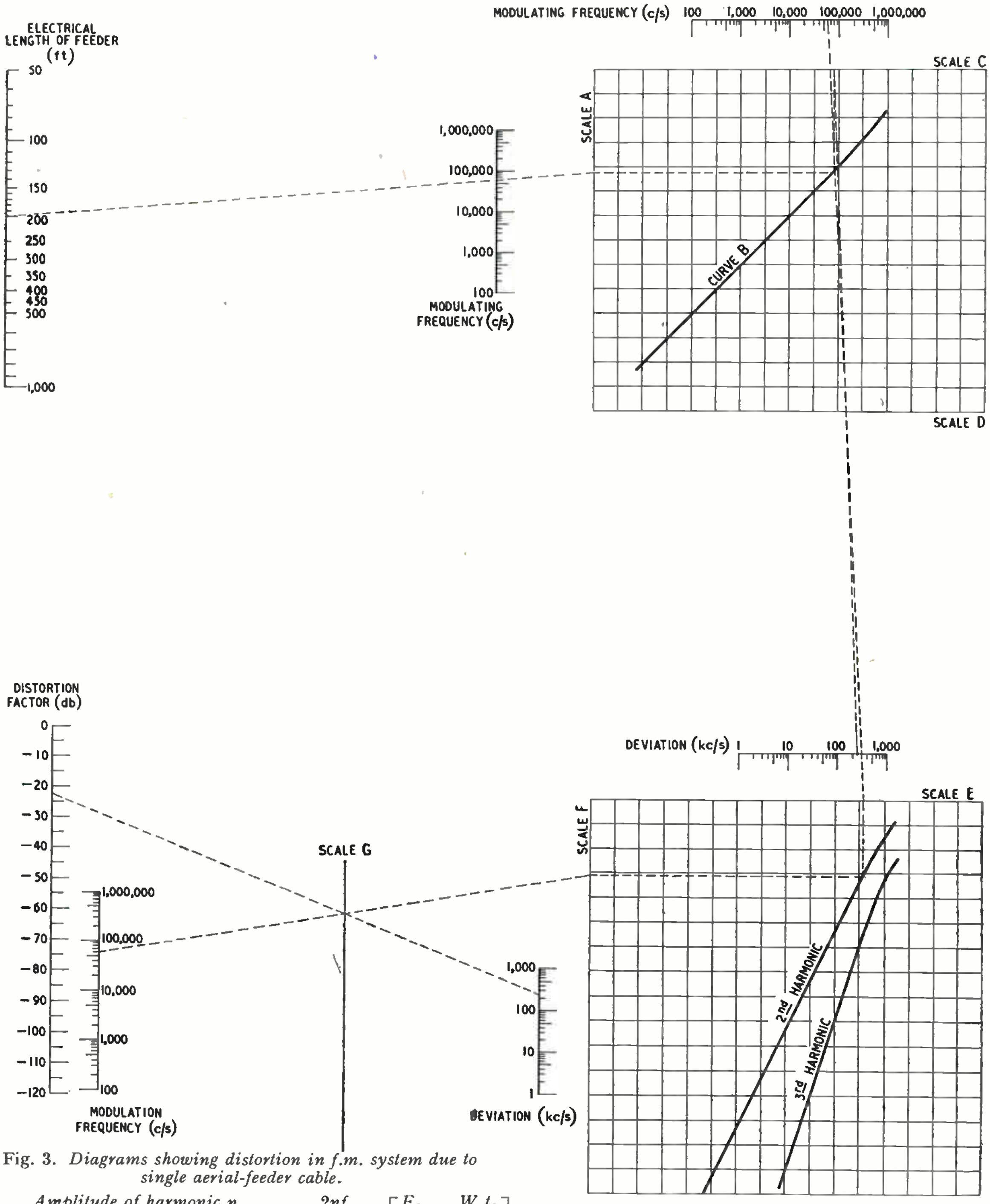

Fig. 3. *Diagrams showing distortion in f.m. system due to single aerial-feeder cable.*

$$\frac{\text{Amplitude of harmonic } n}{\text{Amplitude of fundamental}} = r \times \frac{2nf_a}{F_0} J_n \left[\frac{F_0}{fa} \sin \frac{W_a t_0}{2} \right]$$

To find distortion factor:—Draw line through electrical length of feeder and modulating frequency, find intersection on scale A. Transfer from scale A to scale C using curve B, join intersection of scale C to deviation. Find intersection on scale D, join intersection on scale D to modulating frequency. Obtain intersection on scale E, transfer from scale E to scale F, using curve for required harmonic. Join scale F to modulating frequency to obtain intersection on scale G. Join deviation and scale G to obtain intersection on distortion-factor scale. Example shown for—$f_a = 60$ kc/s; $F_D = 250$ kc/s; feeder = 200 ft, distortion for 2nd harmonic.

Again the actual value of the distortion varies cyclically with changes of carrier phase angle between the echo and main signals. In this case, however, when the even harmonics have a maximum value the odd harmonics are zero and vice-versa. There is also a slow cyclic change as t_0 is increased, but in all practical cases the first maximum is never reached.

Fig. 3 is a nomogram for equations (3) and (4) where t_0 has been expressed in the form of the electrical length of feeder. The total distortion in this case is the sum of the reflection factor (as obtained from Fig. 1) and the distortion factor. It can be seen from Fig. 3 that over most of the range $\sin \frac{\omega_a t_0}{2} \approx \frac{\omega_a t_0}{2}$, and the Bessel functions are sufficiently small to be approximately equal to the first term of the series expansion. In this case

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = 2\ r f_a F_D \pi^2 t_0^2 \quad \ldots \quad \ldots \quad (5)$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = r f_a F_D{}^2 \pi^3 t_0{}^3 \quad \ldots \quad \ldots \quad (6)$$

and Fig. 4 is a nomogram for the approximate solution expressed by equations (5) and (6).

Conclusions

From the equations and nomograms it can be seen that a frequency-modulated system is much more vulnerable to the effects of echo signals than an amplitude-modulated system. In the case of an aerial feeder the deleterious effects of echo signals may be minimized by careful matching of the feeder at one or both ends. For any given carrier frequency the distortion components may be minimized, or in certain cases reduced to zero, by selecting the length of feeder, or controlling the angles of the reflection coefficients.

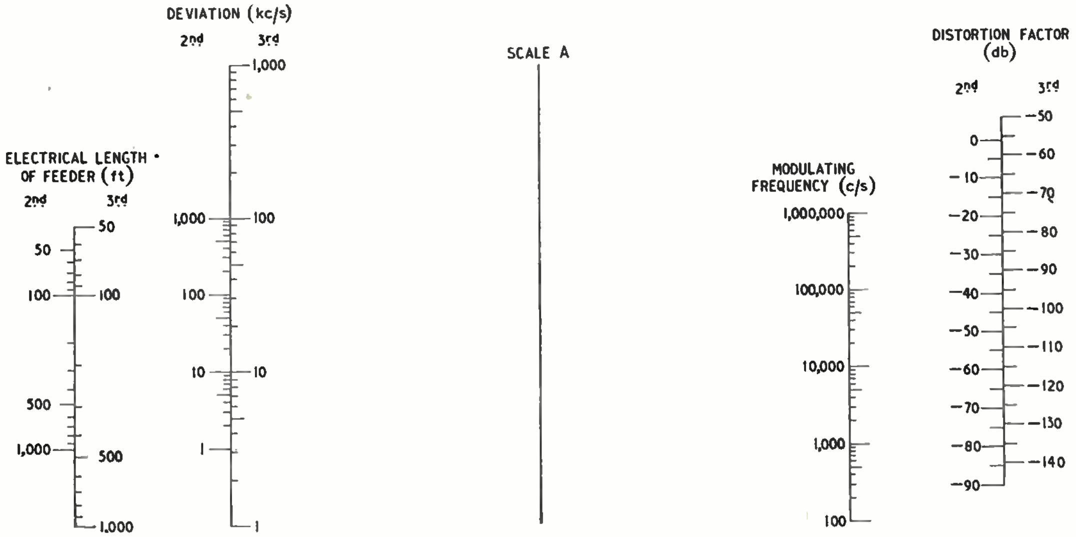

Fig. 4. *Diagram showing distortion in f.m. system due to single feeder cable. To find distortion factor:—Draw line through appropriate deviation and modulating frequency and find intersection on scale A. Join intersection on scale A to appropriate electrical length and produce to obtain intersection on appropriate distortion-factor scale.*

APPENDIX 1

Distortion of Amplitude-Modulated Signals.

Original Signal $= E(1 + m \sin \omega_a t) \sin \omega_c t$

Echo Signal $= rE[1 + m \sin \omega_a(t + t_0)] \sin \omega_c(t + t_0)$

The combined signal if $\cos \omega_a t_0$ is assumed equal to 1 is

$$S = E\{[\sin \omega_c t + r \sin \omega_c(t + t_0)][1 + m \sin \omega_a t] + rm \sin \omega_a t_0 \cos \omega_a t \sin \omega_c(t + t_0)\}$$

This is seen to be a combination of two carrier waves of the same frequency with different amounts of modulation

If, $k^2 = 1 + r^2 + 2r \cos \omega_c t_0$

$K = rm \sin \omega_a t_0$

$$\tan \alpha = \frac{r \sin \omega_c t_0}{1 + r \cos \omega_c t_0}$$

$$S = E\ \{k \sin (\omega_c t + \alpha)[1 + m \sin \omega_a t] + K \cos \omega_a t \sin \omega_c(t + t_0)\}$$

and neglecting any effects due to frequency modulation of the carrier wave, and taking the equation of the envelope[5] we have:

$$\text{Envelope} = E\{k^2(1 + m \sin \omega_a t)^2 + K^2 \cos^2 \omega_a t + 2kK \cos \omega_a t(1 + m \sin \omega_a t) \cos (\omega_c t_0 + \alpha)\}^{\frac{1}{2}}$$

and this is the equation of the resultant output assuming a perfect detector circuit.

Collecting terms we have:—

$$\text{Envelope} = Ek\{1 + [2m \sin \omega_a t + \frac{2K}{k} \cos (\omega_c t_0 - \alpha) \cos \omega_a t] + [m^2 \sin^2 \omega_a t + \frac{K^2}{k^2} \cos^2 \omega_a t + \frac{2Km}{k} \cos (\omega_c t_0 - \alpha) \sin \omega_a t \cos \omega_a t]\}^{\frac{1}{2}}$$

This may now be expanded by the Binomial theorem, and collecting the various terms, and neglecting contributions from high order terms we have:

fundamental component =

$$Ek\left[m^2 + \frac{K^2}{k^2}\cos^2(\omega_c t_0 - \alpha)\right]^{\frac{1}{2}}$$

2nd harmonic component $= Ek\left[\frac{K^2}{4k^2}\sin^2(\omega_c t_0 - \alpha)\right]$

3rd harmonic component =

$$Ek\left[\frac{K^2}{8k^2}\sin^2(\omega_c t_0 - \alpha)\right]\left[m^2 + \frac{K^2}{k^2}\cos^2(\omega_c t_0 - \alpha)\right]^{\frac{1}{2}}$$

From these equations it will be seen that as t_0 increases there is a slow increase in the amount of the distortion, superimposed on a more rapid change.

If $\sin(\omega_c t_0 - \alpha) = 0$

fundamental component =

$$Ek\left(m^2 + \frac{K^2}{k^2}\right)^{\frac{1}{2}} \approx E(1 + r)m(1 + r^2\sin^2\omega_a t_0)^{\frac{1}{2}}$$

2nd harmonic component = 0

3rd harmonic component = 0

If $\cos(\omega_c t_0 - \alpha) = 0$

fundamental component $= Ekm = Em(1 - r^2)^{\frac{1}{2}}$

2nd harmonic component $= Ek\,\frac{K^2}{4k^2} = \frac{Em^2r^2\sin^2\omega_a t_0}{4(1 - r^2)^{\frac{1}{2}}}$

3rd harmonic component $= Ek\,\frac{mK^2}{8k^2} = \frac{Em^3r^2\sin^2\omega_a t_0}{8(1 - r^2)^{\frac{1}{2}}}$

Taking these maximum values of distortion we have if r is small

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = \frac{mr^2}{4}\sin^2\omega_a t_0$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = \frac{m^2r^2}{8}\sin^2\omega_a t_0$$

It can easily be seen that the maximum value of the 2nd harmonic will occur when $\cos(\omega_c t_0 - \alpha) = 0$; and when considering small values of r then $\frac{K^2}{k^2}\cos^2(\omega_c t_0 - \alpha)$ will be small and the maximum value of the 3rd harmonic will occur when $\cos(\omega_c t_0 - \alpha)$ is nearly equal to zero.

APPENDIX 2

Distortion of Frequency-Modulated Signal.

Original Signal $= E\sin\left[\omega_c t + \frac{F_D}{f_a}\cos\omega_a t\right]$

Echo Signal $= rE\sin\left[\omega_c(t + t_0) + \frac{F_D}{f_a}\cos\omega_a(t + t_0)\right]$

where F_D = frequency deviation.

The combined signal is

$$S = E[(1 + r\cos A)\sin B - r\cos B\sin A]$$

$$\text{if } A = 2\frac{F_D}{f_a}\sin\left(\omega_a t + \frac{\omega_a t_0}{2}\right)\sin\frac{\omega_c t_0}{2} - \omega_c t_0$$

$$\text{and } B = \omega_c t + \frac{F_D}{f_a}\cos\omega_a t$$

$$\text{Let } \tan\theta = \frac{r\sin A}{1 + r\cos A}$$

Then $S = E[r^2\sin^2 A + (1 + r\cos A)^2]^{\frac{1}{2}}\sin(\theta - B)$

Now the first part of this represents the amplitude-modulated term which will be removed by limiting, and the only term of interest is that containing the varying frequency component.

Hence the signal received, if we assume a perfect discriminator, is proportional to the time derivative of $\sin(\theta - B)$, B is the component due to the original signal and θ is the added distortion component due to the echo signal.

If we assume r is small then

$$\tan\theta = \frac{r\sin A}{1 + r\cos A} \approx r\sin A$$

$$\text{and } \theta = r\sin\left[2\frac{F_D}{f_a}\sin\left(\omega_a t - \frac{\omega_a t_0}{2}\right)\sin\frac{\omega_a t_0}{2} - \omega_c t_0\right]$$

$$= r\left[\sin\left\{2\frac{F_D}{f_a}\sin\left(\omega_a t + \frac{\omega_a t_0}{2}\right)\sin\frac{\omega_a t_0}{2}\right\}\cos\omega_c t_0 - \cos\left\{2\frac{F_D}{f_a}\sin\left(\omega_a t + \frac{\omega_a t_0}{2}\right)\sin\frac{\omega_a t_0}{2}\right\}\sin\omega_c t_0\right]$$

and expanding these terms using Bessel expansions:—

$\cos(x\sin\phi) = J_0(x) + J_2(x)2\cos 2\phi + J_4(x)2\cos 4\phi \ldots$ etc.

$\sin(x\sin\phi) = 2J_1(x)\sin\phi + 2J_3(x)\sin 3\phi \ldots$ etc.

$$\text{Then } \theta = r\cos\omega_c t_0[2J_1(x)\sin\phi + 2J_3(x)\sin 3\phi \ldots.] + r\sin\omega_c t_0[J_0(x) + 2J_2(x)\cos 2\phi \ldots.]$$

$$\text{if } x = 2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}$$

$$\text{and } \phi = \omega_a t + \frac{\omega_a t_0}{2}$$

and differentiating to give the actual components of audio distortion we have

fundamental $= \frac{F_D\omega_a}{f_a}\sin\omega_a t + r.\cos\omega_c t_0.\,\omega_a.\,J_1(x)\sin\phi$

2nd harmonic $= 4r\omega_a\sin\omega_c t_0.\,J_2(x).\,\sin 2\phi$

3rd harmonic $= 6r\omega_a\cos\omega_c t_0 J_3(x)\cos 3\phi$

Neglecting the phase relationships of the harmonics and the fundamental, and the small contribution to the fundamental from the distortion terms we have if r is small

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = \frac{4rf_a}{F_D}\sin\omega_c t_0\,J_2\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right)$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = \frac{6rf_a}{F_D}\cos\omega_c t_0 J_3\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right)$$

and the maximum values of these expressions are:—

$$\frac{\text{2nd harmonic}}{\text{fundamental}} = \frac{4rf_a}{F_D}J_2\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right)\text{when } \sin\omega_c t_0 = 1$$

$$\frac{\text{3rd harmonic}}{\text{fundamental}} = \frac{6rf_a}{F_D}J_3\left(2\frac{F_D}{f_a}\sin\frac{\omega_a t_0}{2}\right)\text{when } \cos\omega_c t_0 = 1$$

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SPECTRUM OF A FREQUENCY-MODULATED WAVE

By W. C. Vaughan, M.B.E., B.Sc., Ph.D., A.M.I.E.E.

1. Introduction

ALMOST without exception, textbooks describing the frequency spectrum of a frequency-modulated wave do so without explanation, usually suggesting instead that the reader should refer to one or more of the many works dealing with the applications of Bessel functions. Disappointment with the outcome of his efforts to obtain help from these sources often causes the student to regard the results of the analysis merely as factual information, the derivation of which he makes no further attempt to understand. Thereafter the apparent complexity of the problem remains in striking contrast to the simplicity of the analysis of the amplitude-modulated wave. Moreover, failure to understand how the multiple-sideband spectrum of the frequency-modulated wave has been derived has the unfortunate consequence of requiring blind acceptance of a seemingly implausible concept. While some of the difficulties in the way of accepting the results of this particular process may often be traced to fallacious ideas of 'instantaneous frequency,' the latter are less easily avoided when the analysis itself remains a mystery. Some knowledge of the theory and uses of Bessel functions is undoubtedly very useful to the engineer and more so to the physicist, but against the value of this knowledge must be set the serious demands which such study inevitably makes both on time and mathematical ability. Quite apart from this, the Bessel functions do not in themselves offer the immediate appeal of many other branches of mathematics.*

Bessel functions are solutions of Bessel's equation, a differential equation associated with problems in many fields of physics and engineering. Their connection with the subject of frequency-modulation is quite incidental, since in this instance the required analysis involves nothing more than the expansion of a trigonometrical function into a particular form of series, the coefficients of which happen to be Bessel functions. We shall show that the derivation of general expressions for those functions need presume knowledge only of those elementary concepts with which all radio engineers should be thoroughly familiar. Evaluation of these coefficients, although necessary, is a matter on which mathematical textbooks for engineers are curiously reticent. The methods of evaluation to be described yield results of any desired accuracy by simple arithmetical processes which rest on elementary mathematical arguments. Though these arguments may fail to satisfy the rigour of the pure mathematician's concern for the general case, this shortcoming should not detract from their appeal to the engineer, for whom they are, of course, intended.

2. The Nature of the Problem

A voltage of peak value V varying sinusoidally with time and having a frequency $f = \frac{\omega}{2\pi}$ may be represented by a rotating vector. The instantaneous voltage v is then given by $v = V \sin \phi$, where ϕ defines the position of the vector at the instant under consideration. The angular velocity of the rotating vector is $\frac{d\phi}{dt} = 2\pi f$, so that

$$\phi = \int 2\pi f \, dt \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

In the case of sinusoidal amplitude-modulation, f is maintained constant and V is varied sinusoidally at the modulation frequency. On the other hand, sinusoidal frequency-modulation at a frequency f_m is effected by keeping V constant and varying f about a mean value f_c in accordance with the relation,

$$f = f_c \left(1 + k \cos 2\pi f_m t\right)$$

k being a numerical constant.

In this instance,

$$\phi = \int 2\pi f_c \left(1 + k \cos 2\pi f_m t\right) dt$$

so that if the time origin is chosen to make the constant of integration zero,

$$\phi = 2\pi f_c t + \frac{k f_c}{f_m} \sin 2\pi f_m t$$

Now $\frac{k f_c}{f_m}$ is termed the modulation index m_f, so that,

$$\phi = \omega_c t + m_f \sin \omega_m t \qquad \ldots \quad \ldots \quad (2)$$

and the instantaneous voltage is, therefore,

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*"It has often been said by analysts with a taste for elegance that no mathematician left to his own devices would ever have dreamed of inventing anything so uncouth mathematically as Bessel functions; or if by chance he had imagined such things in a nightmare, he would have done his utmost to forget them on coming to his senses."
"The Development of Mathematics," E. T. Bell.

MS accepted by the Editor, October 1951

$$\begin{aligned} v &= V \sin(\omega_c t + m_f \sin \omega_m t) \\ &= V[\sin \omega_c t \cos(m_f \sin \omega_m t) + \cos \omega_c t \sin (m_f \sin \omega_m t)] \quad \ldots \quad (3) \end{aligned}$$

In order to determine the response of a circuit to a voltage of this description, it is necessary to find some means of representing it as a series of constant-frequency, constant-amplitude sinusoidal components. Using the well-known power-series forms we may write $\cos(m \sin\theta)$ and $\sin(m\sin\theta)$ as,

$$\cos(m\sin\theta) = 1 - \frac{(m\sin\theta)^2}{2!} + \frac{(m\sin\theta)^4}{4!} \quad \ldots \quad (4)$$

and

$$\sin(m\sin\theta) = m\sin\theta - \frac{(m\sin\theta)^3}{3!} + \frac{(m\sin\theta)^5}{5!} \quad \ldots \quad (5)$$

Furthermore, the expansion of $\left(\frac{\epsilon^{j\theta} - \epsilon^{-j\theta}}{2j}\right)^n$ by means of the binomial theorem leads to the expression of $\sin^n\theta$ in terms of the sines and cosines of the multiples of θ for any particular value of n. By employing these identities, we may write the foregoing power series in a more useful form. Thus equation (4) becomes,

$$\begin{aligned} \cos(m\sin\theta) = 1 + \Big\{ & \frac{m^2}{2!}\left[\frac{\cos 2\theta - 1}{2}\right] \\ & + \frac{m^4}{4!}\left[\frac{\cos 4\theta - 4\cos 2\theta + 3}{8}\right] \\ & + \frac{m^6}{6!}\left[\frac{\cos 6\theta - 6\cos 4\theta + 15\cos 2\theta - 10}{32}\right] \\ & + \ldots\ldots\ldots \Big\} \end{aligned}$$

If we now collect terms in $\cos 2\theta$, $\cos 4\theta$, etc., this expansion becomes,

$$\begin{aligned} \cos(m\sin\theta) = \Big\{ & \left[1 - \frac{m^2}{2\times 2!} + \frac{3m^4}{8\times 4!} - \ldots\right] \\ & + \left[\frac{m^2}{2\times 2!} - \frac{4m^4}{8\times 4!} + \frac{15m^6}{32\times 6!} - \ldots\ldots\right]\cos 2\theta \\ & + \left[\frac{m^4}{8\times 4!} - \frac{6m^6}{32\times 6!} + \ldots\right]\cos 4\theta \\ & + \left[\frac{m^6}{32\times 6!} - \ldots\ldots\right]\cos 6\theta \\ & + \ldots\ldots\ldots \Big\} \end{aligned}$$

In the same manner, substitution in equation (5) leads to a similar expression for $\sin(m\sin\theta)$, although in this case the expansion contains only the sines of odd multiples of θ. The two functions may now be written in the form,

$$\cos(m\sin\theta) = J_0(m) + 2J_2(m)\cos 2\theta + 2J_4(m)\cos 4\theta + \ldots \quad (6)$$

and

$$\sin(m\sin\theta) = 2J_1(m)\sin\theta + 2J_3(m)\sin 3\theta + \ldots \quad \ldots \quad (7)$$

where $J_0(m)$, $J_1(m)$, $J_2(m)$, etc., are numerical coefficients defined by,

$$J_0(m) = \left[1 - \frac{m^2}{2^2} + \frac{m^4}{2^2\times 4^2} - \frac{m^6}{2^2\times 4^2\times 6^2} + \ldots\ldots\right]$$

$$J_1(m) = \frac{m}{2}\left[1 - \frac{m^2}{2\times 4} + \frac{m^4}{2\times 4\times 4\times 6} - \frac{m^6}{2\times 4\times 6\times 4\times 6\times 8} + \ldots\ldots\right]$$

$$J_2(m) = \frac{m^2}{2\times 2!}\left[1 - \frac{m^2}{2\times 6} + \frac{m^4}{2\times 4\times 6\times 8} - \ldots\ldots\right]$$

etc., etc.

It may be readily verified that all these coefficients are represented by the general expression,

$$\begin{aligned} J_n(m) &= \frac{m^n}{2^n\times n!}\left\{1 - \frac{m^2}{2(2n+2)} + \frac{m^4}{2\times 4(2n+2)(2n+4)} \ldots\ldots\right\} \\ &= \sum_{z=0}^{\infty} \frac{(-1)^z}{z!(z+n)!}\left(\frac{m}{2}\right)^{n+2z} \quad \ldots \quad (8) \end{aligned}$$

provided that unit value* is ascribed to 0!

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*The interpretation of 0! will be familiar to those acquainted with the properties of gamma functions, which provide in the relation,

$$n! = \Gamma(n+1)$$

a generalized definition of the factorial. This gives,

$$0! = \Gamma(1)$$

and since by definition,

$$\Gamma(n) = \int_0^\infty x^{n-1}\epsilon^{-x}\,dx$$

$$\Gamma(1) = \int_0^\infty \epsilon^{-x}\,dx = -\left[\epsilon^{-x}\right]_0^\infty = 1$$

To those not so acquainted, the reasonableness of this identity will be apparent from the symmetry which it provides in many series expansions. Thus, for example, the power-series for $\cos x$ may be written,

$$\cos x = \frac{x^0}{0!} - \frac{x^2}{2!} + \frac{x^4}{4!} - \ldots\ldots = \sum_{z=0}^{\infty} (-1)^z \frac{x^{2z}}{(2z)!}$$

3. The Sideband Pattern

We may now express the function in equation (3) as a series of constant-frequency constant-amplitude sinusoidal components by employing relations (6) and (7). Thus

$$v = V\,\{\sin \omega_c t\,[J_0(m_f) + 2J_2(m_f)\cos 2\omega_m t + 2J_4(m_f)\cos 4\omega_m t + \ldots\ldots] + \cos \omega_c t\,[2J_1(m_f)\sin \omega_m t + 2J_3(m_f)\sin 3\omega_m t + \ldots\ldots\ldots\ldots]\}$$

which may be rearranged as follows,

$$\begin{aligned} v = V\{&J_0(m_f)\sin \omega_c t \\ &+ J_1(m_f)\,[\sin(\omega_c + \omega_m)t - \sin(\omega_c - \omega_m)t] \\ &+ J_2(m_f)\,[\sin(\omega_c + 2\omega_m)t + \sin(\omega_c - 2\omega_m)t] \\ &+ J_3(m_f)\,[\sin(\omega_c + 3\omega_m)t - \sin(\omega_c - 3\omega_m)t] \\ &+ \ldots\ldots\} \end{aligned} \qquad \ldots \quad (9)$$

It has now been shown that the frequency-modulated waveform expressed in equation (3) may be represented as a series of sinusoidal components comprising a carrier of frequency f_c and amplitude $J_0(m_f)V$, together with an infinite number of pairs of sidebands of frequencies $(f_c + f_m)$ and $(f_c - f_m)$, $(f_c + 2f_m)$ and $(f_c - 2f_m)$, $(f_c + 3f_m)$ and $(f_c - 3f_m)$, etc., having respective amplitudes $J_1(m_f)V$, $J_2(m_f)V$, $J_3(m_f)V$, etc. The positions of these sidebands in the frequency spectrum depend only on the modulating frequency; their amplitudes, as well as that of the carrier are determined by the modulation index m_f. It should be noted that since no sideband can carry infinite power, the summation in equation (8) must be finite for all values of n.

4. Relation between Adjacent-Order Bessel Coefficients

The general coefficient $J_n(m)$ is termed an nth-order Bessel function of the first kind, of the argument m, and it is readily shown by differentiation that,

$$y = J_n(m) = \frac{1}{2^n n!}\left\{m^n - \frac{m^{n+2}}{2(2n+2)} + \frac{m^{n+4}}{2 \times 4(2n+2)(2n+4)} + \ldots\ldots\right\}$$

satisfies Bessel's equation,

$$\frac{d^2y}{dm^2} + \frac{1}{m}\frac{dy}{dm} + \left(1 - \frac{n^2}{m^2}\right) = 0$$

The direct calculation of the values of Bessel coefficients from the expansion shown in equation (8) presents a formidable problem which seems to lie in the province of the electronic computor, since account must be taken of very many terms to obtain even an approximate result. The more advanced works on the theory of these functions show the possibility of deriving satisfactory approximations when the argument greatly exceeds n, but these cover only part of the range with which we are here concerned. The methods of computation about to be described may, on the other hand, be used to determine the value of $J_n(m)$ for all magnitudes of n and m in which the radio engineer has an interest. For this purpose we shall employ as a basis the relation,

$$J_n(m) = \frac{m}{n}\left[\frac{J_{n+1}(m) + J_{n-1}(m)}{2}\right] \qquad \ldots \quad (10)$$

This identity can be established quite simply by replacing n by $(n - 1)$ and by $(n + 1)$ in equation (8) to obtain expansions for $J_{n-1}(m)$ and $J_{n+1}(m)$.

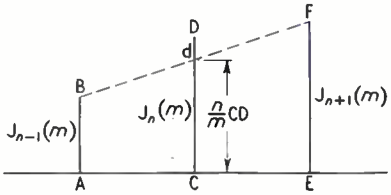

Fig. 1. *Graphical construction showing the relation between adjacent Bessel coefficients.*

Equation (10) leads to a particularly useful graphical construction. Suppose that the equally spaced vertical lines AB, CD and EF in Fig. 1 respectively represent the magnitudes of $J_{n-1}(m)$, $J_n(m)$ and $J_{n+1}(m)$. If B is joined to F, Cd will be the mean of AB and EF, and according to equation (10), Cd/CD must equal n/m. In the special case* when $n = m$, the line BF must pass through D, and the diagram then takes the form of Fig. 2(a), in which AB, CD, EF, etc., represent $J_{m-1}(m)$, $J_m(m)$, $J_{m+1}(m)$, etc., respectively. It follows from this that if $J_{m-1}(m)$ is less than, or differs in sign from $J_m(m)$, $J_m(m)$ will be less than $J_{m-1}(m)$. Taking this argument a stage further, since the line joining D and H must intersect EF produced at f to make $Ef/EF = m + 1/m$, it will be seen that $J_{m+1}(m)$ must be less than $J_{m+2}(m)$, and it thus becomes evident that higher orders of $J_n(m)$ cannot change sign and must rapidly increase in magnitude. This divergence is inconsistent with equation (9), and would give an infinite value to $\sum_{n=0}^{\infty} J_n(m)$. We must, therefore, conclude that $J_{m-1}(m) > J_m(m) > J_{m+1}(m)$, etc. Fig. 2(b) shows a further limitation which must be placed on the magnitude of $J_{m-1}(m)$ in relation to $J_m(m)$. If, for example, AB is more than twice CD, then EF will

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*Although m may, of course, have any value, it has here been assumed to be a whole number. It will be evident that subsequent conclusions regarding the magnitude of $J_n(m)$ for values of n greater than m are not invalidated by this assumption.

differ in sign from CD. In such a case since Ef/EF $= m + 1/m$, GH will represent $J_{m+2}(m)$ and once more we find the higher-order coefficients increasing in magnitude with n. It may therefore be concluded that $J_{m-1}(m)$ cannot exceed $2J_m(m)$.

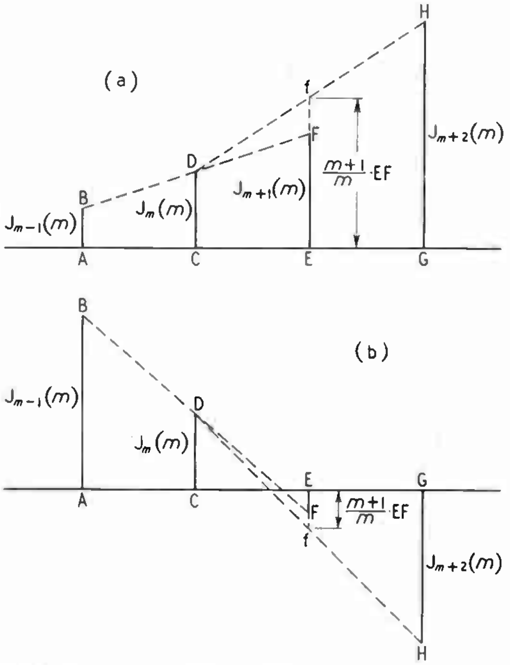


Fig. 2. *Diagram showing coefficients adjacent to* $J_m(m)$. *It is assumed that* $J_{m-1}(m) < J_m(m)$ *in* (*a*), *and* $J_{m-1}(m) > 2\,J_m(m)$ *in* (*b*).

5. Evaluation of Bessel Coefficients (Method 1)

The construction we have just discussed indicates that because $J_n(m)$ is finite for all values of n, then in the range $n > m$, $J_{n-1}(m)$ must always be greater than, and of the same sign as, $J_n(m)$. It should be noted also that when $n \geqslant m$, $J_n(m)$ cannot become zero nor can it change its sign. The first method of computation which we shall describe employs these conclusions to determine a limiting value for the ratio between two coefficents of adjacent orders. In the example selected, the argument m is a whole number, but precisely similar treatment may be employed when this is not the case. In applying this method it will be found advantageous in all instances to use the pair of functions for which n lies nearest to m.

Let us now proceed to find the values of $J_n(5)$ up to, say, the 8th order. The first objective is the determination of the value of the ratio $\rho = \dfrac{J_4(5)}{J_5(5)}$.

From equation (10) we have,

$$2J_5(5) - J_4(5) = J_6(5)$$

or, $$J_5(5) = \frac{1}{(2-\rho)}\,J_6(5)$$

Now $J_6(5)$ cannot be zero and must be less than $J_5(5)$. The required ratio thus lies in the range $2 > \rho > 1$. Similarly,

$$\frac{12}{5}\,J_6(5) - J_5(5) = J_7(5)$$

or, $$J_6(5) = \frac{(2-\rho)}{3{\cdot}8 - 2{\cdot}4\rho}\,J_7(5)$$

This relation shows that $1{\cdot}583 > \rho > 1{\cdot}285$, since the upper value makes $J_7(5)$ zero, while the lower makes $J_7(5)$ equal to $J_6(5)$. Proceeding in this way,

$$\frac{14}{5}\,J_7(5) - J_6(5) = J_8(5)$$

or $$J_7(5) = \frac{3{\cdot}8 - 2{\cdot}4\rho}{8{\cdot}64 - 5{\cdot}72\rho}\,J_8(5)$$

i.e., $1{\cdot}512 > \rho > 1{\cdot}458$

Similarly,

$$\frac{16}{5}\,J_8(5) - J_7(5) = J_9(5)$$

or $$J_8(5) = 8{\cdot}64 - 5{\cdot}72;$$

i.e., $1{\cdot}500 > \rho > 1{\cdot}494$

The number of steps necessitated in this process of narrowing the limits for ρ will depend on the accuracy which is sought. From the manner in which the limits are converging in the present example we may accept a value of 1·498 for ρ.

Using the conclusion that $J_4(5) = 1{\cdot}498J_5(5)$, we may now progressively evaluate each of the other significant coefficients in terms of $J_5(5)$.

TABLE 1

n	0	1	2	3	4	5	6	7	8
$\dfrac{J_n(5)}{J_5(5)}$	−0·6796	−1·254	0·178	1·397	1·498	1·000	0·502	0·205	0·072
$J_n(5)$ (calculated)	−0·1779	−0·3283	0·0466	0·3657	0·3922	0·2618	0·1314	0·0536	0·0188
$J_n(5)$ (from Tables)	−0·1776	−0·3276	0·0466	0·3648	0·3912	0·2611	0·1310	0·0534	0·0184

Thus,

$$\begin{aligned}J_3(5) &= \frac{8}{5}J_4(5) - J_5(5)\\ &= \left[\frac{8}{5}\times 1{\cdot}498 - 1\right]J_5(5) = 1{\cdot}397J_5(5)\\ J_2(5) &= \frac{6}{5}J_3(5) - J_4(5)\\ &= \left[\frac{6}{5}\times 1{\cdot}397 - 1{\cdot}498\right]J^{5}(5)\\ &= 0{\cdot}178J_5(5).\end{aligned}$$

The results for all coefficients up to $J_8(5)$ are shown in Table 1.

It now remains to determine the value of $J_5(5)$, for which purpose we may employ the relation,

$$1 = J_0(5) + 2[J_2(5) + J_4(5) + J_6(5) + \ldots.]$$

which follows from putting $\theta = 0$ in equation (6). The values of coefficients above the 8th order are negligibly small, so that from Table 1 we have,

$$1 = \{-0{\cdot}6796 + 2[0{\cdot}178 + 1{\cdot}498 + 0{\cdot}502 + 0{\cdot}0720]\}J_5(5)$$

Thus, $$J_5(5) = 0{\cdot}2618$$

Using this as a multiplying factor, the values of the other coefficients are readily obtained. Table 1 shows the computed values in comparison with those given in standard tables of Bessel coefficients.

6. Evaluation of Bessel Coefficients (Method 2)

The second method of computation which we shall describe seeks to determine the ratio between coefficients of adjacent orders in a somewhat different manner. From equation (10)

$$\begin{aligned}J_{n+1}(m) &= \frac{2n}{m}J_n(m) - J_{n-1}(m)\\ J_{n+2}(m) &= \frac{2(n+1)}{m}J_{n+1}(m) - J_n(m)\\ &= \left[\frac{4n(n+1)}{m^2} - 1\right]J_n(m)\\ &\quad - \frac{2(n+1)}{m}J_{n-1}(m)\end{aligned}$$

so that in the general case we may write,

$$J_\lambda(m) = \alpha_\lambda J_n(m) - \beta_\lambda J_{n-1}(m)$$

where α_λ and β_λ are numerical factors. When λ is much larger than m, $J_\lambda(m)$ becomes vanishingly small, and $\alpha_\lambda J_n(m)$ and $\beta_\lambda J_{n-1}(m)$ tend to equality. It follows that $\frac{\alpha_\lambda}{\beta_\lambda}$ approaches a constant value as λ becomes large. The constancy of this ratio becomes evident for orders of λ which are only a little greater than m. Its value, which is of course, equal to $\frac{J_{n-1}(m)}{J_n(m)}$ is therefore readily determined in any given instance.

In order to compare the two methods we shall determine the value of $J_n(5)$ by first computing the ratio $\frac{J_4(5)}{J_5(5)}$ in the following manner,

$$\begin{aligned}J_6(5) &= 2J_5(5) - J_4(5)\\ J_7(5) &= \frac{12}{5}J_6(5) - J_5(5)\\ &= \frac{12}{5}[2J_5(5) - J_4(5)] - J_5(5)\\ &= 3{\cdot}8J_5(5) - 2{\cdot}4J_4(5)\end{aligned}$$

Similar calculations give the values of α_λ and β_λ shown in Table 2.

TABLE 2

λ	α_λ	β_λ	$\frac{\alpha_\lambda}{\beta_\lambda}$
6	2	1	2·000
7	3·8	2·4	1·583
8	8·64	5·72	1·510
9	23·85	15·90	1·500
10	77·22	51·52	1·498
11	285·0	190·2	1·498

This shows the value of the ratio $\frac{J_4(5)}{J_5(5)}$ to be 1·498. Individual coefficients are now evaluated by the procedure indicated in the second part of the previous method.

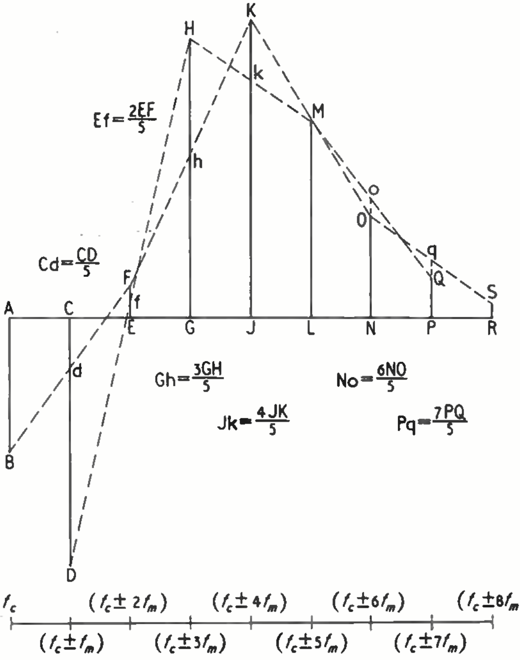


Fig. 3. *Amplitude of components of the frequency spectrum of a frequency-modulated wave* ($m_f = 5$).

The results of the analysis in the selected example are shown in graphical form in Fig. 3 which also demonstrates the simple geometrical construction linking the magnitudes of alternate coefficients. With the addition of an appropriate frequency scale, this diagram represents the amplitudes of the sinusoidal components of the spectrum of a frequency-modulated wave having a modulation index of 5. It will be seen that in this case all prominent components lie in the frequency range from $(f_c - 8f_m)$ to $(f_c + 8f_m)$.

7. Conclusion

In the preparation of this article consideration was given to the possibility of developing general solutions for both methods, instead of showing processes of computation in a particular example. The general expressions, however, rapidly become cumbersome and it is impossible to avoid inelegance in handling them. Moreover, since the purpose has been to show how Bessel coefficients enter into the analysis of a frequency-modulated wave, and to demonstrate simple methods of evaluating these apparently intractable series, the qualitative aim is perhaps better served by treating a particular instance. While this explanation has sought to bridge the gap commonly left by textbooks dealing with the subject, it may also provide a useful introduction to other aspects of the use of Bessel functions in radio engineering.

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.

Resonant Circuit with Periodically-Varying Parameters

SIR,—I have been very puzzled by the article by Messrs. Bura and Tombs, in *Wireless Engineer* for April 1952, pp. 95–100, and by the phenomenon of 'multiple-resonance' which they describe. I have come to the conclusion that they must be restricting their discussion to one particular mode of operation of the circuit, which I have listed as mode (*b*) below, and which is based on a possibly-linear, but interrupted, oscillatory regime. In the hope of clarifying the subject, I would like to give a little general discussion of the modes of operation of a tuned circuit with periodically-varying resistance (see Fig. 1), with the limitation that the frequency (ω_1) of resistance variation is very much smaller than the natural frequency of resonance (ω_0), as in the article under discussion. The opposite condition, that $\omega_1 > \omega_0$, is no doubt relevant to Part 2 of the article.

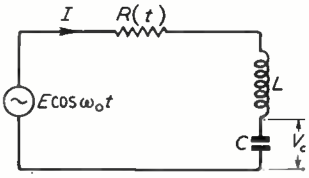


Fig. 1.

The resistance is taken to be sinusoidally modulated; i.e.,

$$R(t) = R_0 + R_1 \cos \omega_1 t$$

As I see it, there are three modes:—

Mode	$R(t)$	R_0 and R_1	Linearity
(*a*)	Always positive	R_0 positive, $R_1 < R_0$	Linear
(*b*)	Positive over most of cycle, negative over remainder	R_0 positive, $R_1 > R_0$	Linear or non-linear
(*c*)	Negative over most or all of cycle	R_0 negative	Non-linear

Mode (*a*).—This is a case of a modulator with frequency-selective terminations, and can be analyzed by steady-state theory. In discussing it, however, I would like first of all to consider the resonant circuit decoupled from the modulator $R(t)$, which therefore works into a plain resistance load, as shown in Fig. 2. This circuit has, in fact, several of the properties discussed by Bura and Tombs for mode (*b*). Here the voltage V_L across the resistance load R_L is

$$V_L = IR_L = \frac{ER_L \cos \omega_0 t}{R_L + R(t)} = \frac{ER_L \cos \omega_0 t}{R_L + R_0 + R_1 \cos \omega_1 t} \quad .. \quad .. \quad .. \quad (1)$$

Put $\dfrac{R_1}{R_L + R_0} = x$ and assume $x < 1$.

$$\text{Then } V_L = \frac{ER_L \cos \omega_0 t}{R_L + R_0} (1 + x \cos \omega_1 t)^{-1}$$
$$= \frac{ER_L \cos \omega_0 t}{R_L + R_0} \Big[\left(1 + \frac{x^2}{2} + ...\right) - \left(x + \frac{3x^3}{4} + ...\right) \cos \omega_1 t + \left(\frac{x^2}{2} + ...\right) \cos 2\omega_1 t - \left(\frac{x^3}{4} + ...\right) \cos 3\omega_1 t + ... \Big] \quad .. \quad .. \quad (2)$$

which shows quite clearly that (a) sidebands are produced at all frequencies $\omega_0 \pm n\omega_1$ although the resistance is sinusoidally modulated, and also (b) the output amplitude is increased relative to its value when the resistance is unmodulated; e.g., the output at frequency ω_0 is increased in the ratio $1 + \frac{1}{4}x^2 + \ldots$. Both these points demonstrated for Fig. 1 in mode (*b*) by Bura and Tombs thus also apply (qualitatively) in a purely resistive circuit in mode (*a*).

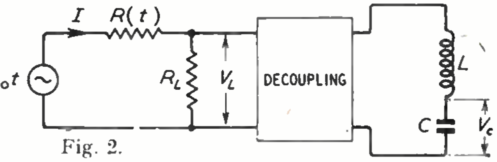


Fig. 2.

Now consider the effect of the tuned circuit in Fig. 2. If this has a high enough Q-value, it will act as a wave analyzer and give a peak of output at each frequency

$(\omega_0 \pm n\omega_1)$ in the spectrum of V_L, just as Bura and Tombs found in Fig. 1 under certain conditions and called 'multiple resonance'. But if the Q-value is too low, separate peaks cannot be found except at the frequency ω_0 if it is predominant.

Returning to the more difficult case of Fig. 1, this is an example of a modulator with frequency-selective terminations as discussed in a previous paper[1] and particularly the discussion thereon.[2] For simplicity assume that the tuned circuit presents an infinite impedance at all component frequencies in (2) except ω_0 and $\omega_0 + \omega_1$, but has an impedance Z_0 at ω_0, and Z_{01} at $\omega_0 + \omega_1$. This is not a very unpractical condition. Then if I_0 is the current at ω_0, and I_{01} that at $\omega_0 + \omega_1$, we can write down equations for the voltages at the two frequencies, noting that no other currents can exist:—

$$\text{At freq. } \omega_0,\ E = I_0(Z_0 + R_0) + \tfrac{1}{2}I_{01}R_1 \quad \ldots \quad \ldots \quad (3)$$

$$\text{At freq. } \omega_0 + \omega_1,\ 0 = \tfrac{1}{2}I_0R_1 + I_{01}(Z_{01} + R_0) \quad \ldots \quad (4)$$

Then

$$I_0 = \frac{-E(Z_{01} + R_0)}{\tfrac{1}{4}R_1^2 - (Z_0 + R_0)(Z_{01} + R_0)} \quad \ldots \quad \ldots \quad (5)$$

and

$$I_{01} = \frac{\tfrac{1}{2}ER_1}{\tfrac{1}{4}R_1^2 - (Z_0 + R_0)(Z_{01} + R_0)} \quad \ldots \quad \ldots \quad (6)$$

Here, on putting in numerical values as used by Bura and Tombs, except that R_1 is made less than R_0, no trace of 'multiple resonance' can be found. Taking $L = 0{\cdot}0851$ H, $\omega_1 = 314$ rads/sec, $\omega_0 = 2 \times 10^4$ rads/sec, $R_0 = 9$ ohms, and $R_1 = 7{\cdot}1$ ohms, and considering three tuning conditions thus:—

(1) circuit tuned to natural resonance at ω_0
(2) circuit tuned to natural resonance at $\omega_0 + \omega_1/2$
(3) circuit tuned to natural resonance at $\omega_0 + \omega_1$

then we obtain the following values for I_0 and I_{01}:—

Tuning	I_0	I_{01}
(1)	$0{\cdot}112E$	$0{\cdot}007E$
(2)	$0{\cdot}037E$	$0{\cdot}005E$
(3)	$0{\cdot}019E$	$0{\cdot}007E$

(Of course, the voltage V_r observed by Bura and Tombs is proportional to the current over such a very small frequency range).

It will be seen that neither the average nor the peak value of I_0 and I_{01} taken together is less at $\omega_0 + \frac{1}{2}\omega_1$ than at $\omega_0 + \omega_1$ or low enough at $\omega_0 + \frac{1}{2}\omega_1$ to suggest there could be a dip between that frequency and $\omega_0 + \omega_1$.

Mode (b).—Here the circuit resistance is negative for a time which may be large compared with one period of the natural resonance frequency (since it is specified that $\omega_1 \ll \omega_0$). Since only intermittent free oscillation can occur, it is possible for purely linear conditions to apply.

Bura and Tombs mention the super-regenerative circuit, and this is presumably mode (b). Oscillation occurs in the absence of an input signal (ω_0), but it is said in text-books[3] that the magnitude of the signal, when present, controls the amplitude to which oscillations build up. Bura and Tombs also refer to their system as linear, so that presumably free and forced oscillations exist together without interaction. No doubt, for this mode, Bura and Tombs' analysis is the most suitable, as it is doubtful if the equations of mode (a) apply to it.

Mode (c).—Here continuous oscillation occurs as even if $R_1 > R_0$, there is not time for the oscillation to die away during 'off' periods, and it must, therefore, be non-linear. It is possible that a form of 'multiple resonance' occurs now, due to the phenomenon of 'sideband pull-in' described and analysed graphically by Bab[4] in 1934. This effect is that a modulated oscillator will synchronize to a tone of frequency approximately equal to that of either the main oscillation *or* its sidebands, so that as the circuit is detuned, synchronization occurs, with an amplitude peak, at each sideband frequency in turn; but for such a circuit there is a threshold value for the injected voltage, below which synchronization fails at some parts of the frequency range.

It is hoped that this discussion of the circuit has helped to set out a background; but it does appear that some further elucidation of the physical performance of Bura and Tombs' system is desirable.

D. G. TUCKER.

H.M. Underwater Detection Establishment,
Portland, Dorset.
18th April 1952.

REFERENCES

[1] D. G. Tucker. "Rectifier Modulators with Frequency-Selective Terminations". *Proc. Instn elect. Engrs.* Vol. 96, Part III, p. 422, 1949.

[2] Discussion on the above paper, *ibid*, Vol. 97, Part III, p. 205, 1950.

[3] See F. E. Terman. "Radio Engineers Handbook", McGraw-Hill. 1943, p. 662.

[4] U. Bab. "Graphische Behandlung von Mitnahme-erscheinungen" (Graphical Treatment of Pull-in Phenomena), *Elektrische Nachrichten Technik*, Vol. 11, p. 187, 1934.

SIR,—In reply to Dr. Tucker's letter:

1. Dr. Tucker is right in supposing that the harmonics are generated by $R(t)$ and that they are selected by the tuned circuit (see p.99 of April issue).

2. The linearity or otherwise of a circuit is expressed by the differential equation that describes the behaviour of the circuit. The Hill equation (p.95 of April issue) is linear and appears to lead to results which agree closely with our experiments. We therefore conclude that under our conditions the linear approximation is satisfactory.

3. The steady-state solution (omitting transients since, for any positive mean resistance, the transients are bound to die away) is given in our equation (10) without any restriction on R_1 being greater or less than R_0 [modes (a) and $(b)$of the letter]. We do not consider there is any significant difference in principle between (a) and (b). The amplification in a super-regenerative circuit in its linear mode is due to the increase in the amplitude of the forced oscillations during the portion of the resistance cycle when the instantaneous circuit resistance is negative.

4. We agree with Dr. Tucker that with small amplitude variations of $R_1 (= 7{\cdot}1\Omega)$, no multiple resonance (as judged by humps on the flank of the resonance curve) occurs. Even taking greater values of $R_1 (= 64\Omega)$ as used by us experimentally, and inserting them into Dr. Tucker's equations (3)-(6) we still find no evidence of a minimum between tuning condition (1) and (3), in fact a *maximum* current I_0 and I_{01} occurs, which for the particular case of $R_1 = 56{\cdot}4\Omega$ proceeds to an infinity, clearly showing the limitations of **Dr. Tucker's** equations.

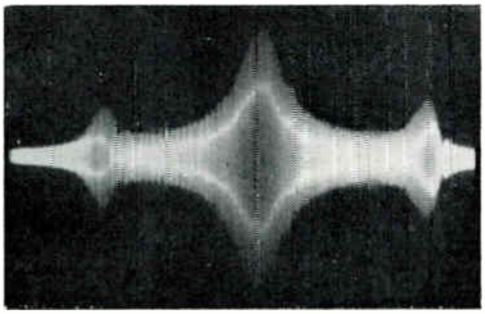
However, applying Dr. Tucker's method and including more terms (using the symbols of his letter), and solving the five simultaneous equations corresponding to the two equations (3) and (4), the following table emerges if E is put equal to unity.

Tuning	$\lvert I_{0(-2)}\rvert$	$\lvert I_{0(-1)}\rvert$	$\lvert I_0\rvert$	$\lvert I_{0(+1)}\rvert$	$\lvert I_{0(+2)}\rvert$	$\Sigma\lvert I\rvert$
(1)ω_0	0·03109	0·1043	0·20692	0·1043	0·03109	0·4777
(2)$\omega_0 + \frac{\omega_1}{2}$	0·006452	0·02697	0·07442	0·05944	0·02363	0·1909
(3)ω_1	0·003878	0·01945	0·06925	0·11303	0·06659	0·2722

Here, in the last column, is clear evidence of multiple resonance since the value for (2) is less than either (1) or

(3). Thus Dr. Tucker's method (with sufficient terms) leads to results similar to our own, the main difference being that ours is the asymptotic value of his when infinite terms are taken.

Added evidence of the reality of the phenomenon is shown in the figure—an oscillogram[1] taken to show multiple resonance.



The situation seems, therefore, to be very much simpler in principle than Dr. Tucker's letter suggests, all phenomena within the stated restrictions being accounted for by our solution of Hill's equation.

P. BURA,
D. M. TOMBS.

Imperial College of Science and Technology,
South Kensington,
London, S.W.7.
17th June 1952.

REFERENCE

[1] Bura and Tombs. '*Nature*', 16th September 1950.

The Dromgoole Effect

SIR,—We have seen with interest your Editorial in the May 1952 issue of *Wireless Engineer* on the experiments of Mr. Dromgoole.

It is well known that the permeability of ferrous materials is affected by stress, and recently we considered using this fact to examine the stress distribution in steel discs. These discs had been subjected to differential cooling to produce pre-stressing at the centre.

The method used in our experiments was dictated by the materials at hand. A number of small 'U'-shaped mu-metal stampings were wound with 200 turns of 34 s.w.g. enamelled copper wire. This coil was connected to an inductance bridge which was supplied from an oscillator at frequencies in the range 1,000 to 50,000 c/s. The stampings were placed so that the pre-stressed material closed the magnetic circuit. On moving the stampings over the surface, variations of the inductance of the coil were measured up to 3%. The arrangement appears to be more sensitive at the higher frequencies.

As a further check on the method, a piece of steel was clamped in the jaws of a large bench vice, and again the inductance of the coil varied as the grip of the vice jaws was increased.

V. C. H. BAILEY.
National Gas Turbine Establishment, R. CHAPLIN.
Leicester.
3rd July 1952.

S.I.M.A. SYMPOSIUM AND EXHIBITION

The fourth symposium and exhibition of the Electrical and Electronics Section of the Scientific Instrument Manufacturers' Association is to be held at the Examination Hall, Queen Square, London, W.C.1, from Tuesday, 2nd September, to Friday, 5th September. Entrance to the exhibition can be obtained on presentation of a trade card but, for the symposium, tickets are needed and are obtainable from the Secretary, 20 Queen Anne Street, London, W.1.

The symposium comprises the following papers:—

"Electronic Control Systems for Large Astronomical Telescopes," by G. H. Hickling.

"The Development of Resistive Elements and Wave-Guide Attenuators from a Semi-Conducting Ceramic," by J. M. Herbert.

"Electronics in Strain Measurement," by D. L. Johnston and D. W. Hobbs.

"Electronic Instruments Developed by the E.R.A. for Research in the Electrical Industry," by G. Mole.

"Electronic Measurement and Control for Industry," by J. R. Boundy.

"Electronics in Temperature Control," by D. K. Das Gupta and R. J. Russell-Bates.

"The Application of High Power Ultrasonics," by B. E. Noltingk.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for June 1952

Date 1952 June	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1029-1130 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	— 0·2	— 2	N.M.
2	+ 0·1	— 1	— 35·2
3	+ 0·1	— 3	— 34·8
4	+ 0·1	— 2	— 35·7
5	+ 0·1	— 2	— 36·1
6	+ 0·1	— 1	— 36·8
7	+ 0·2	— 2	— 36·8
8	N.M.	N.M.	N.M.
9	+ 0·2	— 2	— 38·6
10	+ 0·2	— 2	— 37·9
11	+ 0·3	— 1	— 38·4
12	+ 0·3	0	— 39·0
13	+ 0·3	— 1	— 38·9
14	+ 0·2	— 1	— 39·9
15	+ 0·3	— 1	— 39·0
16	+ 0·4	— 1	— 39·1
17	+ 0·5	0	— 40·0
18	+ 0·4	0	— 38·4
19	+ 0·5	0	— 39·7
20	+ 0·5	0	— 38·7
21	+ 0·5	— 1	— 39·1
22	+ 0·6	— 10	— 37·4
23	+ 0·6	0	— 39·2
24	+ 0·6	0	— 40·1
25	+ 0·7	0	— 38·4
26	+ 0·7	0	— 37·9
27	+ 0·8	— 2	— 37·1
28	+ 0·9	— 2	— 36·4
29	+ 0·9	N.M.	— 34·6
30**	—	— 2	—

The values were estimated on 1st July 1952.

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

N.M. = Not measured.

* = No MSF transmission at 1029 G.M.T. Results for 1429-1530 G.M.T.

** = No MSF transmission at 1029 or 1429 G.M.T.

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 the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

ACOUSTICS AND AUDIO FREQUENCIES

534.321.9 **2099**

The Measurement of Ultrasonic Attenuation in Solids by the Pulse Technique and some Results in Steel.—R. L. Roderick & J. Truell. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 267–279.)

534.756 **2100**

Signal Translation in Hearing.—L. O. Schott. (*Bell Lab. Rec.*, Jan. 1952, Vol. 30, No. 1, pp. 2–8.) Account of investigations of the process by which sound waves are converted to electrical pulses in the ear, using a circuit analogue for the hair-cell pulse generators.

534.771 **2101**

Audiometer Measurement of Hearing Loss.—F. J. Meister. (*Arch. tech. Messen*, Jan. 1952, No. 192, pp. 15–18.) Two audiometers are briefly described, one, of the heterodyne oscillator type, providing a continuous frequency range from 20 c/s to 20 kc/s, and the other, a RC generator, providing 10 frequencies between 64 and 11 584 c/s. Practical details are discussed.

534.839 : [621-1 + 629.13 **2102**

Sonic and Ultrasonic Noise in Engineering and in Aviation.—P. Bugard, M. Guennec & J. Selz. (*Ann. Télécommun.*, Jan. 1952, Vol. 7, No. 1, pp. 47–55.) Report of measurements made with a piezoelectric microphone sensitive up to 80 kc/s, (*a*) using a frequency analyser of range 1–100 kc/s and a pen recorder, (*b*) using a recording c.r.o.

534.843 : 534.7 **2103**

Room Acoustics from the Point of View of the Singer and Speaker.—R. Husson. (*Ann. Télécommun.*, Feb. 1952, Vol. 7, No. 2, pp. 58–74.) Analysis of physiological processes of articulation and the effects of room acoustics on the vocal system. A short reverberation time suitable for an audience may cause rapid fatigue in the performer.

534.846 **2104**

Acoustical Researches in the Municipal Theatre of Budapest.—T. H. Tarnóczy. (*Acta tech. Acad. Sci. hungaricae*, 1952, Vol. 2, Nos. 2–4, pp. 285–301. In English.) Report of investigations of reverberation, echo paths, and attenuation at different locations, with a view to structural modifications to improve the acoustical characteristics of the theatre.

534.85/.86 : 534.32 **2105**

Critical Study of the Concept of High Fidelity in the Response of Electroacoustic Systems.—A. Moles. (*Onde élect.*, Jan. 1952, Vol. 32, No. 298, pp. 11–25.) Three criteria of acoustic quality are examined: (*a*) the minimum change in intensity perceptible by the ear at different frequencies and sound levels; this leads to a definition of an acoustic information unit; (*b*) the inability of the ear to detect a change in quality between two positions not far apart in a good concert hall; (*c*) the maximum variation that can be introduced unnoticed by an audience in the reproduction of a piece of music heard normally just previously. Experimental investigations of (*b*) and (*c*) are described. In practice a system has 'perfect' fidelity if its response characteristic is level to within ± 2 db over the a.f. range.

621.317.7.018.78.029.4 **2106**

A Distortion Analyser for the Audio-Frequency Range.—G. Hoffmann. (*Fernmeldetech. Z.*, Jan. 1952, Vol. 5, No. 1, pp. 31–38.) An explanation of the basic principles of analysers suitable for measurement of all harmonics, combination tones, etc., within a range of about 30 c/s–30 kc/s, and an outline description of a possible type of equipment.

621.395.61/.62 **2107**

Coupled Electroacoustic Transducers.—F. A. Fischer. (*Arch. elekt. Übertragung*, Jan. 1952, Vol. 6, No. 1, pp. 35–36.) Correction to paper noted in 571 of March.

621.395.623.7 **2108**

A Step towards the High-Fidelity Reproducer: the Focusing Baffle.—P. Forestier. (*TSF et TV*, Feb. 1952, Vol. 28, No. 280, pp. 66–68.) Description of a commercially available loudspeaker unit comprising a baffle formed of part of an ellipsoid, effective for the middle and upper frequencies, together with a spherical resonator effective for the lower frequencies.

621.395.625.3 **2109**

Explanation of the Inner Mechanism of Magnetic-Tape Recording.—K. Schwarz. (*Frequenz*, Feb. 1952, Vol. 6, No. 2, pp. 37–44.) Longitudinal and transverse magnetization of a permanent-magnet material are distinguished and the frequency characteristic in each case is discussed. Transverse magnetization of the tape is predominant in normal recording techniques. Causes of nonlinear distortion in recording and reproduction are noted.

621.395.625.3 **2110**

New Magnetic-Recording Head.—M. Camras. (*J. Soc. Mot. Pict. Televis. Engrs*, Jan. 1952, Vol. 58, No. 1, pp. 61–66.) Description of a type of recording head with a third magnetic pole mounted above the air gap between the other two. It is claimed that such heads produce a more uniform field throughout the thickness of the magnetic layer, with a more rapid field decay at the trailing edge.

621.395.92 **2111**

The Medresco Hearing Aid.—C. J. Cameron & E. W. Ayers. (*P.O. elect. Engrs' J.*, Jan. 1952, Vol. 44, Part 4, pp. 153–158.) A broad outline is given of the work done by the Post Office in redesigning the hearing aid developed for the Medical Research Council (see 1245 and 1246 of 1948), incorporating a magnetic receiver and a new microphone, with modification of the amplifier circuit. A description is given of typical Medresco aids and their principal component parts. Production and performance of these aids is briefly mentioned and possible future improvements in design are noted.

AERIALS AND TRANSMISSION LINES

621.392.015.3 : 517.432 **2112**

A Particular Application of Operational Calculus to Transients on Transmission Lines.—R. Codelupi. (*Alta Frequenza*, Feb. 1952, Vol. 21, No. 1, pp. 27–38.) The quadripole equations of a uniform line with negligible attenuation and phase distortion are put into a form suitable for application of operational calculus. As examples, the expression for the far-end voltage of a line is transformed to one which shows the reflected waves, and the waveform corresponding to a unit step voltage input is determined (*a*) for the input end of an infinitely long coaxial cable, (*b*) for the output from a 300-m length of 75-Ω coaxial cable terminated by a 75-Ω resistor in parallel with a 0·01-μF capacitor. For (*b*) the results are in good agreement with measurements.

621.392.21 **2113**

Very Flexible Two-Wire Line with Rapid Radial Diminution of Field, for U.H.F. Transmission.—H. Kleinwächter & H. Weiss. (*Onde élect.*, Feb. 1952, Vol. 32, No. 299, pp. 46–50.) Investigation of a transmission line consisting of a twisted pair of very thin insulated wires, for use as a low-power feeder for cm waves. Optimum wire diameter and insulation thickness are determined, taking flexibility and attenuation into account. The radial diminution of the radiation field is calculated approximately. Results of measurements on a line consisting of 1-mm copper wires with polythene coating of outer diameter 2 mm agree well with calculations, the attenuation being of the order of 0·7 db/m. For a line using 0·05-mm wires the calculated attenuation is about 5 db/m.

621.392.26 **2114**

Compilation of the Propagation Constants of an Inhomogeneously Filled Waveguide.—L. G. Chambers. (*Brit. J. appl. Phys.*, Jan. 1952, Vol. 3, No. 1, pp. 19–21.) Application to the waveguide problem of the Rayleigh-Ritz method developed for determining the natural vibration frequencies of a mechanical system.

621.396.67 : 535.42 : 538.56 **2115**

Radiation or Diffraction Patterns Close to Receiving Antennas.—L. S. Palmer. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 289–290.) Comments on 2159 of 1951 (Andrews).

621.396.67 : 621.392 **2116**

On the Transmission Efficiency of Long Feeder Lines for Very High Frequencies.—K. Nagai, T. Omori, R. Sato & G. Sato. (*Technol. Rep. Tohoku Univ.*, 1950, Vol. 15, No. 1, pp. 57–69.) Experiment shows that the transmission efficiency of a two-wire open feeder 500 m long may be $> 50\%$ for wavelengths of 4–5 m. Wires should be 6–7 m above the ground and 30–40 cm apart; for a length of 5λ at either end spacing may be 10 cm if necessary. Distance ($2l$) between insulators should be about 30 m and such that $2l = (2n + 1)\ \lambda/4$, where n is any integer.

621.396.67.018.424 **2117**

Broad-Band Antenna Element.—M. W. Scheldorf. (*Tele-Tech*, Jan. 1952, Vol. 11, No. 1, pp. 50–51.) Short description of the development of a slot-fed cylindrical system of rods.

621.396.67.018.424 **2118**

Radiation Damping, Resistance and Directional Characteristics of Ultrawide-Band Aerials.—H. Wolter. (*Z. angew. Phys.*, Feb. 1952, Vol. 4, No. 2, pp. 60–70.) An extension of van der Pol's theory of unloaded aerials leads to a fundamental relation between characteristic impedance and radiation damping. Aerials with very low characteristic impedance, such as wide cages, double cones, dipole triangles, etc., have high radiation damping, and their feedpoint resistances and directional characteristics are practically constant for all frequencies for which the total length of the aerial is not appreciably less than $\lambda/2$. V-shaped aerials of low characteristic impedance produce a concentrated beam and have wide-band properties. For producing a highly concentrated beam, arrays of wide-band aerials with special wide-band feed may be used.

621.396.671 **2119**

Current Distribution along a Cylindrical Aerial.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 28th Jan. 1952, Vol. 234, No. 5, pp. 513–515.) The cylinder is assumed to have a finite radius and be bounded by two planes normal to the axis, about which the current distribution is symmetrical. Starting from Maxwell's equations and using methods which have been applied in the study of transients in electrical circuits, a relation is obtained for transmitting aerials which represents an infinite set of equations involving an infinite number of unknowns. Discussion of this relation shows that the distribution of current (regarded as the algebraic sum of the internal and external current sheets) is perfectly sinusoidal for the cases where the length of the aerial is $\lambda/2$ or an odd multiple of $\lambda/2$, and also for very short aerials. For receiving aerials, similar results are obtained, thus justifying the methods of calculating their impedance and radiation characteristics. In the case of a long thin wire, end effects can be neglected and the above results will apply.

621.396.677 **2120**

Superdirective Aerials.—P. Aigrain. (*Onde élect.*, Feb. 1952, Vol. 32, No. 299, pp. 51–54.) These are defined as having a main-lobe width $< \sin^{-1} \lambda/2a$, where $2a$ is the aerial length. An approximate calculation is made of the ratio of the energy stored in the aerial and its near field to that radiated in unit time, which ratio is of the nature of a Q factor. The theoretical requirements for an aerial to have a minimum Q value for a given direction are determined. Discussion of numerical results for a particular case indicates that the extreme reduction of bandwidth and radiation efficiency makes super-directive aerials impracticable.

621.396.677 : 537.226 **2121**

An Experimental Investigation of the Dielectric Rod Antenna of Circular Cross Section Excited in Rotationally Symmetrical Modes.—C. M. McKinney. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 11–13.) Report of measurements on three series of dielectric aerials excited in the TM_{01} mode at 9·275 kMc/s. Maximum attenuation of secondary lobes was obtained with rods of relatively large diameter, but with rods of small diameter sharper and deeper central nulls occurred in the radiation pattern. For uniformly tapered rods the maximum secondary-lobe attenuation and also the deepest central null were obtained with the longest rod (length 10λ). Similar results were obtained with TE_{01} excitation.

621.396.677 : 621.396.65 **2122**

Antennas for the TD-2.—A. H. Lince. (*Bell Lab. Rec.*, Feb. 1952, Vol. 30, No. 2, pp. 49–55.) Details of the construction and assembly of lens aerials for the TD-2 relay system [1109 of April (Roetken, Smith & Friis)].

621.396.677.029.64 **2123**

Obstacle-Type Artificial Dielectrics for Microwaves.—C. Süsskind. (*J. Brit. Instn Radio Engrs*, Jan. 1952, Vol. 12, No. 1, pp. 49–60. Discussion, pp. 61–62.) A survey of analytical and practical design techniques. 38 references.

621.396.677.2 + 621.396.97 **2124**

High Frequency Broadcast Transmission with Vertical Radiation.—Adorian & Dickinson. (See 2335.)

621.392.015.3 **2125**

Travelling Waves on Transmission Systems. [Book Review]—L. V. Bewley. Publishers: Chapman & Hall, 2nd edn 1951, 544 pp., 96s. (*Beama J.*, Jan. 1952, Vol. 59, No. 175, p. 11.) The first edition, published in 1933, has been used as the basis of a university course on transmission-line transients; this revised edition takes account of developments in theory and practice in relation to surge phenomena.

621.396.67 : 621.397.6 **2126**

TV Master Antenna Systems. [Book Review]—I. Kamen & R. H. Dorf. Publishers: J. F. Rider, New York, 1951, 368 pp., $5.00. (*Electronic Engng*, April 1952, Vol. 24, No. 290, p. 183.) "Recommended not only to those who are concerned with the provision and maintenance of communal aerial systems, but also to others who are interested in television reception and distribution."

621.396.67.029.64 : 621.396.932 **2127**

Centimetric Aerials for Marine Navigational Radar. [Book Notice]—Publishers: H.M. Stationery Office, London, 156 pp., 15s. (*Govt Publ.*, *Lond.*, Jan. 1952, p. 28.) "Proceedings of a conference held June 15–16, 1950, in London."

CIRCUITS AND CIRCUIT ELEMENTS

512 : 621.3.06 **2128**

The Algebra of Chains of Contacts.—H. Schwab. (*Ann. Télécommun.*, Jan. 1952, Vol. 7, No. 1, pp. 2–16.)

621.3.015.7 **2129**

Pulse Discrimination by means of Delay Elements.—H. Laett. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Jan. 1952, Vol. 30, No. 1, pp. 1–6. In German.) Discussion of the method in which the pulse train is slightly delayed and then added to the original series of pulses. The problem resolves itself into the physical realization of an open or short-circuited transmission line satisfying the requirements as regards echo time, impedance and bandwidth. The necessary design formulae are derived for several types of line.

621.3.015.7 : 621.387.422 **2130**

Ten-Channel Pulse Analyser.—M. Langevin & G. Allart. (*C. R. Acad. Sci.*, *Paris*, 28th Jan. 1952, Vol. 234, No. 5, pp. 515–518.) Circuit details and description of equipment for amplitude sorting of the pulses given by a proportional scintillation counter.

621.314.2 : 621.3.012 **2131**

New Conductance Diagram for Transformers.—H. Kafka. (*Arch. Elektrotech.*, 1952, Vol. 40, No. 4, pp. 219–230.) Analysis of the transformer by a graphical construction based on a simplified circuit diagram (1324 of 1951). A numerical example is given. A generalization of the method is illustrated.

621.314.3 **2132**

Magnetic Amplifiers and their Recent Improvements.—B. Pistoulet. (*Rev. gén. Élect.*, Jan. 1952, Vol. 61, No. 1, pp. 45–52.) Analysis of the characteristics of transductors under different operating conditions, with particular reference to the use of high-permeability materials.

621.314.3 **2133**

An Analysis of Transients in Magnetic Amplifiers.—D. W. Ver Planck, L. A. Finzi & D. C. Beaumariage. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 498–503.) Analytical expressions are derived for the envelope of the transient output current of a magnetic amplifier when the control voltage is varied. The analysis applies to transients lasting for a few cycles of the supply frequency, or longer. Experimental results are in good agreement with theory.

621.316.86.096 **2134**

Use of Thermistors as Variable R.F. Resistance Standards.—M. Soldi. (*Alta Frequenza*, Feb. 1952, Vol. 21, No. 1, pp. 3–26.) The possibility is examined of using thermistors as r.f. resistance standards, particularly for high values of resistance. Measurements show that at constant temperature the r.f. resistance of thermistors is less than the d.c. resistance, the difference (%) increasing with frequency and with resistance value, but being small enough to be neglected over a wide range of both frequency and resistance. A description of the bridge circuit used in the measurements is given.

621.318.5 : 512 **2135**

Some Results on the Application of Boole's Algebra to the Synthesis of Relay Circuits.—C. Cardot. (*Ann. Télécommun.*, Feb. 1952, Vol. 7, No. 2, pp. 75–84.)

621.319.45 : 536.48 **2136**

Electrolytic Capacitors at Low Temperatures.—C. D. Crater. (*Tele-Tech*, Jan. 1952, Vol. 11, No. 1, pp. 44–45, 72.) Analysis of tests carried out on the products of four manufacturers revealed considerable variation in effective capacitance at low temperature among units of different makes, of different voltage ratings, and among different lots from the same manufacturer. The results indicate that plain-foil types have a much greater stability at low temperature than etched-foil types. Extended storage or operation at temperatures down to −55°C produces no significant permanent changes in operating characteristics.

621.385.029.62 : 621.3.012.8 **2137**

A Systematic Method of Linear Small-Signal V.H.F. Analysis for Valve Circuits.—I. A. Harris. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, pp. 79–89.) The analysis takes account of electron inertia effects and treats the triode valve as a passive circuit element described by a set of linear equations. These express the mesh current associated with each adjacent pair of electrodes in terms of the external voltages applied between a common point and each electrode. The basic system can be applied to such problems as the calculation of (*a*) the input and output admittances of anode-, grid-, or cathode-separation triode amplifiers, (*b*) the noise factor of a single stage.

621.392 **2138**

Introduction to Formal Realizability Theory: Part 1.—B. McMillan. (*Bell Syst. tech. J.*, March 1952, Vol. 31, No. 2, pp. 217–279.) A general approach to the theory of the realizability of networks with many accessible terminals. The methods developed are applied to give a complete characterization of all finite passive networks.

621.392.012.8 : 517.562.2 **2139**

Network Representation of Transcendental Impedance Functions.—M. K. Zinn. (*Bell Syst. tech. J.*, March 1952, Vol. 31, No. 2, pp. 378–404.) The admittance or impedance of certain structures, such as a finite length of transmission line or a resonant cavity, can be represented at all frequencies by that of a network comprising lumped resistance, inductance, capacitance and conductance. In general the network contains an infinite number of branches, although a finite number may be used if only certain modes are to be represented. The procedure for the network synthesis is based on use of Mittag-Leffler's theorem, which provides a tool for breaking up a transcendental meromorphic function into an infinite series of simple fractions. The method is applied to (*a*) an open-circuited twin-wire transmission line, (*b*) a short-circuited coaxial line (or toroidal cavity with E radial), (*c*) a toroidal cavity with E axial.

621.392.4/.5 : 621.396.822 **2140**

Noise Factor of Networks.—O. E. Keall. (*Marconi Rev.*, 1st Quarter 1952, Vol. 15, No. 104, pp. 25–34.) Normal methods of circuit analysis are used in the estimation of noise factor, some of the terms being defined so as to facilitate the use of these methods. Circuits are divided into two types, depending upon whether or not a valve or other isolating device is included, and theorems appropriate to the two types are presented.

621.392.43 : 621.396.67 **2141**

Methods of Calculation relating to Inductive Aerial Couplings.—V. Familier. (*Onde élect.*, Feb. 1952, Vol. 32, No. 299, pp. 39–45.) A treatment of matching problems based on simple geometry. Similar geometrical methods have been applied by Storch (571 of 1950) to the case of capacitive coupling. In the complex plane, the point representing the input impedance of a network describes a circle when an element of the network is varied, the element being purely resistive or purely reactive. From the intersections of this circle with straight lines and circles determined by circuit parameters, optimum matching conditions for aerial and tuned secondary circuit can be found. The matching ranges of quadripoles can be determined in a similar manner.

621.392.5 **2142**

Discontinuous Low-Frequency Delay Line with Continuously Variable Delay.—J. M. L. Janssen. (*Nature, Lond.*, 26th Jan. 1952, Vol. 169, No. 4291, pp. 148–149.) The network described consists of a number of sections each comprising a clamping circuit; step variations of voltage occur at instants controlled by application of switching pulses, whose frequency determines the time delay.

621.392.5 **2143**

Rise Time of Artificial Delay Lines.—R. Génin. (*C. R. Acad. Sci., Paris*, 7th Jan. 1952, Vol. 234, No. 2, pp. 193–195.) Using analysis involving Bessel functions, an expression is derived according to which the rise time is proportional to $n^{1/3}$, where n is the number of π sections in the line. This expression is of the same form as that obtained experimentally by Elmore & Sands (2007 of 1950). Since the delay time is proportional to n, it is possible to reduce relative distortion by making n large.

621.392.5 **2144**

Ladder Development of RC Networks.—E. A. Guillemin. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 482–485.) Darlington and Cauer have described a method for the synthesis of a lossless quadripole network from a single driving-point impedance and knowledge of the zeros of transmission. The procedure can readily be extended to RL and RC networks provided the zeros of transmission are restricted to the negative real axis of the complex frequency plane. The method is illustrated by numerical calculations, starting from an assumed pair of functions.

621.392.5 : 512.831 **2145**

Notes on the Application of Matrix Calculus to Linear and Pseudolinear Feedback Systems.—J. Salmon. (*J. Phys. Radium*, Feb. 1952, Vol. 13, No. 2, Supplement, pp. 25A–28A.) Conditions for the initiation of oscillations in a linear feedback system are found and extended to certain pseudolinear systems.

621.392.5 : 621.3.015.3 **2146**

Study of Transient Processes in Linear Quadripoles.—F. Brunner. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, Jan./Feb. 1952, Vol. 6, Nos. 1/2, pp. 1–9.) In the analytical method described, a relation is derived between the steady-state and initially variable components of output voltage, and the duration of the transient is defined as the time taken for the initially variable component to fall to a given fraction of the steady-state component. The method is illustrated by application to a RC element and to a parallel-resonant circuit.

621.392.52 **2147**

Generalized Ideal Filters.—L. A. Zadeh & K. S. Miller. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 223–228.) A definition is formulated which extends the concept of ideal filter to both linear varying-parameter and nonlinear types of system; a filter is said to be ideal if it can extract a signal from its combination with another signal, even when the two frequency bands overlap. The basic properties of ideal filters are investigated using function-space techniques.

621.392.52 **2148**

The Double-T RC Filter.—W. Schmidt. (*Elektrotech. Z.*, 15th Jan. 1952, Vol. 73, No. 2, pp. 35–38.) The action of a high-pass and a low-pass filter in parallel for suppression of a single frequency is analysed and design procedure indicated. When the filter is used in the feedback network of an amplifier, the circuit may operate either as an oscillator or as a tuned amplifier with prescribed bandwidth.

621.392.52.029.64 : 621.396.611.4 **2149**

Cavity Band-Pass Filters for Centimetre Waves.—H. Döring & W. Klein. (*Arch. elekt. Übertragung*, Feb. & March 1952, Vol. 6, Nos. 2 & 3, pp. 47–57 &

119–125.) A theoretical treatment of filters comprising a number of cavity resonators in the form of flat cylindrical boxes coupled by windows in the common walls. The alteration of the circuit parameters of the end cavities due to the coupling elements causes a mismatch within the filter. The adjustment of this by means of the coupling reactances affords a useful means of obtaining a required transmission characteristic; e.g., in a 4-element filter the group delay can be made practically constant over a large part of the pass band. The design procedure and method of measurement of the attenuation and phase constant of such filters are described.

621.392.54 + 621.392.26.072.31 **2150**

Application of Multi-Hole Coupling to the Design of a Variable and Calibrated Waveguide Attenuator and Impedance.—W. J. van de Lindt. (*Philips Res. Rep.*, Feb. 1952, Vol. 7, No. 1, pp. 28–35.) A discussion of the characteristics of two parallel waveguides mutually coupled by *n* equidistant identical directional elements, with a description of the application of such a system to the design of a calibrated variable attenuator, and a calibrated variable impedance capable of changing independently the amplitude and the phase of the reflection coefficient.

621.392.6 **2151**

Generalized Network Theory.—U. Kirschner. (*Arch. elekt. Übertragung*, Feb. 1952, Vol. 6, No. 2, pp. 86–87.) Correction to paper abstracted in 837 of 1951.

621.395.665.1 : 534.86 **2152**

New Principle for Electronic Volume Compression.—H. E. Haynes. (*J. Soc. Mot. Pict. Televis. Engrs*, Feb. 1952, Vol. 58, No. 2, pp. 137–144.) The principle is to modulate the signal with h.f. rectangular pulses of variable duty factor (*k*). Unwanted modulation products are filtered out, leaving the desired signal with amplitude multiplied by *k*. The value of *k* is varied in accordance with an appropriate control voltage. The circuit described incorporates a 45-kc/s pulse generator keying a push-pull amplifier. Advantages of the system are extremely low audio 'thump', very fast action if required, low distortion, and use of components and valves not specially selected. Performance figures are given.

621.396.6 : 061.4 **2153**

R.E.C.M.F. Exhibition Preview.—(*Electronic Engng*, April 1952, Vol. 24, No. 290, pp. 178–181.) Short descriptions of selected exhibits at the Radio and Electronics Component Manufacturers Federation exhibition, London, April 1952. See also *Wireless World*, May 1952, Vol. 58, No. 5, pp. 179–182.

621.396.6 : 061.4 **2154**

The National Components Exhibition.—H. Gilloux. (*Radio franç.*, Feb. 1952, No. 2, pp. 20–24.) Review of the exhibition in Paris, February 1952, and description of certain exhibits. For longer lists of items, including acoustic equipment, measurement sets and valves, see *Toute la Radio*, March/April 1952, Vol. 19, No. 164, pp. 115–122; *TSF et TV*, March & April 1952, Vol. 28, Nos. 281 & 282, pp. 85–88 & 139–144; *Radio prof., Paris, l'Exportation Électricité-Radio franç., Supplement*, March 1952, 16 pp.

621.396.611.018.3 **2155**

Subharmonic Oscillations in Electric Circuits containing Iron-Core Reactors.—J. P. Schouten & H. J. Heijn. (*Appl. sci. Res.*, 1952, Vol. B2, No. 4, pp. 301–319.) Investigation of the flux variation in an iron-cored reactor connected in series with an inductor and capacitor and fed by a sinusoidal e.m.f. In the theoretical treatment the flux/current curve is represented approximately by three straight lines of different slope, and the associated differential equation is solved graphically, with particular reference to the third-order subharmonic. Oscillograms reproduced confirm the theory.

621.396.611.21 : 534.133 **2156**

Forced Thickness-Shear and Flexural Vibrations of Piezoelectric Crystal Plates.—R. D. Mindlin. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 83–88.) An approximate theory is presented which includes the interaction between the elastic and electric fields. Computed frequencies for rectangular AT-cut quartz plates are compared with measurements by Sykes, and formulae are derived relating resonance frequencies to dimensions, elastic and electric constants, and orientation of cut.

621.396.611.3.011.21 **2157**

Input-Admittance Characteristics of a Tuned Coupled Circuit.—R. A. Martin & R. D. Teasdale. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, p. 459.) Correction to paper abstracted in 1241 of May.

621.396.611.3.029.64 **2158**

The Application of Window Coupling at Centimetre Wavelengths.—H. Döring & W. Klein. (*Elektrotech. Z.*, 1st Jan. 1952, Vol. 73, No. 1, pp. 5–9.) Basic principles are briefly described and applications in directional couplers, coupled cavity resonators in klystrons, and multistage cavity filters, are considered.

621.396.611.4 **2159**

The Loop-Excited Cavity Resonator comprising Two Confocal Paraboloids of Revolution.—H. Buchholz. (*Arch. elekt. Übertragung*, Jan. & Feb. 1952, Vol. 6, Nos. 1 & 2, pp. 6–16 & 67–72.) A cavity having the shape of a double-convex lens is formed from two paraboloids and is excited by a coaxial circular loop. Analysis of the equations for the field leads to a complete solution in the form of two loop integrals. On developing into an infinite series, this is found to correspond to the superposition of an infinite number of wave trains whose basic form is that of the wave in an infinitely long parabolic horn. The solution has poles for an infinite series of discrete values of the wave number, corresponding to the natural resonance frequencies; calculation of these frequencies is easy for the case of equal paraboloids. The magnetic-field lines and Q are investigated for near-resonance conditions. Numerical values are tabulated for a higher transcendental function used in the solution.

621.396.611.4.029.64 **2160**

Electromagnetic Resonant Behavior of a Confocal Spheroidal Cavity System in the Microwave Region.—J. C. Simons & J. C. Slater. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 29–30.) Discussion of the resonance of a small spheroidal object in a large spheroidal cavity, approximating to a needle-like aerial in a large cavity. If the needle is thin enough, the magnetic field on its surface is greatly enhanced when the cavity is tuned to a resonance frequency determined by the needle. This can be applied to measurement of the surface impedance of the needle material.

621.396.615 **2161**

The Build-Up Process in Oscillators.—W. Herzog. (*Arch. elekt. Übertragung*, Feb. 1952, Vol. 6, No. 2, pp. 58–66.) Assuming a cubic form of triode characteristic, the types of grid-voltage and anode-current swing for class A, B and C operation are explained and the mean anode currents determined. The effect of feedback is considered. An approximate theory of the Meissner oscillator is developed and the build-up period and the 'inertia' of the circuit are determined. Nonlinear feedback is also considered.

621.396.615 : 621.396.822 **2162**

The Effect of Background Noise on the Amplitude of [valve-] Maintained Oscillators.—A. Blaquière. (*C. R. Acad. Sci., Paris*, 11th Feb. 1952, Vol. 234, No. 7, pp. 710–712.) The noise is treated as a succession of impulses and each pulse is regarded as the sum of two components, one producing, to the first order, a change of phase without change of amplitude, the other a change of amplitude but no phase shift. Second-order analysis shows that the noise power is equally divided between a periodic component and a continuous spectrum. Formulae are derived which permit comparison between the noise power of a valve oscillator and that of a passive circuit.

621.396.615 : 621.396.822 : 529.786 **2163**

The Effect of Background Noise on the Frequency of Valve Oscillators. Ultimate Accuracy of Electronic Clocks.—A. Blaquière. (*C. R. Acad. Sci., Paris*, 21st Jan. 1952, Vol. 234, No. 4, pp. 419–421.) An analysis is made of the disturbing effect of a single noise pulse on the signal generated by an amplitude-stabilized oscillator, using a method described by Rice (2219 of 1948); both the amplitude and the phase of the signal are affected. The mean square of the phase shift due to noise increases linearly with time; an expression is derived for the mean square of the error in time measurement when the oscillator is used as a clock.

621.396.615.17 **2164**

An Amplitude-Comparator Multivibrator.—S. Fedida. (*Marconi Rev.*, 1st Quarter 1952, Vol. 15, No. 104, pp. 35–43.) Description of a method of rendering the output-pulse waveform, amplitude, delay, etc., and the flip-flop recovery time, independent of the amplitude of the input pulse. This is effected by suitable connection of a double diode in a subsidiary feedback loop in the flip-flop circuit.

621.396.615.17 **2165**

Triangular-Waveform (Sawtooth) Generator.—R. Peretz. (*HF, Brussels*, 1952, Vol. 2, No. 1, pp. 16–24.) A mathematical analysis is made of a circuit in which the voltage across a capacitor in the cathode circuit of a triode is applied through a directly coupled amplifier to the grid of the valve. From a pulsed input across the capacitor, a waveform of any positive or negative exponential type can be obtained by variation of the amount of feedback used. Application of the circuit in a sawtooth generator giving frequencies from 0·001 to 1 000 c/s is described and performance characteristics are shown.

621.396.619.2 + 621.396.622.6/.7] : 621.395.44 **2166**

Frequency Converters as Quasilinear Quadripoles.—W. Klein. (*Arch. elekt. Übertragung*, Jan. 1952, Vol. 6, No. 1, pp. 29–35.) Theory is developed for the various modulator and demodulator circuits used in s.s.b. and d.s.b. carrier-wave technique. A basic feature, whose introduction enables the superposition principle to be used, is the quasilinear circuit, an idealized equivalent for the actual frequency-converter circuit, in which two quadripoles are in effect switched alternately into use. The corresponding modulation function is a square wave. As a particular case, the switching may be simply a reversal of a single circuit. Quasilinear quadripoles incorporating modulated rectifiers are described; switching is performed by the periodic polarity reversals of the carrier voltage. Conditions to be satisfied by carrier waveform and rectifier characteristics are discussed. Calculations are made for various known modulator circuits.

621.396.645 **2167**

Distributed Amplification.—A. Cormack. (*Electronic Engng*, April 1952, Vol. 24, No. 290, pp. 144–147.) Basic principles are outlined and design details are given of two amplifiers with flat response curves from l.f. to 170 Mc/s. The first, a single-stage amplifier, has a gain of 18 db; the other has two stages in cascade, each with four valves, and has a gain of 28 db. Several methods of obtaining an output impedance of 75Ω in the final stage are considered briefly.

621.396.645 **2168**

The Treatment of Amplifier Circuits by means of V_a/V_g Characteristics.—A. Simon. (*Fernmeldetech. Z.*, Jan. 1952, Vol. 5, No. 1, pp. 11–16.) Direct determination of V_a/V_g characteristics (curves of constant anode and grid currents as dependent on anode and grid voltages) is difficult; hence they are usually derived from V_a/i_a or V_g/i_a characteristics. Particular features of these characteristics are noted and working characteristics are given for class A, B and C amplifiers, circuit parameters being tabulated for specified anode voltages and output powers, and control power estimated. Data are also tabulated for a class B amplifier with low-level modulation of an input stage.

621.396.645 **2169**

Some Rules for the Construction of I.F. Amplifiers.—W. Hasselbeck. (*Funk u. Ton*, Jan. 1952, Vol. 6, No. 1, pp. 1–7.) Review of the principal conditions which must be fulfilled for satisfactory operation.

621.396.645 : 621.392.52 **2170**

New Method of Calculating High-Frequency Filters with Tchebycheff Type of Amplification.—H. Edelmann. (*Arch. elekt. Übertragung*, Feb. 1952, Vol. 6, No. 2, p. 87.) Correction to paper abstracted in 88 of January.

621.396.645.029.3 **2171**

Equipment for Acoustic Measurements: Part 5—A Portable 7·5-W Loudspeaker Amplifier.—D. E. L. Shorter & W. Wharton. (*Electronic Engng*, Jan. 1952, Vol. 24, No. 287, pp. 7–9.) Description of a power amplifier used to drive the loudspeaker used in tests of room acoustics. Part 4: 1817 of July.

621.396.645.36 **2172**

The Cathamplifier.—C. A. Parry. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 460–465.) Reprint. See 78 of 1951.

621.396.645.37 **2173**

Complex Feedback.—W. Oesterlin. (*Arch. tech. Messen*, Feb. 1952, No. 193, pp. 39–42.) Investigation of the possibility of compensating both the phase and amount of amplification of a pentode by means of frequency-dependent complex feedback. Application of this principle in a pentode amplifier resulted in uniform amplification up to 12 kc/s, with negligible phase change.

621.396.822 : [621.315.5 + 621.385.2 **2174**

Thermal and Shot Fluctuations in Electrical Conductors and Vacuum Tubes.—S. S. Solomon. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 109–112.) A new derivation of Nyquist's equation relative to the amount of the thermal fluctuations generated in an electrical conductor, together with a generalization to include any arbitrary impedance function. This shows that the original Nyquist equation is valid only for physically realizable impedances of the minimum-reactance type. A short derivation of the shot-noise formula for temperature-limited diodes is also presented.

621.397.645 : 621.385.4 **2175**

Coaxial Tetrode as a TV Amplifier at V.H.F. and U.H.F.—D. H. Preist. (*Tele-Tech*, Jan. 1952, Vol. 11, No. 1, pp. 52–53 . . 88.) Three alternative arrangements for use with a coaxial tetrode are considered and a detailed discussion is given of a power amplifier using an Eimac type-4X150G valve. The network connecting the screen grid, control grid and cathode is basically a folded coaxial line, connected between control grid and screen grid at one end and between control grid and cathode at the other. At the point of folding there is a variable series inductance, provided by the stub with its adjustable short-circuiting bar, plus two shunt capacitors which are not required in all cases. The output circuit is a conventional two-section band-pass filter; the drive is applied via a loop between control grid and screen grid. Under class-B linear conditions this amplifier has a bandwidth (at – 3 db) of 5 Mc/s, peak power 107 W at 815 Mc/s and 220 W at 500 Mc/s, and power gain of 8–10. A cross-section is shown through a similar amplifier using an Eimac type-4W20000A tetrode with a water-cooled anode capable of dissipating 20 kW.

621.397.645.018.424 **2176**

Wide-Band Amplifiers with Stagger-Tuned Circuits.—J. de Vos. (*Funk u. Ton*, Feb. 1952, Vol. 6, No. 2, pp. 69–74.) Discussion of the operation of i.f. wide-band amplifiers such as are used in television receivers. Resonance frequencies for the different circuits are determined which give an optimum shape to the transmission curve.

621.3.015.3 : 517.432.1 **2177**

Transients in Electric Circuits, using the Heaviside Operational Calculus. [Book Review]—W. B. Coulthard. Publishers: Pitman & Sons, London, 2nd edn, 32s. 6d. (*Engineering, Lond.*, 18th Jan. 1952, Vol. 173, No. 4486, pp. 67–68.) "For this second edition the opportunity has been taken to revise the whole text . . . The wide range and representative character of the problems dealt with should commend the book to all electrical engineers."

GENERAL PHYSICS

531 : 537].001.362 **2178**

Analogies between Mechanical and Electrical Magnitudes.—W. Reichardt. (*Frequenz*, Jan.–March 1952, Vol. 6, Nos. 1–3, pp. 25–29, 50–55 & 72–87.)

534.22 **2179**

The Concept of Group Velocity.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 4th Feb. 1952, Vol. 234, No. 6, pp. 599–602.) Analysis justifying accepted ideas on the subject.

535.22 + 621.396.11 **2180**

The Velocity of Light.—L. Essen. (*Sci. Progr.*, Jan. 1952, Vol. 40, No. 157, pp. 54–70.) Review of the various methods that have been used to determine the velocity, and analysis of the results obtained.

535.34 : 621.315.61 **2181**

The Structure of the Long Wave Absorption Edge of Insulating Crystals.—I. C. Cheeseman. (*Proc. phys. Soc.*, 1st Jan. 1952, Vol. 65, No. 385A, pp. 25–32.) Theoretical study of a process by which light can be absorbed in insulators at frequencies below that corresponding to the energy gap. Theoretical and experimental results are in good agreement for CdS.

535.42 **2182**

A Rigorous Formulation of the Classical Diffraction Problem.—H. Hönl. (*Z. Phys.*, 19th Feb. 1952, Vol. 131, No. 3, pp. 290–304.) The methods of Sommerfeld, Schwarzschild, and Levine & Schwinger are briefly reviewed and a treatment by means of a Fourier representation of the wave function is given; this leads to two simultaneous integral equations for slits of arbitrary shape in the plane screen. Differences from Kirchhoff's theory are particularly considered.

535.42 **2183**

The Diffraction of Electromagnetic Waves at a Slit: Part 1.—E. Groschwitz & H. Hönl. (*Z. Phys.*, 19th Feb. 1952, Vol. 131, No. 3, pp. 305–319.) Application of the general theory [2182 above (Hönl)] to a straight slit, assuming the wave function to be zero at the slit boundary.

535.42 : 538.56 **2184**

Diffraction of Electromagnetic Waves by Apertures in Plane Conducting Screens.—J. P. Vasseur. (*Onde élect.*, Jan.–March 1952, Vol. 32, Nos. 298–300, pp. 3–10, 55–71 & 97–112.) Classical methods of direct integration of Maxwell's equations are reviewed and a detailed study is made of Kottler's formulae, showing in what respects they are incorrect. A system of magnetic dipoles distributed over the plane of the aperture gives a diffraction field which satisfies all the boundary conditions. These dipoles are determined by a system of two integro-differential equations more complete than the analogous equations of Copson. Reciprocally, the diffraction field can be produced by a system of electric dipoles distributed over the surface of the screen; these also are determined by a system of two integro-differential equations. These systems of equations enable a rigorous statement of Huyghens' principle under several equivalent forms, whose comparison leads to the exact expression of Babinet's principle, which can also be established directly. In the cases of diffraction by a small circular hole and by a half plane, this method of treatment gives results found in other ways by Bethe and Sommerfeld. It appears impossible to solve explicitly the integro-differential equations concerned, even in simple cases, but when the aperture is large enough (several times the wavelength) the calculations are simplified and lead to formulae analogous to those of Kottler. These formulae have many reciprocal expressions, the integration extending over the aperture or over the metal part of the screen. Experimental results for the diffraction of 3-cm e.m. waves are in satisfactory agreement with the simplified formulae proposed. 242 references.

535.42 : 538.56 **2185**

Diffraction by a Wave-Guide of Finite Length.—D. S. Jones. (*Proc. Camb. phil. Soc.*, Jan. 1952, Vol. 48, Part 1, pp. 118–134.) Starting from the electric intensities on the planes containing the walls, integral equations are derived which, after application of the Laplace transform, can be solved by successive substitutions. The series thus obtained is too complex for practical purposes, and an approximate solution is found for the case of waveguide length large compared with the wavelength.

535.42 : 538.56 **2186**

Diffraction Measurements at 1.25 Centimeters.—R. D. Kodis. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 249–255.) Brass and polystyrene cylinders with diameters comparable with λ were used as diffracting objects. Radiation source, obstacle and detector were mounted above a 4-ft × 6-ft horizontal conducting sheet, the auxiliary apparatus being located below it. Both the phase and amplitude of the electric field near the obstacle were measured; the results are compared with values calculated from theory for the conducting cylinders. The technique was also applied to investigate diffraction by an edge.

535.42 : 538.56 : 621.396.67 **2187**

Radiation or Diffraction Patterns Close to Receiving Antennas.—L. S. Palmer. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 289–290.) Comments on 2159 of 1951 (Andrews).

537.1 : 530.12 **2188**

Classic Theory of the Point Charge.—B. Jouvet. (*C. R. Acad. Sci., Paris*, 11th Feb. 1952, Vol. 234, No. 7, pp. 712–714.) Study and geometrical representation of the relativistic and electromagnetic invariances of the classic theory of the point charge.

537.213 : 537.562 **2189**

Cell constituted by Gas Ionized at High Frequency.—M. Chenot. (*C. R. Acad. Sci., Paris*, 4th Feb. 1952, Vol. 234, No. 6, pp. 608–610.) A discharge tube, excited by metre waves applied to external electrodes, can furnish a constant difference of potential between two electrodes in contact with the ionized gas within the tube, if these electrodes are arranged asymmetrically so as to produce a deformation of the luminous discharge column, such as a large dark space near one of the external electrodes. This effect is discussed.

537.311.1 **2190**

The Mean Free Path of Electrons in Metals.—E. H. Sondheimer. (*Advances in Physics*, Jan. 1952, Vol. 1, No. 1, pp. 1–42.) A survey of the theory of electrical conduction in metals (based on Sommerfeld's quantum-mechanical treatment) with reference to size effects in which the mean free path is comparable with some significant linear dimension. The evaluation of the mean free path (independently of electron density) from measurements of the resistivity of thin films or wires is discussed. Study of the influence of a magnetic field on the resistivity of thin specimens enables the momentum of electrons at the surface of the Fermi distribution to be deduced. The anomalous skin effect enables the mean free path to be compared experimentally with the penetration depth of h.f. electric fields.

537.311.31 **2191**

Metallic Conduction—The Internal Size Effect.—D. K. C. MacDonald. (*Phil. Mag.*, Jan. 1952, Vol. 43, No. 336, pp. 124–125.) Addendum to 649 of March.

537.311.31 : 539.23 **2192**

Law of Variation of the Resistance of Very Thin Metal Films as a Function of the Applied Potential.—N. Mostovetch, B. Vodar & T. Duhautois. (*C. R. Acad. Sci., Paris*, 14th Jan. 1952, Vol. 234, No. 3, pp. 305–308.) The increase of conductivity observed with high field strengths is attributed to a lowering of the potential barrier between the metal grains, due to the Schottky effect.

537.311.33 **2193**

Theory of Conductivity of Semiconductors.—G. Jaffé. (*Phys. Rev.*, 15th Jan. 1952, Vol. 85, No. 2, pp. 354–363.) The author's earlier theory (1933 Abstracts, p. 287) for polarizable media is extended to include both the electronic and ionic components of conduction; the latter component is found to cause significant departures from the Mott-Schottky theory of rectification. The a.c. admittance is derived and the frequency dependence of the susceptance and conductance is shown to be markedly influenced by the ionic component at low frequencies. Comparison of the theory with measurements on Se disks shows good agreement for the frequency range 2–200 kc/s if two species of ion of different mobilities be assumed.

537.311.33 : 537.311.4 **2194**

The Relation between Contact Resistance and Contact Potential Difference.—M. A. Krivoglaz & K. B. Tolpygo. (*Zh. tekh. Fiz.*, April 1951, Vol. 21, No. 4, pp. 417–426.) The tunnel effect through the potential barrier formed by the curvature of the conduction band of a semiconductor on which a metal electrode has been deposited is calculated. The influence of this effect on the contact resistance is discussed for various forms of the potential barrier.

537.525.72 : 538.63 **2195**

Interaction of Travelling Magnetic Fields with Ionized Gases.—P. C. Thonemann, W. T. Cowhig & P. A. Davenport. (*Nature, Lond.*, 5th Jan. 1952, Vol. 169, No. 4288, pp. 34–35.) Experiments are described in which the magnetic field associated with a r.f. current in a helix produces an electrodeless d.c. discharge of the order of amperes in an ionized gas in a tube surrounded by the helix. Both straight and toroidal tubes were investigated. The magnitude of the d.c. was not sensitive to either phase velocity or frequency, but was highly sensitive to gas pressure. Evidence was also obtained for the amplification of a r.f. signal injected into a line enclosing the plasma of a d.c. arc.

537.533 : 546.655.4-31 **2196**

Some Results on the Optical Emissivity and Thermionic Emission of Ceria.—R. Uzan. (*Le Vide*, Jan. 1952, Vol. 7, No. 37, pp. 1139–1140.) The thermionic emission follows Richardson's law, with a mean value of ϕ of 2·7 eV for a coating thickness of 60 μ. Values of A lie between 0·005 and 10 A/cm^2.

537.533.8 **2197**

On the Theory of Secondary-Electron Emission.—J. L. H. Jonker. (*Philips Res. Rep.*, Feb. 1952, Vol. 7, No. 1, pp. 1–20.) Starting from (*a*) Whiddington's law concerning the velocity reduction of electrons penetrating into a solid substance, (*b*) the experimental law of absorption, (*c*) the assumption that the distribution of the secondary electrons within matter is isotropic, the dependence of the properties of secondary electrons on various parameters is calculated and found in good agreement with experimental results.

537.533.8 **2198**

Secondary Emission from Composite Surfaces.—H. Jacobs, J. Martin & F. Brand. (*Phys. Rev.*, 1st Feb. 1952, Vol. 85, No. 3, pp. 441–447.) Investigation of various compounds indicates that each has its own threshold energy below which primary electrons do not yield true secondary electrons. It is concluded that secondary electrons originate from the filled band of a compound rather than from electron traps.

538.11 : 538.124 **2199**

The Lowest Energy State of a Linear Antiferromagnetic Chain.—P. W. Kasteleijn. (*Physica*, Feb. 1952, Vol. 18, No. 3, pp. 104–113.)

538.11 : 539.132 **2200**

On the Quantum Theory of Antiferromagnetism.—H. A. Kramers. (*Physica*, Feb. 1952, Vol. 18, No. 2, pp. 101–103.)

538.3 **2201**

Fundamentals of a New Theory of Electromagnetism.—B. Jouvet. (*C. R. Acad. Sci., Paris*, 18th Feb. 1952, Vol. 234, No. 8, pp. 819–822.) The laws of composition of fields are examined on the hypothesis that there is an upper limit to the value of field strength. A fundamental invariant is deduced.

538.3 **2202**
Dirac's New Electrodynamics.—K. J. Le Couteur. (*Nature, Lond.*, 26th Jan. 1952, Vol. 169, No. 4291, pp. 146–147.) Discussion on 1574 of June.

538.3 : 535.13 **2203**
Is there an Aether?—H. Bondi & T. Gold: P. A. M. Dirac. (*Nature, Lond.*, 26th Jan. 1952, Vol. 169, No. 4291, p. 146.) Comment on 1573 of June and author's reply.

538.521 **2204**
Electromagnetic Induction and Magnetoelectric Induction.—G. Vallauri. (*Alta Frequenza*, Dec. 1951, Vol. 20, No. 6, pp. 227–246.) The two laws of induction are compared and experiments are described which prove the law of magnetoelectric induction directly.

538.56 : 537.533 : 523.72 **2205**
Condition for Radiation from a Solar Plasma.—J. Feinstein. (*Phys. Rev.*, 1st Jan. 1952, Vol. 85, No. 1, pp. 145–146.) Theory developed by Bailey (105 of January) is critically discussed.

538.566 **2206**
The Interaction of Electromagnetic Waves within Matter.—R. Lucas. (*C. R. Acad. Sci., Paris*, 7th Jan. 1952, Vol. 234, No. 2, pp. 191–193.) Relativistic kinematic theory is used to examine the conditions necessary for interaction. The relative directions of propagation and the new frequencies resulting from interaction are determined.

538.569.4.029.64 **2207**
Beam System for Reduction of Doppler Broadening of a Microwave Absorption Line.—H. R. Johnson & M. W. P. Strandberg. (*Phys. Rev.*, 1st Feb. 1952, Vol. 85, No. 3, pp. 503–504.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 551.510.535 : 621.396.9 **2208**
Theory of Radio Reflections from Meteor Trails: Part 1.—T. R. Kaiser & R. L. Closs. (*Phil. Mag.*, Jan. 1952, Vol. 43, No. 336, pp. 1–32.) A comprehensive treatment of the reflection of e.m. waves from an ionized column with cylindrical symmetry, which diffuses radially outwards. Different distributions of electron density n are considered and the Gaussian distribution $n = n_o \exp[-(r/r_o)^2]$ is discussed in detail, r being the radius of the ionized column. The theory predicts two qualitatively different types of meteor-trail echo, depending on the magnitude of the electron line density α. With $\alpha \gg 10^{12}$ per cm, the column is expected to reflect in a manner similar to that of a metal cylinder. The reflection coefficients for the incident electric vector E parallel and perpendicular respectively to the axis of the column, are equal and substantially constant with increasing radius, except initially. These predictions are in agreement with experimental data on long-duration echoes. With $\alpha < 10^{12}$ per cm, for E parallel to the axis the echo amplitude decays exponentially with increasing radius, while for E perpendicular to the axis the amplitude may pass through a resonance value before decaying exponentially. The short-duration echoes observed in most cases are of these types.

523.7 **2209**
Solar Observations.—A. Behr & H. Siedentopf. (*Naturwissenschaften*, Jan. 1952, Vol. 39, No. 2, pp. 28–38.) Detailed report of methods of observation of sunspot activity, solar flares, eruptions and corona effects and of results obtained at many different stations over a number of years.

523.7 : 538.122 **2210**
Measurements of the Sun's General Magnetic Field.—G. Thiessen. (*Nature, Lond.*, 26th Jan. 1952, Vol. 169, No. 4291, p. 147.) Data are presented in support of the conclusions stated in 1141 of 1951.

523.72 : 538.56 : 537.533 **2211**
Condition for Radiation from a Solar Plasma.—Feinstein. (See 2205.)

523.72 : 621.396.822 **2212**
Excess Radio Noise from Solar Flares and Sunspots.—R. Q. Twiss. (*Nature, Lond.*, 2nd Feb. 1952, Vol. 169, No. 4292, pp. 185–186.) Theoretical discussion of the noise generated in and escaping from an ionized atmosphere, with particular reference to plasma oscillations and to the amplification of transverse field waves under suitable boundary conditions.

523.72 : 621.396.822] : 550.385 **2213**
Possible Identification of a Solar M-Region with a Coronal Region of Intense Radio Emission.—A. Maxwell. (*Observatory*, Feb. 1952, Vol. 72, No. 866, pp. 22–26.) On 14th June 1950 a sunspot group associated with unusually intense metre-wave radio emission crossed the sun's central meridian. For the next six months there was, at 27-day intervals, a sequence of moderate geomagnetic storms of the type known to follow the formation of an M-region. Since metre-wave solar radiation originates in the corona, the correlation between M-regions and the corona is close. The connection between aurorae, M-regions and geomagnetic storms is shown by the occurrence of radar auroral echoes at the time of the storms or 14 days out of phase with them.

523.8 : 621.396.822 **2214**
Radio Stars or Radio Nebulae.—R. Bracewell. (*Observatory*, Feb. 1952, Vol. 72, No. 866, pp. 27–29.) Brief consideration of available evidence.

523.852.3 : 621.396.822 **2215**
Extra-Galactic Radio-Frequency Radiation.—R. H. Brown & C. Hazard. (*Phil. Mag.*, Feb. 1942, Vol. 43, No. 337, pp. 137–152.) Description of measurements on λ 1·89 m using a paraboloid aerial system 218 ft in diameter giving a beam width of 2°. Three bright nebulae have been examined; three radio sources are associated with them. The radio intensity to be expected from eight of the major clusters has been calculated; two of the clusters have been surveyed and identified with radio sources. There appears to be a relation between the apparent photographic magnitude and apparent radio magnitude of nebulae.

550.384 : 551.510.535 **2216**
Interpretation of the Eleven-Yearly Variation of the Horizontal Component of the Earth's Magnetic Field.—P. Bernard. (*C. R. Acad. Sci., Paris*, 18th Feb. 1952, Vol. 234, No. 8, pp. 866–868.) The general mean of observations made at 37 stations shows a minimum of the horizontal component 1·1 years after the maximum of sunspot activity. The nature and height of ionospheric currents capable of giving rise to these variations are discussed.

551.510.535 **2217**
Physics of the Ionosphere.—K. Rawer. (*Phys. Blätter*, 1952, Vol. 8, No. 1, pp. 15–23.) A review of present-day theory.

551.510.535 : 621.396.11.029.56 **2218**

Reflection of Short Waves at Heights less than 100 km.—Dieminger & Hoffmann-Heyden. (See 2317.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 **2219**

Pictorial Radio.—C. D. Tuska. (*J. Franklin Inst.*, Jan. & Feb. 1952, Vol. 253, Nos. 1 & 2, pp. 1–20 & 95–124.) Defined as 'multi-coordinate or graphic indicating radio systems for obtaining bi- or tri-dimensional information transmitted or reflected from a plurality of geometrically related points'. The development is traced from the time of Hertz to the end of the second world war. 37 references, including many patents.

621.396.9 **2220**

Development of an Experimental Electromagnetic Detector.—J. Moline. (*Radio franç.*, Feb. & March 1952, Nos. 2 & 3, pp. 1–5 & 11–15.) Description of radar equipment designed primarily for instruction in the principles of the art. Operating frequency is 145 Mc/s ($\lambda = 2 \cdot 07$ m), pulse duration $3\,\mu$s and recurrence frequency 500/sec, peak power about 0·5 kW, and range 30 km. Type-A display is used.

621.396.9 : 519.2 **2221**

The Statistical Properties of Noise Applied to Radar-Range Performance.—S. M. Kaplan & R. W. McFall. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 487–489.) Discussion on 1390 of 1951.

621.396.932 **2222**

Radar Equipment at the Port of Le Havre.—(*TSF et TV*, Jan. 1952, Vol. 28, No. 279, p. 30.) Note of features of the 3-cm equipment to be installed. These include a 14-m paraboloidal aerial weighing about 5 tons and with a beam angle of only 42 minutes, a special system incorporating a semi-reflecting surface for rapid determination of the speed and direction of a moving vessel, and a cm-wave telephony system for ship communication.

621.396.933 **2223**

Some Navigational and Air Traffic Control Problems of Civil Aviation and the Application of Radio to their Reduction.—G. W. Stallibrass. (*J. Brit. Instn Radio Engrs*, Jan. 1952, Vol. 12, No. 1, pp. 3–20. Discussion, pp. 21–22.)

621.396.933.1.087.9 **2224**

The Design and Development of the Decca Flight Log.—G. E. Roberts. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, pp. 117–131.) The presentation of navigational data is discussed. The information given by the Decca Navigator receiver is displayed in a convenient form in the instrument described. The choice of coordinates is discussed and details are given of the various units used in the Mark 01 model.

621.396.933.2 : 623.451 **2225**

Miniature Transponder Beacon for Guided Missiles.—B. H. Sinclair. (*TV Engng, N.Y.*, Feb. & March 1952, Vol. 3, Nos. 2 & 3, pp. 8–9, 30 & 14–17, 28.) The unit comprises receiver, decoding, trigger, modulator, transmitter and duplexing circuits. It is housed in a pressurized container 2½ in. in diameter and 6⅜ in. long and weighs < 2 lb. A similar container holds the Zn-Ag_2O_2 battery unit and remotely controlled switching mechanism. For stability, 'etched-plate' circuits are used and small components are secured to the chassis with plastic compound.

534.88 **2226**

Echo Sounding at Sea (British Practice). [Book Review]—H. Galway. Publishers: Pitman & Sons, London, 35s. (*Marconi Rev.*, 1st Quarter 1952, Vol. 15, No. 104, p. 44.) "The book can be confidently recommended not only to shore staff and sea-going radio officers who will be responsible for fitting or maintenance of echo-sounding equipment, but also to ships' navigating officers."

621.396.93 **2227**

Radio Research Special Report No. 22. Siting of Direction Finding Stations. [Book Notice]—W. Ross & F. Horner. Publishers: H.M. Stationery Office, London, 1951, 42 pp., 1s. 6d. (*Govt Publ., Lond.*, Feb. 1952, p. 26.)

621.396.93 **2228**

Funkpeiler. Grundlagen und Anwendungen (Direction Finders, Fundamentals and Applications). [Book Review]—H. Gabler. Publishers: Deutsches Hydrographische Institut, Hamburg, 1951, 70 pp., 5.40 DM. (*Fernmeldetech. Z.*, Jan. 1952, Vol. 5, No. 1, p. 43.) A summarized treatment of the essentials of all the important theoretical and technical problems of radio direction finding in ship navigation.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 **2229**

The Ionization Gauge — Two Modifications.—J. H. Burrow & E. W. J. Mitchell. (*J. sci. Instrum.*, Jan. 1952, Vol. 29, No. 1, pp. 27–28.) Description of a gauge forming part of the main pumping line, giving increased sensitivity; and of a modification permitting determination of the time taken to contaminate a surface.

535.323 + 535.341] : 546.24 : 539.234 **2230**

Optical Properties of Tellurium in the Infra-Red.—T. S. Moss. (*Proc. phys. Soc.*, 1st Jan. 1952, Vol. 65, No. 385B, pp. 62–66.)

535.343.2-15 : [546.28 + 546.289 **2231**

Far-Infrared Transmission of Silicon and Germanium.—R. C. Lord. (*Phys. Rev.*, 1st Jan. 1952, Vol. 85, No. 1, pp. 140–141.)

535.37 **2232**

Electron Levels in Phosphorescent Crystals with Fe, Ni or Co Poison Centres.—N. Arpiarian & D. Curie. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1952, Vol. 234, No. 1, pp. 75–77.)

537.311.33 : 546.28 **2233**

The Drift Mobility of Electrons in Silicon.—J. R. Haynes & W. C. Westphal. (*Phys. Rev.*, 15th Feb. 1952, Vol. 85, No. 4, p. 680.) Measurements were made using a pulse-injection technique adapted from that described by Haynes & Shockley (1928 of 1951) for investigating Ge. An average value of 1 210 cm/s per V/cm was obtained. This is more than four times as great as the value obtained from Hall-effect data by Pearson & Bardeen (*Phys. Rev.*, 1949, Vol. 75, p. 865), using multicrystal specimens; in the present experiments single crystals were used.

537.311.33 : 546.289 **2234**

Rectification Phenomena and Transistor Action in Germanium.—P. Aigrain. (*Ann. Phys., Paris*, Jan./Feb. 1952, Vol. 7, pp. 140–184.) Full account of experiments carried out and theory developed, which were briefly reported in 1305, 1306, 1310 and 1818 of 1950, and 1529 of 1951.

537.311.33 : 546.289 **2235**
Properties of Thermally Produced Acceptors in Germanium.—C. S. Fuller, H. C. Theuerer & W. van Roosbroeck. (*Phys. Rev.*, 15th Feb. 1952, Vol. 85, No. 4, pp. 678–679.) Brief account of experiments extending the investigation of the effects of heat treatment on *n*-type Ge [2201 of 1951 (Theuerer & Scaff)]. Results indicate that (*a*) the *n*-to-*p* type conversion is characterized by the diffusion of a *p-n* boundary from the surface of the specimen to the interior, (*b*) the concentration of acceptor centres approaches an equilibrium value dependent on the heating temperature over the range from about 550°C to the melting point within which the conversion is possible.

538.221 **2236**
Properties of Ferromagnetic Powders at Frequencies up to 24 kMc/s.—B. Pistoulet. (*Ann. Télécommun.*, Jan.–March 1952, Vol. 7, Nos. 1–3, pp. 27–45, 85–97 & 127–138.) Measurements of the complex permeability of various metal powders embedded in dielectric material are reported for a wide range of frequencies. The preparation of the mixtures and the measurement methods and apparatus are described. A study is made of the variations of permeability as a function (*a*) of the proportion of magnetic powder in the test samples, (*b*) of the frequency. The results obtained enable the permeability characteristics of the powders themselves to be deduced. A general method for the study of magnetic resonance under the influence of a constant magnetic field is described, with results obtained at wavelengths of 3·2 and 1·25 cm. 52 references.

538.248 **2237**
Investigation of the After-Effect Constant over the Total Hysteresis Range.—J. C. Barbier. (*C. R. Acad. Sci., Paris*, 21st Jan. 1952, Vol. 234, No. 4, pp. 415–417.)

539.23 : 546.72 : 537.311.31 **2238**
Measurements on Thin Iron Films.—A. van Itterbeek, L. de Greve & F. Heremans. (*Appl. sci. Res.*, 1952, Vol. B2, No. 4, pp. 320–324.) Resistivity/thickness curves for sputtered and evaporated iron films are found to have different shapes, due to differences of texture revealed by electron-microscope photographs.

539.231 : 546.57 **2239**
The Structure of Sputtered Silver Films.—C. E. Ells & G. D. Scott. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 31–34.) Electron-microscope study shows that sputtered films have a more continuous structure than evaporated films prepared at the same rate, but if the evaporated films are produced at much faster rates the thinner sputtered films are more continuous and the thicker less continuous than evaporated films of the same thickness.

539.234 **2240**
Vacuum Evaporation Equipment with High Pumping Speed.—R. Bernard & F. Davoine. (*Le Vide*, Jan. 1952, Vol. 7, No. 37, pp. 1136–1138.) Increased fineness of film structure is achieved, since the pumping speed of 200 *l*/sec is sufficient to maintain the pressure at about 10^{-5} mm Hg.

546.561.221 : [537.311.33 + 538.632 **2241**
The Electrical Conductivity and Isothermal Hall Effect in Cuprous Sulphide, Semiconductor.—E. Hirahara. (*J. phys. Soc. Japan*, Nov./Dec. 1951, Vol. 6, No. 6, pp. 428–437.)

546.561.221 : 537.311.33 **2242**
The Physical Properties of Cuprous-Sulphide Semiconductors.—E. Hirahara. (*J. phys. Soc. Japan*, Nov./Dec. 1951, Vol. 6, No. 6, pp. 422–427.)

621.315.612.4 : 546.431.824-31 **2243**
Properties of Powdered $BaTiO_3$.—V. E. Derr & M. D. Earle. (*Phys. Rev.*, 15th Jan. 1952, Vol. 85, No. 2, pp. 384–385.) Discharge curves for capacitors with dielectric of tightly packed powdered $BaTiO_3$ are shown and discussed. The change of current with time differs from that of the usual type of capacitor, the terminal voltage falling quickly at first and then more slowly for many minutes.

621.315.612.4 : 546.431.824-31 **2244**
Some Factors Influencing the Dielectric Properties of Barium Titanates.—W. R. Eubank, F. T. Rogers, Jr, L. E. Schilberg & S. Skolnik. (*J. Amer. ceram. Soc.*, Jan. 1952, Vol. 35, No. 1, pp. 16–22.) The permittivity of ceramics of various grades of purity and fired at various temperatures was determined as a function of temperature in the region of the Curie point. The temperature coefficient of permittivity is largest for the purer titanates. The dependence of the Curie point on firing temperature, cooling rate and impurity content is described. The optimum firing temperature appears to be about 1400 °C. Contact with Pt or ZrO_2 during firing may be detrimental.

621.315.612.4.011.5 : 546.431.824-31 **2245**
Adiabatic Thermal Changes in Barium Titanate Ceramic at Low Temperatures.—R. W. Schmitt. (*Phys. Rev.*, 1st Jan. 1952, Vol. 85, No. 1, pp. 1–4.) Measurements at 4°K indicate that the polarization process in the ceramic is thermodynamically irreversible; this accounts for the large variation of dielectric constant with temperature at 4°K.

621.318.1 **2246**
Magnetic Cores and Sheaths in the Field of Telecommunications.—P. M. Prache. (*Câbles & Transmission, Paris*, Jan. & April 1952, Vol. 6, Nos. 1 & 2, pp. 22–64 & 124–164.) Detailed analysis of the characteristics of ribbon, wire and powder cores and of magnetic loading sheaths for cables, with all the relevant formulae for apparent permeability, eddy-current and hysteresis losses, etc. Calculated values of apparent permeability of powder cores suitable for Pupin coils are in good agreement with measured values. A method of cable loading by means of half-sheaths formed from insulated high-permeability wires enables operation at frequencies up to 200 kc/s. 60 references.

621.318.33 **2247**
The Use of Current Sheets in the Design of Magnets to give Bounded Fields of Required Form, free from Edge Distortion.—H. O. W. Richardson. (*Proc. phys. Soc.*, 1st Jan. 1952, Vol. 65, No. 385B, pp. 5–14.)

621.791.343 **2248**
Welding and Soldering Aluminum.—S. Freedman. (*Radio & Televis. News, Radio-Electronic Engng Section*, Feb. 1952, Vol. 47, No. 2, pp. 16, 18.) Using Chemalloy, a compound of various metals and chemicals, Al or any Zn-base metal can be soldered or welded simply by heating to about 800°F and applying a rod of the alloy.

666.1.037 : 539.319 **2249**
Theory of Stresses in Glass Butt Seals.—N. L. Svensson: M. Zaid: H. Rawson. (*Brit. J. appl. Phys.*, Jan. 1952, Vol. 3, No. 1, pp. 30–32.) Comments on 3032 of 1951 and author's reply.

MATHEMATICS

512.99 **2250**
Theory and Applications of Wave Vectors.—F. Dahlgren. (*Acta polyt., Stockholm*, 1951, No. 97,

68 pp.) Rules for the mathematical treatment of such vectors are given, as well as physical definitions of quantities which can be conveniently represented by them, these quantities mainly relating to electrical machines.

512.99 : 519.21 **2251**

The Probability Distribution of the Phase of the Resultant Vector Sum of a Constant Vector Plus a Rayleigh Distributed Vector.—K. A. Norton, E. L. Shultz & H. Yarbrough. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 137–141.) Formulae, tables and graphs are given of the cumulative probability distribution of a function frequently occurring in the theory and practice of radio wave propagation as well as in the study of the influence of noise in phase modulation systems.

517.63 **2252**

On Approximate Expressions for the Exponential Integral and the Error Function.—R. Bellman. (*J. Math. Phys.*, Jan. 1952, Vol. 30, No. 4, pp. 226–231.)

681.142 **2253**

A Simple Electronic Digital Computer.—W. L. van der Poel. (*Appl. sci. Res.*, 1952, Vol. B2, No. 5, pp. 367–400.) Description of a computer which has been simplified to the utmost practical limit at the sacrifice of speed, with examples showing the use made of sub-programmes in its operation.

681.142 **2254**

A Direct-Current Network Analyzer for Solving Wave-Equation Boundary-Value Problems.—G. W. Swenson, Jr, & T. J. Higgins. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 126–131.)

681.142 **2255**

New Techniques on the Anacom — Electric Analog Computer.—E. L. Harder & J. T. Carleton. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 547–556.) The direct-analogue method used in the Anacom is outlined and a description given of the sigma amplifier, which combines adding, integrating, delay, and other operations and results in improved computing technique. Various applications of the equipment are described.

681.142 : 512.25 **2256**

New Principle of Construction of Machines for Solution of Systems of Linear Equations by Electrical Analogy.—D. Mitrovic, R. Huron & R. Tomovic. (*C. R. Acad. Sci., Paris*, 4th Feb. 1952, Vol. 234, No. 6, pp. 589–591.)

681.142 : 517.942.9 **2257**

Three-Dimensional Electrical Potential Analyser.—V. E. Gough; S. C. Redshaw. (*Brit. J. appl. Phys.*, Feb. 1952, Vol. 3, No. 2, p. 58.) Comment on 1358 of May and author's reply.

681.142 : 621.3.042.14/.15 **2258**

A Coincident-Current Magnetic Memory Cell for the Storage of Digital Information.—W. N. Papian. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 475–478.) Binary information can be stored in small ferromagnetic cores, three-dimensional arrays of which may be built up so that 'writing' or 'reading' of a desired unit may be effected by exciting the appropriate coordinate lines. Criteria for core materials are set up and experimental results with some selected materials are described.

MEASUREMENTS AND TEST GEAR

53.081.4 **2259**

Fundamental Considerations regarding the Use of Relative Magnitudes.—J. W. Horton. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 440–444.) There are two number systems, conforming concurrently to the decimal system and related by the basic quantity $10^{0.1}$, by which relative magnitudes may be evaluated. The term 'logit' is suggested for the quantity $10^{0.1}$, which plays a similar part in computations dealing with relative magnitudes to that of the unit in computations involving absolute magnitudes. Methods of using the logit are outlined and the resulting advantages are discussed.

535.322.1.029.65 : 537.228.5 **2260**

Development of a Spectroscope for Millimetre Waves.—É. Roubine. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1951, No. 16, pp. 21–44.) An instrument for the K and J bands.

621.3.018.41(083.74) **2261**

The Transmission of Time Signals and Standard Frequencies by the I.E.N. [Istituto Elettrotecnico Nazionale Galileo Ferraris], **Turin.**—C. E. (*Alta Frequenza*, Oct. 1951, Vol. 20, No. 5, pp. 219–223.) A weekly experimental service with 300-W power commenced on May 15th, 1951. 5-Mc/s transmissions are made every Tuesday from 0900 to 1200 and from 1400 to 1700 (C.E.T.), each hour being subdivided into five-minute periods, with time signals alternating with either 440-c/s or 1 000-c/s modulated signals. Spoken announcements are made at the beginning of each hour and Morse-code announcements every ten minutes. A horizontal dipole aerial is used. Over short ranges frequency is guaranteed to within ± 2 parts in 10^8 and time to within ± 25 ms.

621.3.018.41(083.74) **2262**

Frequency Multiplier giving a 1 000-c/s Signal Synchronized by a Pendulum Chronometer.—P. Parcelier. (*C. R. Acad. Sci., Paris*, 7th Jan. 1952, Vol. 234, No. 2, pp. 190–191.)

621.317.3.027.3 : 621.385.3 **2263**

Some Properties and Applications of the Inverted Triode.—A. Rogozinski & J. Weill. (*J. Phys. Radium*, Feb. 1952, Vol. 13, No. 2, Supplement, pp. 28A–30A.) The main application is in the direct measurement of high voltages, for which a bridge-type circuit with two inverted triodes is suitable. The primary range 5–200 V can be extended to about 5 kV by use of a high-resistance bridge. The I/V characteristic of a Type-100 Th inverted triode suitable for use up to about 20 kV is shown, and a megohmmeter-voltmeter described which can also be used for the measurement of very low currents passed through a high resistance.

621.317.328 : 621.384.62† **2264**

Determination of Field Strength in a Linear-Accelerator Cavity.—L. C. Maier, Jr, & J. C. Slater. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 78–83.) One theoretical and two experimental methods are described for determining the accelerating field in the M.I.T. linear-accelerator cavity in terms of the input power. One experimental method is based on measurement of the power leaking out through a small hole in the end wall of the cavity. The other method depends on perturbation of the resonance frequency of the cavity by a small conducting sphere located on the axis. The three methods give consistent results.

621.317.328 : 621.396.611.4 **2265**

Field-Strength Measurements in Resonant Cavities.—L. C. Maier, Jr, & J. C. Slater. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 68–77.) The perturbation of the resonance frequency of a cavity due to insertion of ellipsoidal objects is calculated for objects of needle, sphere, and disk types. The perturbations for the three types of object depend on different components of the electric and magnetic fields, and by making measurements with all three it is theoretically possible to measure all the field components. Experimental verification of the theory was satisfactory for spheres and disks, but for needles the perturbation is very sensitive to needle shape and the needles used were not accurate enough ellipsoids to give satisfactory quantitative results.

621.317.333.4.015.7 : 621.315.212 **2266**

A Pulse-Echo Test Set for the Quality Control and Maintenance of Impedance Uniformity of Coaxial Cables.—E. Baguley & F. B. Cope. (*P.O. elect. Engrs' J.*, Jan. 1952, Vol. 44, Part 4, pp. 164–168.) The equipment, designed primarily for use on 0·375-in. coaxial cables, can detect impedance irregularities and differences in end impedance of the order of 0·05%. Impedance irregularities and mismatches can be located to within 1% on a direct-reading scale of yards or metres. A technique for cable testing during manufacture and installation is also described.

621.317.333.8 : 621.392.5 **2267**

Impulse Measurements by Repeated-Structure Networks.—C. L. Dawes, C. H. Thomas & A. B. Drought. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 571–580. Discussion, pp. 580–583.) Analysis shows that the ladder type of network considered possesses the ideal characteristic of an attenuating network for passing any kind of signal without distortion. Experimental results on a simple T structure indicate that a practical h.v. divider of high accuracy may be constructed. An appendix gives the Laplace-transform solution for the parallel-T type of ladder network.

621.317.335.3 + 621.317.372 **2268**

An Improved Method of Measuring Dissipation Factor and Dielectric Constant using the Susceptance Variation Principle.—C. F. Miller & F. G. Whelan. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 491–497.) Full paper. Summary abstracted in 2277 of 1950.

621.317.335.3.029.64 : 546.212-13 **2269**

The Dielectric Constant of Water Vapor in the Microwave Region.—G. Birnbaum. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 220–223.) Using a cavity method described previously [1426 of 1951 (Birnbaum et al.)], measurements were made over the temperature range 32–103°C at a frequency of 9·28 kMc/s and at the single temperature of 24·5°C at 24·8 kMc/s. Results are discussed.

621.317.335.3.029.64.012.3 **2270**

New Chart for the Determination of the Permittivity of Dielectrics at U.H.F.—A. Lebrun. (*C. R. Acad. Sci., Paris*, 28th Jan. 1952, Vol. 234, No. 5, pp. 518–520.) Description of the construction and use of a chart applicable to the method of measurement in which a short-circuited section of waveguide is filled with the dielectric and measurements are made of the width of the resonance curve or of the voltage s.w.r.

621.317.35 **2271**

Analysers for Aperiodic Phenomena.—G. Francini. (*Alta Frequenza*, Dec. 1951, Vol. 20, No. 6, pp. 247–261.) Direct and indirect electrical methods available for low frequencies are discussed and possible simplifications in design of apparatus are considered from the point of view of their effect on performance. Sources of possible error are discussed.

621.317.361 **2272**

Frequency Measurement by the Capacitor Charge Method: Part 2.—H. Weidemann. (*Arch. tech. Messen*, Jan. 1952, No. 192, pp. 11–14.) Conditions that must be satisfied for very rapid frequency variations to be followed by the meter, and practical limitations of the method are considered. Alternative arrangements are described in which stability of operation need be ensured only in the auxiliary devices used. These involve either multivibrator, thyratron or transitron circuits. Part 1: 732 of March.

621.317.373 **2273**

Phase Measurement.—J. Henry. (*Radio franç.*, Jan. & Feb. 1952, Nos. 1 & 2, pp. 13–22 & 9–17.) Discussion of the functions of the various components of phase meters, and also of complete phase meters of different types.

621.317.431 **2274**

A Two-Fluxmeter Method of Measuring Ferromagnetic Hysteresis Loss.—H. Aspden. (*J. sci. Instrum.*, Jan. 1952, Vol. 29, No. 1, pp. 5–7.) The time required for hysteresis-loop measurements can be greatly reduced by using a separate fluxmeter to integrate successive capacitor discharges, which are initiated by a chosen change in flux density and are proportional to the magnetizing field.

621.317.44 **2275**

Magnet-Steel Test Unit with Fluxmeter Compensated for Controlling Force.—E. Steingroever. (*Arch. Elektrotech.*, 1952, Vol. 40, No. 5, pp. 275–279.)

621.317.444 : 621.396.645.35 : 621.317.755 **2276**

An Integrating Amplifier for the Oscillographic Recording of Magnetic Flux.—S. Ekelöf, L. Bengtson, G. Kihlberg & P. Leithammel. (*Acta polyt., Stockholm*, 1951, No. 98, 23 pp.) Theory of operation and circuit details of a push-pull integrating amplifier consisting of a d.c. amplifier with capacitive feedback. Precautions are taken to obtain stable gain and low output-voltage drift. The instrument was designed for use in obtaining oscillographic records of transient magnetic fluxes lasting 0·01–0·5 sec, the flux change being 500–5 000 maxwells.

621.317.6.029.3 : 621.317.755 **2277**

Automatic Audio-Frequency Response-Curve Tracer.—(*Radio tech. Dig., Édn franç.*, 1951, Vol. 5, No. 6, pp. 339–347 & 1952, Vol. 6, No. 2, pp. 77–87.) Adaptation of articles by Hamburger (148 of 1949 and 1047 of April), with supplementary data from other sources.

621.317.7 : 621.3.015.7 **2278**

A Video Probe.—R. R. Rathbone. (*Radio & Televis. News, Radio-Electronic Engng Section*, Jan. 1952, Vol. 47, No. 1, pp. 16..29.) Description of a probe suitable for testing pulse equipment; it comprises an attenuator and cathode follower feeding a terminated coaxial cable with a characteristic impedance of 93 Ω. The cable may have any length up to 100 ft without introducing reflections.

621.317.7.088 **2279**

Scale-Overlap Errors and Frequency Errors in Instruments with Barrier-Layer Rectifiers.—J. Hajek. (*Arch. tech. Messen*, Jan. 1952, No. 192, pp. 19–22.) Scale-overlap errors are calculated from the shape of the rectifier characteristic, (*a*) taking account of and (*b*) neglecting back current. Frequency errors and frequency dependence in Graetz full-wave rectifier circuits are also considered.

621.317.715 : 523.723 **2280**

Brownian Fluctuations in Galvanometers and Galvanometer Amplifiers.—R. V. Jones & C. W. McCombie. (*Philos. Trans. A.*, 24th Jan. 1952, Vol. 244, No. 881, pp. 205–230.) The effects of molecular bombardment of the galvanometer mirror and of noise in the circuit resistance are studied, using the correlation function of the random force. The inclusion of circuit inductance is shown to cause no change in the r.m.s. values of the deflection and angular velocity. The magnitude and correlation function of the fluctuations in a galvanometer amplifier can be obtained from the results of a simple experiment. Close agreement has been obtained between theoretical and experimental values.

621.317.715.082.742 2281

Use of a Moving-Coil Galvanometer to measure the Mean Charging Current of a Periodically Discharged Capacitor.—R. Legros. (*Ann. Phys., Paris*, Jan./Feb. 1952, Vol. 7, pp. 5–29.) The galvanometer and capacitor form part of an electronic frequency meter (3187 of 1948 and 1899 of 1949). When frequencies are to be measured which are close to the natural frequency of the moving parts, oscillations are set up. These are discussed in detail and the resultant reading errors evaluated, various types of scale graduation being considered.

621.317.715.082.742 2282

Moving-Coil Galvanometers with High Voltage Sensitivity. Applications to Problems of Biological Physics.—J. Coursaget. (*Ann. Phys., Paris*, Jan./Feb. 1952, Vol. 7, pp. 30–90.)

621.317.729 : 537.291 2283

Automatic Tracer for Electron Trajectories and its Use for Determining the Current Lines in an Electrolyte Trough.—J. Marvaud. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1952, Vol. 234, No. 1, pp. 45–47.)

621.317.729 : 538.311 2284

The Extension of the Electrolytic Tank Method to the Study of Magnetic Fields due to Iron-clad Current Sheets in Three Dimensions.—H. O. W. Richardson. (*Proc. phys. Soc.*, 1st Jan. 1952, Vol. 65, No. 385B, pp. 15–18.)

621.317.75 : 621.396.615.14 2285

Sweep-Frequency Generator for U.H.F. Television Band.—J. A. Cornell & J. F. Sterner. (*Tele-Tech*, Feb. 1952, Vol. 11, No. 2, pp. 38–40 .. 88.) Description of an instrument designed for laboratory investigations of the characteristics of filters, tuning units and other components used in the 470–890-Mc/s band. It comprises a sweep-frequency oscillator modulated by a vibrating mechanism, a variable-frequency marker oscillator, a crystal calibrator providing 1-Mc/s check points, and an arrangement of mixers which superimposes marker and calibration pips upon the response curves displayed on a c.r.o.

621.317.755 : 621.317.74.018.782.4† 2286

A Scanner for Rapid Measurement of Envelope Delay Distortion.—L. E. Hunt & W. J. Albersheim. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 454–459.) Description of c.r.o. equipment for display of the envelope delay/frequency characteristic of a transmission system. It has proved very useful for measurements on the TD-2 radio-relay system and for adjustment of the equalizers used in the system.

621.383 : 621.396.645.35 2287

Photoelectric Amplifiers.—A. Schaller. (*Arch. tech. Messen*, Feb. 1952, No. 193, pp. 43–46.) Basic principles are outlined of photocell applications for the measurement and amplification of very small voltages and currents, and short descriptions are given of amplifiers developed by Leo & Hübner, Lawson, Gall, and Siemens & Halske A.-G. 18 references.

621.385.001.4 2288

Dynamic Measurements on Receiving Valves.—A. J. Heins. (*J. Brit. Instn Radio Engrs*, Jan. 1952, Vol. 12, No. 1, pp. 63–68.) Three sets of direct-reading dynamic test equipment, designed for measurements of equivalent noise resistance, power output and distortion, and cross-modulation, are described. The tests can be made by relatively unskilled operators on batches of receiving valves in quantity production.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.771 2289

A New R.P.M. Indicator.—(*Overseas Engr*, Feb. 1952, Vol. 25, No. 290, pp. 250–251.) Description of a direct-reading pulse-counter tachometer for measuring the rotational speed of prime movers over the range 500–10 000 r.p.m. accurate to within ± 1 r.p.m. An electrically maintained tuning fork or low-frequency crystal serves as time standard.

535.336.2.05 : 621.389 2290

Defects in the Mass Spectrometer due to Space Charge and Magnetic-Field Saturation.—E. W. Becker & W. Walcher. (*Z. Phys.*, 19th Feb. 1952, Vol. 131, No. 3, pp. 395–407.)

536.587 : 537.312.6 : 621.316.86 2291

Temperature Regulation using Thermistors.—N'Guyen Thien-Chi & J. Suchet. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 75–77.) Three materials are available for covering the range 100°C–1 150°C in three stages. Simple but rugged commercial equipment is described for controlling one or more furnaces simultaneously, the thermistors operating relays directly.

538.569.2.047.029.63/.64 2292

Microwaves in Medical and Biological Research.—J. E. Roberts & H. F. Cook. (*Brit. J. appl. Phys.*, Feb. 1952, Vol. 3, No. 2, pp. 33–40.) Recent work on the absorption of radiation in the frequency range 1–30 kMc/s is reviewed, with particular reference to materials of biological interest. 37 references.

621.316.7 2293

The Cranfield Conference on Automatic Control, 16th–21st July 1951.—J. Loeb. (*Ann. Télécommun.*, Jan. 1952, Vol. 7, No. 1, pp. 17–26.) A review of the proceedings, with short summaries of the most important papers presented. See also 1061 of April (Tustin).

621.384.62† 2294

Experimental Study of a Waveguide Electron Accelerator.—J. Vastel. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 20–33.) Full details of the design, construction and testing of the prototype linear accelerator section described by Grivet & Vastel (1738 of 1951) in *C. R. Acad. Sci., Paris*, where the magnetron peak power was incorrectly given as 0·5 mW instead of 0·5 MW.

621.384.62† 2295

The M.I.T. Linear Electron Accelerator.—P. T. Demos, A. F. Kip & J. C. Slater. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 53–65.)

621.384.62† : 531.314.3 2296

Particle Dynamics in the Linear Accelerator.—J. R. Terrall & J. C. Slater. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 66–68.)

621.384.62† : 621.317.328 **2297**

Determination of Field Strength in a Linear-Accelerator Cavity.—Maier & Slater. (See 2264.)

621.385.833 **2298**

Some Methods for Determination of the Field on the Axis in Electron Optics.—F. Bertein. (*C. R. Acad. Sci., Paris*, 21st Jan. 1952, Vol. 234, No. 4, pp. 417–419.)

621.385.833 **2299**

Trajectories in Electron Lenses: a Method of Approximation.—F. Bertein. (*J. Phys. Radium*, Feb. 1952, Vol. 13, No. 2, Supplement, pp. 41A–49A.)

621.385.833 **2300**

A New Mathematical Model of an Electron Lens.—P. Grivet. (*J. Phys. Radium*, Feb. 1952, Vol. 13, No. 2, Supplement, pp. 1A–9A.) Full paper. See 1394 of May.

621.385.833 **2301**

Cardinal Parameters of a New Model of an Electron Lens.—P. Grivet. (*C. R. Acad. Sci., Paris*, 2nd Jan. 1952, Vol. 234, No. 1, pp. 73–75.) Simple formulae are given to facilitate use of the mathematical model previously described (1394 of May) and values of the cardinal parameters are tabulated.

621.385.833 **2302**

Theory of the Electrostatic [electron] **Lens formed by Two Coaxial Cylinders.**—P. Grivet & M. Bernard. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 3–9.) Mathematical theory leading to simple formulae for calculating parameters. See also 1393 of May.

621.385.833 **2303**

Numerical Ray-Tracing in Electron Lenses.—J. C. Burfoot. (*Brit. J. appl. Phys.*, Jan. 1952, Vol. 3, No. 1, pp. 22–24.) A simple step-by-step method is described whose accuracy can be increased to any desired extent without increasing the complexity. A numerical example is given for a strong e.s. lens.

621.385.833 **2304**

Characteristics of the Hot-Cathode Electron-Microscope Gun.—M. E. Haine & P. A. Einstein. (*Brit. J. appl. Phys.*, Feb. 1952, Vol. 3, No. 2, pp. 40–46.)

621.385.833 **2305**

Summarized Proceedings of a Conference on Electron Microscopy — St. Andrews, June 1951.—D. G. Drummond & G. Liebmann. (*Brit. J. appl. Phys.*, Jan. 1952, Vol. 3, No. 1, pp. 25–29.)

621.385.833 : 061.3 **2306**

Proceedings of the Electron Microscope Society of America.—(*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 156–164.) Summaries are given of 44 papers presented at the annual meeting of the society in Philadelphia, November 1951.

621.385.833 : 537.291 **2307**

A Reduced Equation for the Trajectories in an Electron Mirror.—M. Bernard. (*C. R. Acad. Sci., Paris*, 4th Feb. 1952, Vol. 234, No. 6, pp. 606–608.)

621.387.462 : 549.211 **2308**

Differences between Counting and Non-Counting Diamonds.—G. P. Freeman & H. A. van der Velden. (*Physica*, Jan. 1952, Vol. 18, No. 1, pp. 1–19.)

621.398 **2309**

The Radio-Controlled Aircraft Winner of the International Contest 1950.—A. Wastable. (*TSF et TV*, Jan. 1952, Vol. 28, No. 279, pp. 11–12.) Brief description of the telecontrol system. A superregenerative receiver operates a master relay according to the pulse sequence transmitted. The airborne equipment weighs 750 g.

681.177 **2310**

An Electronic Digital Recording Machine — the SETAR.—N. T. Welford. (*J. sci. Instrum.*, Jan. 1952, Vol. 29, No. 1, pp. 1–4.) Description of the design and principles of operation of a 'serial event timer and recorder' developed for studying human performance. Events are recorded in sequence in digital code on standard teleprinter tape. An event is defined as the making or breaking of one or more input circuits. A continuously running generator provides timing pulses at 100 per sec or 10 per sec.

PROPAGATION OF WAVES

538.566 **2311**

An Integral-Equation Approach to the Problem of Wave Propagation over an Irregular Surface.—G. A. Hufford. (*Quart. appl. Math.*, Jan. 1952, Vol. 9, No. 4, pp. 391–404.) Theoretical discussion of the propagation of radio waves over a surface whose radius of curvature is everywhere much larger than a wavelength. It is assumed that a scalar wave phenomenon is involved and that a homogeneous boundary condition applies at the surface. An integral equation is derived for the attenuation function, at all points on the earth's surface and a formal solution for the field at any point above the earth is obtained. The analysis is applied to the special cases of a plane earth and a spherical earth. Agreement with the earlier work of Norton (33 of 1938), van der Pol & Bremmer (3102 of 1938), and Fock (2891 of 1947) is noted.

538.566.2 **2312**

Determination of the Fine Structure of the Dielectric Constant in a Slightly Heterogeneous Layer by Reflection Measurements.—G. Eckart. (*C. R. Acad. Sci., Paris*, 14th Jan. 1952, Vol. 234, No. 3, pp. 309–311.) The method of analysis previously described (1882 of 1951) for determining the variation of the dielectric constant across the layer demands an impossibly high experimental accuracy; Bremmer's method (205 of 1950), using the WKB approximation, is preferable. Integration of the function derived for the reflected signal leads to an integral equation which is solved by a Fourier transformation. The analysis is performed for a plane incident wave, and the modification necessary for the case of a spherical wave is indicated.

621.396.11 + 535.222 **2313**

A New Determination of the Velocity of Electromagnetic Radiation by Microwave Interferometry.—K. D. Froome. (*Nature, Lond.*, 19th Jan. 1952, Vol. 169, No. 4290, pp. 107–108.) The free-space phase velocity of waves of frequency 24 kMc/s has been determined by means of apparatus which is the microwave equivalent of the Michelson optical interferometer. The apparent wavelength in air was observed by movement of a reflector through a distance corresponding to an exact integral number of energy minima at the detector, and could be determined from a single experiment with an accuracy to within ± 3 parts in 10^6, the total displacement of the reflector being about 162 cm. Frequency was determined by comparison with a high harmonic of a standard quartz oscillator. The results, when referred to vacuum conditions, gave $c_o = 299\,792{\cdot}6 \pm 0{\cdot}7$ km/s.

621.396.11 **2314**

Oblique Reflexion of Radio Waves by Way of a Triangular Path.—J. H. Meek. (*Nature, Lond.*, 23rd Feb.

1952, Vol. 169, No. 4295, p. 327.) Traces due to waves reflected first from the F layer and then from an E_s cloud are shown. From a series of records obtained at 15-sec intervals, a speed of about 330 km/hour was calculated for E_s clouds.

621.396.11 : 551.510.535 **2315**

Application of the Appleton-Hartree Formula to the Determination of Phase Path of an Electromagnetic Wave in the Ionosphere.—É. Argence. (*C. R. Acad. Sci., Paris*, 21st Jan. 1952, Vol. 234, No. 4, pp. 456–458.) A method is described of determining the phase path without the use of the generalized magneto-ionic theory. The relation with Booker's method (422 of 1939) is indicated. Special cases for which the formulae become simplified are (*a*) east-west propagation, (*b*) propagation at the equator, (*c*) propagation at the poles. For the area round the poles the correction necessary to the usual predicted frequencies is large. The theory is applicable to the propagation of wave packets as discussed by Booker (714 of 1950).

621.396.11 : 621.392 **2316**

Transmission Lines as Models for the Study of Electromagnetic Wave Propagation in One Dimension.—L. Lunelli. (*Alta Frequenza*, Oct. & Dec. 1951, Vol. 20, Nos. 5 & 6, pp. 179–199 & 262–282.) Two formal analogies are established between Maxwell's equations for propagation in a medium with constant parameters and the equations for propagation along a transmission line. The first relates electric field intensity to line voltage and magnetic field intensity to line current. The second inverts these relations. Similitude ratio and conditions are developed for both analogies, and tables of parameters are provided for the three types of line considered, viz., twin solid or stranded conductors, and coaxial cables. The practical design of models is explained in detail, the limitations, errors and difficulties involved being fully treated. Tables summarize possible solutions in various typical cases.

621.396.11.029.56 : 551.510.535 **2317**

Reflection of Short Waves at Heights less than 100 km.—W. Dieminger & A. E. Hoffmann-Heyden. (*Naturwissenschaften*, Feb. 1952, Vol. 39, No. 4, pp. 84–85.) Waves of wavelength from 75 to 200 m are reflected at heights of 75–100 km. A diurnal variation of the reflection height occurs, with a minimum about mid-day. The height is practically independent of the frequency used. The echo amplitude varies irregularly with an average period of several seconds without corresponding reflection-height variations. Echoes are strongest in the daytime, certainly in winter, and they are observed on days when ionospheric absorption is particularly high. The characteristics of the reflecting layer concerned are discussed.

621.396.81 **2318**

An Improved Method for the Calculation of the Field Strength of Waves Reflected by the Ionosphere.—K. Bibl, K. Rawer & E. Theissen. (*Nature, Lond.*, 26th Jan. 1952, Vol. 169, No. 4291, pp. 147–148.) In previous calculations the blanketing effect of the E layer has been assumed to occur sharply at a given frequency; because of refraction and selective absorption in the E layer this effect actually takes the form of a gradual transition dependent on amplitude. Numerical values for particular transmission paths of interest have been calculated and are to be published separately.

621.396.81 : 621.396.65 **2319**

Statistics of Propagation in the 5-m Waveband for Distances greater than the Optical Range.—H. Schröder. (*Frequenz*, Jan. 1952, Vol. 6, No. 1, pp. 20–25.) An analysis is made of field-strength measurements, taken over a period of a year, of the signal received at Berlin over the 213-km radio link from Bocksberg, using frequencies of 60 and 68 Mc/s. Results for selected days and mean values for ten consecutive days in each of the four seasons are shown in charts; in general, the hours of densest telephone traffic do not coincide with the times when transmission conditions are best. Graphs show the probability of attainment of specified signal levels and signal/noise ratios. The effects of reducing the number of dipoles in the aerial array and of reducing transmitter power are discussed. Comparison is made between the measured field strengths and values calculated from theory.

621.396.812.4 : 551.510.535 **2320**

A Note on Ionospheric Conditions which may affect Tropical Broadcasting Services after Sunset.—B. W. Osborne. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, p. 110.) Variations in the height and structure of the F_2 layer at sunset may lead to rapid and intense fading in short-distance transmission at low latitudes. The layer may disintegrate entirely at about the time of sunset on the ionosphere. At Singapore these effects are most frequent at the equinoxes between 1900 and 2100 local time and may occur on half the days of any month. See also 989 of April.

RECEPTION

621.396.621 : 621.396.619.13 + 621.392.52 **2321**

A Comparison between Two-Circuit Band-Pass Filters and Modulation Converters in the Riegger Circuit [ratio detector].—A. Nowak. (*Funk u. Ton*, Feb. 1952, Vol. 6, No. 2, pp. 75–83.) Discussion of the primary and secondary voltages in the two-circuit i.f. band-pass filter of a particular frequency discriminator, and of the dependence of the circuit voltages on circuit damping d and coupling coefficient k. The ratio k/d is an important parameter for the design of such filters; a simple method of measuring it is described.

621.396.621.001.11 **2322**

Time Analysis and Filtering.—J. Icole & J. Oudin. (*Ann. Télécommun.*, Feb. 1952, Vol. 7, No. 2, pp. 99–108.) The theory of the detection of information in the presence of random noise is discussed. In cases of (*a*) sinusoidal signals of uncertain frequency, and (*b*) signals in the form of coherent noise, methods based on time correlation analysis are advantageous; these include time-displacement, frequency-displacement, directional and intercorrelation methods. Integration and summation techniques based on mean values are analogous to simple frequency filtering and are less suitable in these cases.

621.396.621.54 **2323**

Tracking Problem in the Superheterodyne.—J. Mohrmann. (*Fernmeldetech. Z.*, Jan. 1952, Vol. 5, No. 1, pp. 24–30.) A critical review of published work on the subject and explanation of a graphical method of solving the problem.

621.396.622.71 : 621.396.619.13 **2324**

Theory and Practice of the Ratio Detector.—H. Marko. (*Frequenz*, Jan. 1952, Vol. 6, No. 1, pp. 1–10.) The operation of the ratio detector, in particular its amplitude-limiting action, is explained simply by substituting for the rectifier circuit a linear equivalent circuit of the type previously described (1120 of April). The method also makes it easy to estimate the effects of circuit asymmetry and deviations of component values from nominal. Results are confirmed by measurements on an actual circuit.

621.396.82 : [621.396.619.11/.13 **2325**

Interference in F.M. and A.M. Reception due to Weak Interfering Transmitters.—M. Kulp. (*Arch. elekt. Übertragung*, Jan. 1952, Vol. 6, No. 1, pp. 17–28.) New relations for Bessel functions are used to calculate the effective value of the total interference in f.m. reception due to an unwanted f.m. transmitter. Cases considered include: common and different carrier frequencies, modulated and unmodulated carriers, low-pass filter used or not used. Interference due to an unwanted a.m. transmitter is also considered. To enable a.m. and f.m. conditions to be compared, corresponding formulae are given for a.m. reception disturbed by a.m. or f.m. interference. Results are tabulated.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **2326**

Instantaneous Power Spectra.—C. H. Page. (*J. appl. Phys.*, Jan. 1952, Vol. 23, No. 1, pp. 103–106.) "The intuitive concept of a changing spectrum is discussed. The instantaneous power spectrum is defined mathematically and used to make the intuitive concepts more precise. It depends upon the past history of a signal, but not upon the future. Integration of the instantaneous power spectrum over time yields the conventional energy spectrum. The instantaneous power spectrum of a random function may be averaged over the ensemble of functions, with a resulting stochastic average instantaneous power spectrum that is equal to the conventional time average power spectrum of a stochastic process."

621.39.001.11 **2327**

The Evaluation of Communication and Transmission Methods in Communication Engineering.—J. Piesch. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, Jan./Feb. 1952, Vol. 6, Nos. 1/2, pp. 13–21.) Discussion of Shannon's theory of communication (1361 of 1949).

621.395 : 061.3 **2328**

The XVIth Plenary Assembly of the C.C.I.F. (Florence, 1951).—(*Poste e Telecomunicazioni*, Feb. 1952, Vol. 20, No. 2, pp. 75–81.) Report on work done by the various commissions. See also *Fernmeldetech. Z.*, April 1952, Vol. 5, No. 4, pp. 186–190, and *P.O. elect. Engrs' J.*, April 1952, Vol. 45, Part 1, pp. 35–38.

621.396.4.018.78 : 621.396.619.13 **2329**

Linearity Limits of Discriminators, particularly for Wide-Band F.M. Radio Beam Links.—P. Barkow. (*Fernmeldetech. Z.*, Feb. 1952, Vol. 5, No. 2, pp. 67–78.) The origins of distortion of f.m. in directional systems are investigated and the results of a series of measurements on a push-pull type of discriminator are presented in tables and numerous diagrams. Discussion of the principles of the process of modulation conversion in discriminators indicates that with careful design it should be possible to reduce distortion below 7·5 neper, a value which should not be exceeded for discriminators used on multichannel wide-band links.

621.396.619.16 **2330**

A New Method of Code Modulation: 'Δ-Modulation'.—L. J. Libois. (*Onde élect.*, Jan. 1952, Vol. 32, No. 298, pp. 26–31.) An analysis of the delta p.c.m. system in which the information transmitted refers to the slope of the input-signal characteristic. The method is based on differential analysis of the input signal and comparison with the signal decoded locally. The signal/noise ratio of the system is proportional to $(F/f_m)^3$, where F is the pulse code repetition frequency and f_m the highest modulation frequency. From the point of view of telephony quality, except when the frequency of the code pulses is very high, the system is equivalent to a 6-unit p.c.m. system of equal bandwidth. Tests with experimental equipment confirm this. Theory indicates that for high pulse-code frequencies the quality obtainable with p.c.m. is much better than with delta modulation. See also *J. Brit. Instn Radio Engrs*, July 1950, Vol. 10, No. 7, pp. 242–243 (Beard).

621.396.619.16 : [621.3.018.78 + 621.396.822 **2331**

Background Noise and Distortions in Code Modulation.—L. J. Libois. (*Câbles & Transmission, Paris*, Jan. 1952, Vol. 6, No. 1, pp. 65–79.) Three sources of noise affecting p.c.m. transmissions are examined: noise of external origin, such as circuit noise, disturbances due to coding errors, and quantization noise. Circuit noise at least 20 db below the signal can be regarded as negligible, so that a signal/noise ratio of 30 db should suffice when fading is taken into account. Effects of coding errors can be practically eliminated by use of a series coder, such as a binary counter. Quantization noise analysis shows that when the sampling frequency is twice the highest modulation frequency to be transmitted, as is practically the case in pulse multiplex, for which the ratio is about 2·5, all the distortion energy is found within the modulation band, and the signal/noise ratio, for 100% modulation, is equal to $\sqrt{6}p$, where $2p$ is the number of quantization steps effectively used. Experimental results on the quality of telephone conversation in the presence of noise indicate that a 6-unit code system, using a series coder and compressor, should enable quite satisfactory quality to be obtained.

621.396.619.16 : 621.3.018.78 **2332**

Analysis of Distortion in Pulse-Code Modulation Systems.—J. P. Schouten & H. W. F. van't Groenewout. (*Appl. sci. Res.*, 1952, Vol. B2, No. 4, pp. 277–290.) Theoretical treatment of the distortion introduced by amplitude quantization. The analysis is general and permits arbitrary choice of quantization level, of sampling frequency and of signal frequency. The results are applied to the case of a sinusoidal input signal.

621.396.822 : 621.396.619.11 **2333**

On the Distribution of Energy in Noise- and Signal-Modulated Waves: Part 1—Amplitude Modulation.—D. Middleton. (*Quart. appl. Math.*, Jan. 1952, Vol. 9, No. 4, pp. 337–354.) Theoretical analysis of the spectral distribution of intensity of a.m. of a carrier by noise or by signal and noise; in the latter case particular attention is paid to the case of a sinusoidal modulating signal. The mean carrier power, the mean total power and the mean continuum power are deduced as functions of the noise and signal modulation indices. Some of the results are shown graphically. The effect on the spectrum of over-modulation by the signal and/or noise is discussed and illustrated qualitatively.

621.396.933 **2334**

Gapless Coverage in Air-to-Ground Communications at Frequencies above 50 Mc/s.—K. A. Norton & P. L. Rice. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 470–474.) There is an optimum height of ground-station aerial for an air-to-ground communication system. With heights less than the optimum, the maximum range is reduced at all aircraft altitudes. When higher aerials are used, interference between the direct and ground-reflected waves causes gaps in the coverage at the higher aircraft altitudes. The optimum aerial height decreases with increasing frequency. Sets of curves show the variation with frequency of the optimum aerial height ensuring gapless communication to the maximum range at all aircraft altitudes less than (*a*) 10 000, (*b*) 25 000, (*c*) 40 000 ft. Other curves show the maximum range for satisfactory communication at

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selected aircraft altitudes from 1 000 to 40 000 ft, assuming optimum ground-station aerial height.

621.396.97 + 621.396.677.2 **2335**

High Frequency Broadcast Transmission with Vertical Radiation.—P. Adorian & A. H. Dickinson. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, pp. 111–116.) The necessity for exact siting of the transmitter is avoided by using vertical transmission, with reflection from the F_2 layer, at frequencies between about 2 and 10 Mc/s. Reasonably good reception can be maintained at ranges up to 150–200 miles in latitudes between 10°N and 20°N with two frequencies, the higher one being used during the day-time. Details are given of three suitable types of aerial.

621.396.97 + 621.396.975 **2336**

Wireless Broadcasting and Rediffusion Systems for Colonial Territories.—A. Cross & F. R. Yardley. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, pp. 91–109.) A description of the broadcasting system now installed in the colony of Trinidad. The system comprises a medium-wave service for the densely populated areas, a short-wave service for rural areas and supplementary facilities given by rediffusion programmes and experimental community reception in small villages. The installation of these services and the reception of B.B.C. and U.S.A. overseas broadcast programmes are discussed. Details are given of the 72-Mc/s links used with the rediffusion services.

SUBSIDIARY APPARATUS

621-526 **2337**

Control-System Synthesis by Root Locus Method.—W. R. Evans. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 66–69.) Description, with examples, of a graphical method of determining all the roots of the differential equation of a control system; the method readily permits synthesis for a desired transient response or frequency response.

621-526 **2338**

A Theory of Multidimensional Servo Systems.—M. Golomb & E. Usdin. (*J. Franklin Inst.*, Jan. 1952, Vol. 253, No. 1, pp. 29–57.)

621-526 **2339**

Stability and Parametric Continuity of a Linear Servomechanism with Time-Varying Coefficients.—J. Brodin. (*C. R. Acad. Sci., Paris*, 18th Feb. 1952, Vol. 234, No. 8, pp. 800–801.)

621.314.634 **2340**

Studies on Selenium and its Alloys: Report 2 — Some Experiments on the Rectifying Characteristics of Selenium Rectifier.—T. Sato & H. Kaneko. (*Technol. Rep. Tohoku Univ.*, 1950, Vol. 15, No. 1, pp. 1–10.) Results of resistance measurements lead to the conclusion that the blocking layer is a thin film of amorphous Se. Report 1: 1336 of May.

621-526 **2341**

Selected Government Research Reports. Vol. 5. Servomechanisms. [Book Notice]—Publishers: H.M. Stationery Office, London, 310 pp., 63s. (*Govt Publ., Lond.*, Feb. 1952, p. 26.)

TELEVISION AND PHOTOTELEGRAPHY

621.397.24 : 621.315.212.4 **2342**

The Birmingham-Manchester Television Link.—(*P.O. elect. Engrs' J.*, Jan. 1952, Vol. 44, Part 4, p. 158.) Vision signals are relayed from Birmingham to the Holme Moss station by special coaxial cables via Telephone House, Manchester. Asymmetric sideband transmission is used, the carrier frequency being 1·056 Mc/s. Amplifiers are installed at 6-mile intervals along the cable, which can also provide 1 200 telephone channels.

621.397.5 : 535.623 **2343**

The National Television Systems Committee Color-Television Transmission: Part 1.—R. M. Bowie & B. F. Tyson. (*Sylvania Technologist*, Jan. 1952, Vol. 5, No. 1, pp. 10–16.) A general description of the N.T.S.C. system in which the necessary colour information is added to the monochrome transmission by vestigial-sideband modulation of two sub-carriers in phase quadrature. See also 1750 of June (Hirsch et al.).

621.397.5(083.74) **2344**

Belgian Television Standards.—G. Hansen. (*HF, Brussels*, 1952, Vol. 2, No. 1, pp. 7–15.) The standards adopted are discussed in relation to the availability of French and Dutch programmes and the cost of commercial receivers. The standards include both 625-line and 819-line definition, positive modulation with 5-Mc/s video bandwidth, and a.m. for the sound channel. The increase in cost of a receiver for two channels over that of a similar single-channel receiver is small.

621.397.5(494) **2345**

Television in Other Lands and Television Planning for Switzerland.—(*Tech. Mitt. schweiz. Telegr. TelephVerw.*, 1st Jan. 1952, Vol. 30, No. 1, pp. 19–32. In German.) A review of progress in various countries throughout the world and an outline of developments proposed for Switzerland up to 1953.

621.397.61 **2346**

The Du-Mitter.—S. R. Patremio. (*Radio & Televis. News, Radio-Electronic Engng Section*, Feb. 1952, Vol. 47, No. 2, pp. 3–5, 31.) A transmitter covering channel 2 or 3, and transforming television line signals to r.f. signals for feeding large groups of receivers in exhibitions, conferences, offices, etc.

621.397.611.2 **2347**

Electromagnetic Scanning Generators for Television.—L. W. Whitaker. (*Marconi Rev.*, 1st Quarter 1952, Vol. 15, No. 104, pp. 1–24.) The basic problems associated with the design of circuits for obtaining the required current waveforms in the deflection coils are considered for the case of both line and frame scanning. The various types of scanning generators are classified and methods of using feedback in order to obtain a linear sweep are dealt with in some detail. A method of obtaining correction in flat-faced tubes is outlined. The design and operation of several types of complete line and frame scanning generators using feedback for linearization are described in detail.

621.397.611.2 **2348**

Charging of Secondary-Emission Surfaces.—R. Colberg. (*Fernmeldetech. Z.*, Feb. 1952, Vol. 5, No. 2, pp. 56–66.) A short review of the phenomena, with descriptions of applications in various iconoscopes, the orthicon and image orthicon. 30 references.

621.397.611.2 **2349**

The Development of Storage-Type Television Camera Tubes.—A. Karolus. (*Z. angew. Phys.*, Feb. 1952, Vol. 4, No. 2, pp. 71–77.) A survey with 32 references.

621.397.62 : 535.88 **2350**

The Fischer Large-Screen Projection System.—E. Baumann. (*J. Brit. Instn Radio Engrs*, Feb. 1952, Vol. 12, No. 2, pp. 69–78.) Account of latest developments of 'eidophor' equipment. See also 485 of 1951.

621.397.621.2 **2351**

Scanning and E.H.T. Circuits for Wide-Angle Picture Tubes.—E. Jones. (*J. Brit. Instn Radio Engrs*, Jan. 1952, Vol. 12, No. 1, pp. 23–48.) A discussion of energy-recovery scanning circuits primarily designed for television receivers in which a c.r. tube with scanning angle of 70° is supplied from a circuit for which the line voltage is restricted to 190 V. Design procedures are established, with particular attention to obtaining linearity of the trace by use of a saturable reactor with a ferrite core.

621.397.621.2 **2352**

Magnetic Beam-Deflection Systems.—H. Bähring. (*Funk u. Ton*, Jan. 1952, Vol. 6, No. 1, pp. 8–24.) The deflection sensitivity, inductance, magnetic efficiency and overall quality factor of various types of deflection system are discussed. A formula derived for the quality factor enables comparison to be made between different types. Of the various types considered, the cylindrical screened coil system and the ring yoke with teeth supporting the windings have the greatest efficiency and highest quality factor.

621.397.621.2 **2353**

The Modulation Characteristic of Cathode-Ray Tubes in Television.—R. B. Mackenzie. (*Brit. J. appl. Phys.*, Feb. 1952, Vol. 3, No. 2, pp. 54–58.) The ambiguity which arises in defining the modulation characteristic in terms of gamma is discussed. A method of eliminating this ambiguity is proposed which is partly empirical, but leads to a simple mathematical treatment.

621.397.621.2 : 535.623 **2354**

Color Television Reproducers.—H. R. Lubcke. (*J. Soc. Mot. Pict. Televis. Engrs*, Jan. 1952, Vol. 58, No. 1, pp. 22–27.) Description of a three-gun tube with a heterogeneous screen of special construction using phosphors with different response times. By altering the speed of traverse of the electron beams, different colours predominate in the composite-phosphor response.

621.397.621.2(73) **2355**

Cathode-Ray Tubes for Television Receivers in U.S.A.—G. G. Esculier. (*Onde élect.*, Jan. 1952, Vol. 32, No. 298, pp. 32–35.) A review of recent developments and methods adopted to ensure good quality in mass production.

621.397.645 : 621.385.4 **2356**

Coaxial Tetrode as a TV Amplifier at V.H.F. and U.H.F.—Preist. (See 2175.)

621.397.645.018.424 **2357**

Wide-Band Amplifiers with Stagger-Tuned Circuits.—de Vos. (See 2176.)

621.397.828 **2358**

F.C.C.'s Plan for Handling TVI.—G. S. Turner. (*QST*, Jan. 1952, Vol. 36, No. 1, pp. 22–24.)

TRANSMISSION

621.392.52 : 621.396.615.029.55/.62 : 621.397.828 **2359**

Practical Applications of Pi-Network Tank Circuits for TVI Reduction.—G. Grammer. (*QST*, Jan. 1952, Vol. 36, No. 1, pp. 10–15.. 106.)

621.396.216 **2360**

Total or Partial Suppression of a Modulation Sideband.—J. Oswald. (*Câbles & Transmission, Paris*, April 1952, Vol. 6, No. 2, pp. 165–173.) Analysis of the characteristics of the envelope of an a.m. signal when the carrier wave and one sideband are suppressed. A definition is given of the mean and the maximum degree of modulation of a stationary aleatory signal with a limited spectrum and Gaussian distribution. Passage of a modulated wave through a filter is considered, the theory showing the existence of the two components in quadrature which characterize the response of an arbitrary linear network to a modulated signal. The probability law of the signal envelope and the degree of modulation are slightly modified by the suppression of a sideband, so that a compression of the envelope levels results. The theory is applied to vestigial-sideband transmission of a television signal.

621.396.61 **2361**

The New Paris-Villebon 100-kW Transmitter, Foremost in the International Technical Field.—P. A. François. (*TSF et TV*, Jan. 1952, Vol. 28, No. 279, pp. 21–24.) Special features of the transmitter, type monobloc TH755, are described, including the gravity-feed water cooling system for the valves, the heat dissipation of which causes the water to boil. The transmitter is modulated at high power and radiates on 1·07 Mc/s from an aerial common to a 150-kW transmission on 863 kc/s. Overall efficiency unmodulated is 53%.

621.396.615.016.22 **2362**

Curves of Equal H.F. Power of a Transmitter.—P. Mourmant. (*Radio franç.*, Jan. 1952, No. 1, pp. 1–10.) The family of equal-h.f.-power curves for a transmitter, expressed in terms of load impedance, give to the 'matching range' (595 and 2097 of 1948) a quite rigorous character. The matching ranges of ideal and of practical matching quadripoles are discussed, with particular reference to power tolerances and power output. Equipment for determining the values of the components of a matching quadripole is mentioned.

621.396.619.27 **2363**

Theory of the Cut-Off [Cowan] **Modulator with Capacitor Shunt.**—V. Belevitch. (*HF, Brussels*, 1952, Vol. 2, No. 1, pp. 1–6.) A method of analysis of circuits with periodically varying parameters is applied in determining the output function of a Cowan modulator, taking account of the capacitance of the rectifier. To a first approximation the effect of this capacitance is equal to that of the same capacitance in a circuit operating at a fixed frequency, the effect being calculated for an equivalent frequency that is a simple function of the input and carrier frequencies.

VALVES AND THERMIONICS

537.533.8 + 537.525.92 **2364**

Equilibrium between an Insulator emitting Secondary Electrons, a Space Charge, and an Enclosure at Constant Potential.—M. Barbier. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 61–67.) The order of magnitude of the surface potential of the insulator to be expected for various intensities of bombardment is calculated for the ideal cases in which the insulator and the collector electrode are two parallel planes or two concentric spheres, taking account of the space charge due to the secondary electrons.

621.314.7 **2365**

Progress in Transistor Technique.—E. H. Hungermann. (*Elektron Wiss. Tech.*, 1951/1952, Vol. 5, Nos. 13/14, pp. 429–439.) Summarizes publications and patents on methods of production of single crystals and on various treatments designed to improve performance. 39 references.

621.314.7 **2366**

Current Multiplication in the Type-A Transistor.—W. R. Sittner. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 448–454.) Discussion of the possibility of high current amplification arising from the trapping of holes in the barrier region of the collector which, due to the requirement of space-charge neutrality, leads to an enhanced electron concentration. Measurements on two transistors over the temperature range 237–298°K suggest trap densities of the order of 10^{13} cm^3, while trapping energies of 0·3 eV are derived both from the absolute magnitude of the trapping ratio (density of trapped holes/density of mobile holes) and from its temperature dependence.

621.314.7 **2367**

Transistor Forming Effects in *n*-Type Germanium.—L. B. Valdes. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 445–448.) An experimental study of the effects of electrically forming the collector of an *n*-type Ge transistor. It is concluded that a region of high-conductivity *p*-type material is produced under the collector, due either to lattice dislocations or to thermal diffusion of impurities. The enhanced current amplification so obtained is attributed to a *p*-*n* hook formed by a very small *n*-region immediately below the contact and surrounded by the larger *p*-region produced by forming.

621.383.27 **2368**

Research on Electron Multiplication and its Applications: Part 1.—D. Charles. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 34–60.) Nine-stage and ten-stage multipliers with crossed electric and magnetic fields were studied. The elementary and the exact theory of their operation are presented, the latter involving the calculation of the equipotential surfaces between two electrodes, and the deduction of the electron trajectories. The factors essential to satisfactory performance are discussed and enumerated. Experiments show that the sensitivity falls off rapidly below 3 000 Å, with a possible lower limit at about 2 200 Å. It is estimated that a photoelectric current of 10^{-14} A can be measured without much difficulty, the main limiting factor being background noise, which can be reduced by operation at low temperature. Comparative measurements at + 28°C and — 23°C with special equipment for determining the absolute limit of sensitivity gave results of 24×10^{-16} A and $1{\cdot}5 \times 10^{-16}$ A respectively.

621.383.27 **2369**

Electron Multiplier for the Ultraviolet Range down to to 1 450 Å.—V. Schwetzoff, S. Robin & B. Vodar. (*C. R. Acad. Sci., Paris*, 21st Jan. 1952, Vol. 234, No. 4, pp. 426–428.)

621.385 : 537.525.92 **2370**

Three Elementary Cases of the Expansion of Space-Charge Clouds.—H. Kleinwächter. (*Funk u. Ton*, Jan. 1952, Vol. 6, No. 1, pp. 25–28.) Solutions are obtained for the rate of expansion of spherical, cylindrical, and plate-shaped space-charge clouds.

621.385 : 537.525.92 **2371**

Space-Charge-Wave Propagation in a Cylindrical Electron Beam of Finite Lateral Extension.—P. Parzen. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 215–219.) The influence of valve configuration is investigated theoretically; the case of the complete-space-charge diode is discussed in detail. Results agree with the experimental findings of Cutler & Quate (1274 of 1951.).

621.385.029.6 **2372**

Noise in Transit-Time Valves.—W. Kleen. (*Frequenz*, Feb. 1952, Vol. 6, No. 2, pp. 45–50.) The noise figure of a klystron or a travelling-wave valve is governed primarily by the transit time of the electron beam. Fluctuations of density and velocity in the plane of the first accelerating electrode cause two space-charge waves of slightly different phase between this electrode and the h.f. input. Interaction of these waves gives rise to unwanted periodic components in the output. There is an optimum spacing of h.f. input and cathode for which the noise figure is a minimum. Experimental results confirm the theory.

621.385.029.6 **2373**

An Internal-Feedback Traveling-Wave-Tube Oscillator.—E. M. T. Jones. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 478–482.) Theory is presented which neglects space-charge effects. Experimental results for an oscillator using a helix as the interaction structure are in good agreement with the theory.

621.385.029.6 **2374**

Optimum Amplification in the Travelling-Wave Valve with Helix.—J. Labus. (*Arch. elekt. Übertragung*, Jan. 1952, Vol. 6, No. 1, pp. 1–5.) A formula is developed giving the valve amplification in terms of helix dimensions and operating parameters, and conditions are investigated for optimum value of amplification. Two cases are distinguished: (*a*) beam current related to helix potential by the space-charge law (helix at anode potential), (*b*) beam current adjustable independently of helix potential. Higher amplification can be obtained with lower power consumption in the latter case. Amplification is proportional to the cube root of wavelength times beam power, and decreases with increasing pitch of helix; it depends also on the ratio between the diameters of beam and helix.

621.385.029.6 : 621.392.5 **2375**

The Delay Line as a Component of Valves.—W. Kleen & W. Ruppel. (*Arch. Elektrotech.*, 1952, Vol. 40, No. 5, pp. 280–304.) Basic formulae and properties of various types of delay line for travelling-wave valves are reviewed, and particular types of homogeneous and inhomogeneous lines are discussed, the latter type defined as consisting of a series of quadripole elements of finite axial extent and being treated in detail from the point of view of their resemblance to filters. Equivalent circuits are derived and solutions of the field equations are obtained for plane-parallel and cylindrical types of line and for the circular type of the travelling-wave magnetron, results being presented graphically to facilitate approximate determination of characteristics.

621.385.029.6.012.6 (083.5) **2376**

Langmuir's ξ, η Tables for the Exponential Region of the I_a-V_a Characteristic.—G. Diemer & H. Dijkgraaf. (*Philips Res. Rep.*, Feb. 1952, Vol. 7, No. 1, pp. 45–53.) "The inter-electrode distances of modern microwave diodes and triodes are often so small that the normal operating point lies in the exponential part of the characteristic. A set of ξ, η tables with the voltage gradient at the anode as parameter is given from which the potential distribution in such cases can be derived."

621.385.029.64 **2377**

Optimum Geometry of Microwave Amplifier Valves.—G. Diemer & K. Rodenhuis. (*Philips Res. Rep.*, Feb. 1952, Vol. 7, No. 1, pp. 36–44.) On the basis of van der Ziel & Knol's theory of feedback amplifiers (1646 of 1950) it is shown that for u.h.f. amplifier valves the upper frequency limit for the amplification is the highest possible if the electrode areas are so chosen that the useful capacitance equals the unavoidable parasitic capacitance. For optimum gain-bandwidth products the useful capacitance should be somewhat higher.

621.385.032.216 **2378**

The Growth and Properties of Cathode Interface Layers in Receiving Valves.—M. R. Child. (*P.O. elect. Engrs' J.*, Jan. 1952, Vol. 44, Part 4, pp. 176–178.) A resistive interface layer tends to grow between the oxide cathode matrix and the supporting core and results in a reduction of mutual conductance by negative feedback. The effects of temperature variation, silicon concentration, and cathode current on the rate of increase of interface resistance are described.

621.385.032.216 **2379**

Emission and Crystal Size of Oxide-Coated Cathodes.—J. Shimazu. (*J. phys. Soc. Japan*, Nov./Dec. 1951, Vol. 6, No. 6, pp. 479–485.)

621.385.032.216 **2380**

Anomalous Distribution of the Velocities of the Electrons emitted by a Pulsed Oxide Cathode.—R. Loosjes & C. G. J. Jansen. (*Le Vide*, Jan. 1952, Vol. 7, No. 37, pp. 1131–1135.) Continuation of work reported in 2067 of 1950 (Loosjes, Vink & Jansen) and back references.

621.385.032.216.011.22 **2381**

Contribution to the Study of the Resistance through the Oxide-Cathode Layer.—C. Biguenet. (*Le Vide*, Jan. 1952, Vol. 7, No. 37, pp. 1123–1130.) Cathodes with coatings of various thicknesses were prepared, and curves of resistance/thickness at various temperatures obtained. By extrapolation, a value of 0·8Ω was obtained for the interface resistance, which was practically independent of cathode temperature and emission current. The coating resistance was greater (of the order of a few ohms), varying according to thickness, temperature and mode of operation of the cathode. Explanations are advanced to account for the different experimental results obtained by various workers.

621.385.2 **2382**

The Electric Field in Diodes and the Transit Time of Electrons as Functions of Current.—P. L. Copeland & D. N. Eggenberger. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 280–286.) Equations equally applicable to parallel-plane, coaxial-cylinder and concentric-sphere configurations are developed, giving approximate solutions for potential distribution and transit time. Calculations for the coaxial-cylinder arrangement indicate that the field at the cathode changes only very slowly as a function of the geometry; hence formulae derived for the parallel-plane arrangement are applicable with only small corrections to coaxial-cylinder arrangements. Functions used in the calculations are tabulated.

621.385.2 **2383**

Space Charge and Transit Time Considerations in Planar Diodes for Relativistic Velocities.—H. F. Ivey. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 208–211.) "The usual Child-Langmuir equation is extended to the case of relativistic velocities for a diode with parallel plane electrodes. The solutions obtained are valid for any value of the accelerating voltage. As the variation of mass with velocity becomes important, the exponent giving the dependence of current density on anode voltage becomes less than three-halves, and for very large values of accelerating potential approaches unity. The variation with anode voltage of the transit time, both for the space-charge-free case and for the space-charge-limited case, has been calculated. The potential distribution across the diode is also discussed."

621.385.2 : 537.525.92 **2384**

Space-Charge-Limited Currents between Inclined Plane Electrodes.—H. F. Ivey. (*J. appl. Phys.*, Feb. 1952, Vol. 23, No. 2, pp. 240–249.) The method of investigation described by Walker (1275 of 1951) is used; solutions are found for the potential distribution, space-charge characteristic ('perveance'), particle trajectories and transit time under space-charge-limited conditions for diodes with angles up to 180° between the electrodes.

621.385.2 : 546.289 **2385**

Germanium Diode Experience.—(*Radio & Televis. News, Radio-Electronic Engng Section*, Jan. 1952, Vol. 47, No. 1, pp. 20–21.) Account of a preliminary study of the performance of the 16 000 Ge diodes in the N.B.S. Eastern Automatic Computer (SEAC) during the latter half of 1950.

621.385.2 : 621.3.015.3 **2386**

Transients in Valves.—H. Fack. (*Frequenz*, Feb. 1952, Vol. 6, No. 2, pp. 33–37.) A mathematical treatment of the response of a plane diode to a transient input. A simple equivalent circuit is derived which comprises resistance and inductance in parallel with a capacitance, the inductance representing electron inertia.

621.385.38 **2387**

Statistical Nature and Physical Concepts of Thyratron Deionization Time.—H. A. Romanowitz & W. G. Dow. (*Trans. Amer. Inst. elect. Engrs*, 1950, Vol. 69, Part I, pp. 368–379.)

621.385.5 **2388**

Circuits of the Balitron Tube.—N. Z. Ballantyne. (*Radio & Televis. News, Radio Electronic Engng Section*, Jan. 1952, Vol. 47, No. 1, pp. 6–8 .. 31.) A beam-deflection valve with a stable negative-resistance type of characteristic is obtained by slight modifications of the arrangement and shape of the electrodes of the positive-resistance type of the 'balitron' valve. Suitable oscillator and frequency-converter circuits using the modified valve are described.

621.385.832 : 621.318.572 **2389**

Ribbon Beam Valves: Principles and some Applications.—G. Piétri. (*Le Vide*, Jan. 1952, Vol. 7, No. 37, pp. 1113–1122.) Theory and brief descriptions of a 10-way switching valve and a pulse generator.

621.396.615.142 **2390**

Retarding-Field Oscillators.—J. J. Ebers. (*Proc. Inst. Radio Engrs*, Feb. 1952, Vol. 40, No. 2, pp. 138–145.) Velocity-variation oscillations are analysed mathematically and distinguished from the Barkhausen-Kurz type. Electron paths and the processes of bunching and drifting are discussed and measurements of power output and efficiency are given for an experimental planar-type oscillator.

621.396.615.142 **2391**

The Variation of the H.F. Power of a Drift-Space Valve with the D.C. Power.—R. Gebauer & H. Kosmahl. (*Z. angew. Phys.*, Dec. 1951, Vol. 3, No. 12, pp. 449–452.) Investigations reported in 1036 of 1951 are continued, measurements being made on a type 0+ valve (input-gap transit angle $\approx \pi$). The value of h.f. power as a function of d.c. power varies first quadratically, then linearly, and, if the beam current is made sufficiently high, finally passes through a maximum and falls to zero. The rising part of the curve is explained on the basis of the variation of efficiency with modulation depth, which in turn depends on beam current; the falling part is attributed to the effect of space charge in increasing transit time and impairing focusing.

621.396.615.142.2 **2392**

A Type of Reflex Klystron with Fixed Load and Wide Frequency Range.—J. Laborderie. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 68–74.) The design of the

Type-KR 142 valve is described. It covers the range 8·45–10·30 cm and has a power of at least 50 mW over the whole range, with a peak power of about 250 mW around 8·8 cm and bandwidth of about 20 Mc/s over most of the range. The load is a 75-Ω coaxial line terminated by its characteristic impedance. Careful dimensioning and assembly of the coupling loop is essential.

621.396.615.142.2 **2393**
Recent Developments in High-Power Klystron Amplifiers.—V. Learned & C. Veronda. (*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 465–469.) The three types of beam focusing used in modern high-power klystrons are discussed, with illustrations of the SAS-28 c.w. 250-W 2·6-kMc/s valve using ion focusing, the SAL-39 valve using space-charge focusing and giving a pulsed output of 20 kW at 1 kMc/s and 1% duty cycle, and the SAC-33 valve with a magnetically focused beam, giving 500 W c.w. at 5 kMc/s. The maximum power output, efficiency, gain, bandwidth, tuning means and temperature compensation of modern klystrons are reviewed.

621.396.615.142.2.029.65 **2394**
Development of a Demountable Klystron for the Generation of Millimetre Waves.—M. Matricon. (*Rev. tech. Comp. franç. Thomson-Houston*, Dec. 1951, No. 16, pp. 45–52.) Details of a reflex klystron designed for use with the spectroscope described in 2260 above. It has a beam current of 10–20 mA and covers several bands in the range 20–28 kMc/s. With some parts changed, bands in the range 28–38 kMc/s are covered.

621.396.622.63 + 621.383.5 **2395**
Tabulated Data of the Most Important Commercially Available Rectifying Crystals (Crystal Valves).—O. Stürzinger. (*Bull. schweiz. elektrotech. Ver.*, 26th Jan. 1952, Vol. 43, No. 2, pp. 41–47. In German.) The data tabulated include information on frequency range, temperature range, physical form, applications and alternative types, as well as the characteristic parameters. Photocells are included.

621.396.822 : 621.385 **2396**
Space-Charge Smoothing of Microwave Shot Noise in Electron Beams.—F. N. H. Robinson. (*Phil. Mag.*, Jan. 1952, Vol. 43, No. 336, pp. 51–62.) "A theoretical analysis is given which takes account of both space-charge interaction between electrons and the multi-valued nature of the flow due to the Maxwellian distribution of initial velocities. The theory is in agreement with the experimental results of Cutler and Quate [1274 of 1951] and makes possible a coherent account of shot noise at all frequencies."

621.396.822 : 621.385 **2397**
The Calculation of Fluctuation Noise in Interelectrode Spaces without Transverse Magnetic Field.—G. Convert. (*Ann. Radioélect.*, Jan. 1952, Vol. 7, No. 27, pp. 10–19.) The assumptions usually made in calculating the noise due to fluctuations of current or velocity in an electron beam are reviewed. Approximate formulae are developed for the noise at a point in a beam with and without space charge. When account is taken of the special conditions which exist in klystrons and travelling-wave valves, the formulae can also be applied to these valves, particularly to low-noise types.

621.383 **2398**
Die Photozellen. [Book Review]—P. Görlich. Publishers: Akad. Verl.-Ges., Leipzig, 1951, 288 pp., 19.80 DM. (*Z. angew. Phys.*, Feb. 1952, Vol. 4, No. 2, p. 79.) A comprehensive work covering theory, methods of manufacture and properties of commercially available photocells.

MISCELLANEOUS

001.891 : 621.396 **2399**
Radio Research.—(*Engineering, Lond.*, 28th Dec. 1951, Vol. 172, No. 4483, pp. 815–816.) An account of the work done by the Radio Research Board in 1950, based on the Director's report (Radio Research, 1950, H.M. Stationery Office, 1s. 9d.).

06.091 Clerk Maxwell **2400**
The Inaugural Clerk Maxwell Memorial Lecture.—G. W. O. Howe. (*J. Brit. Instn Radio Engrs*, Dec. 1951, Vol. 11, No. 12, pp. 545–554.) An account of Clerk Maxwell's life and personality, delivered in the Clerk Maxwell lecture theatre of the Cavendish Laboratory, Cambridge, 24th August 1951, at the conclusion of the Television-Engineering session of the Radio Convention of the British Institution of Radio Engineers.

061.4 : [621.396.6 + 621.317.7 + 621.38 **2401**
Physical Society's Exhibition 1952.—(*Electrician*, 4th April 1952, Vol. 148, No. 3851, pp. 1071–1078; *Elect. Times*, 3rd April 1952, Vol. 121, No. 3152, pp. 615–621; *Elect. Rev., Lond.*, 4th April 1952, Vol. 150, No. 3880, pp. 703–711; *Metal Ind., Lond.*, 4th April 1952, Vol. 80, No. 14, pp. 272–274; *Nature, Lond.*, 17th May 1952, Vol. 169, No. 4307, pp. 816–818; *Engineer, Lond.*, 4th & 11th April 1952, Vol. 193, Nos. 5019 & 5020, pp. 463–465 & 495–499; *Engineering, Lond.*, 28th March, 4th & 11th April 1952, Vol. 173, Nos. 4496–4498, pp. 405–406, 425–427 & 470–472.) Reports with descriptions of some of the exhibits.

061.4 : 621.396 **2402**
Around the Stands at the Fifth Annual R.S.G.B. Amateur-Radio Exhibition.—(*R.S.G.B. Bull.*, Jan. 1952, Vol. 27, No. 7, pp. 312–315.) A brief account of exhibits by the Services and the Radio Industry.

061.4 : 621.396.6 **2403**
The National Components Exhibition.—Gilloux. (See 2154.)

621.396 **2404**
Radio Progress during 1951.—(*Proc. Inst. Radio Engrs*, April 1952, Vol. 40, No. 4, pp. 388–439.) A comprehensive review with a world-wide range of 779 references. Section headings are: (*a*) electron tubes and semiconductors; (*b*) audio techniques; (*c*) electroacoustics; (*d*) information theory and modulation systems; (*e*) circuit theory; (*f*) radio transmitters; (*g*) receivers; (*h*) video techniques; (*i*) television systems; (*j*) facsimile; (*k*) vehicular communications; (*l*) navigation aids; (*m*) wave propagation; (*n*) antennas, waveguides, and transmission lines; (*o*) industrial electronics; (*p*) electronic computers; (*q*) quality control; (*r*) instrumentation; (*s*) piezoelectricity; (*t*) symbols and abbreviations.

621.396.029.63/.64 **2405**
Les Hyperfréquences (Tubes et appareils de mesure. Applications aux télécommunications et au radar). [Book Review]—J. Voge. Publishers: Eyrolles, Paris, 317 pp., 1980 fr. (*Rev. gén. Élect.*, Dec. 1951, Vol. 60 No. 12, p. 473.) "Specialists will fully appreciate the, novelty and completeness of this book . . . which is also of outstanding interest to university students . . . and in general to all those who need to acquire or complete a knowledge of very-short-wave technique (from about 50 to 1 cm wavelength)."

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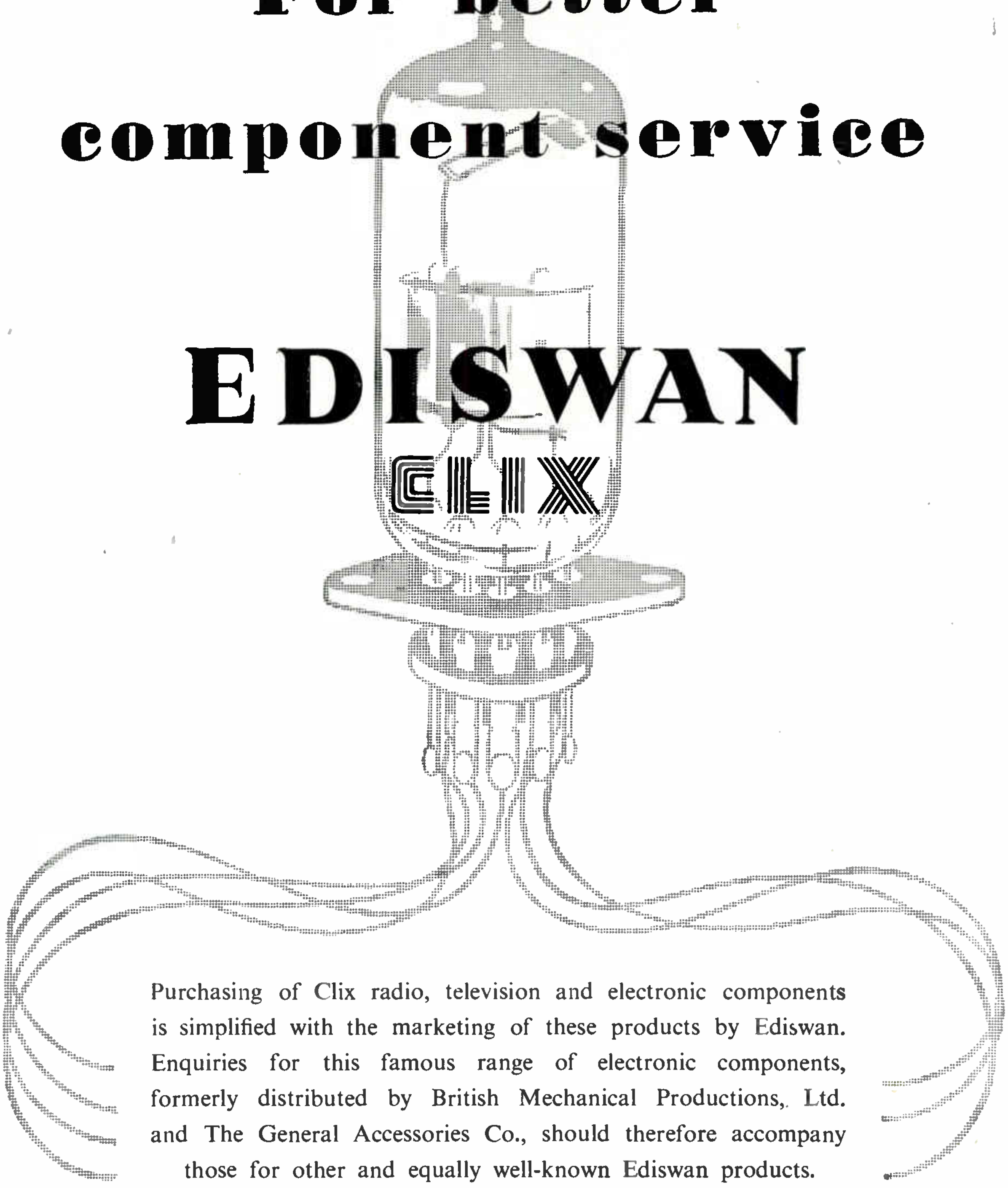

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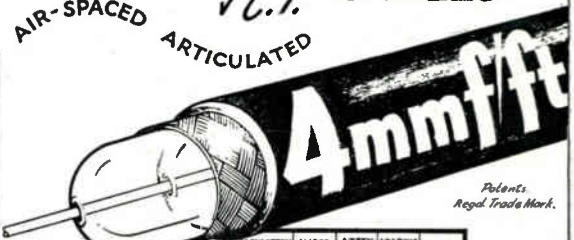
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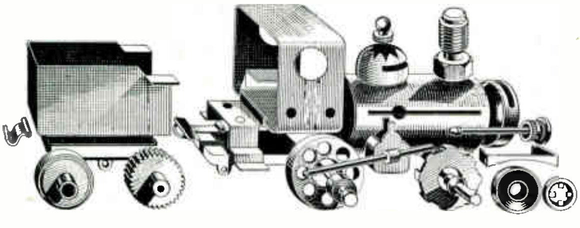

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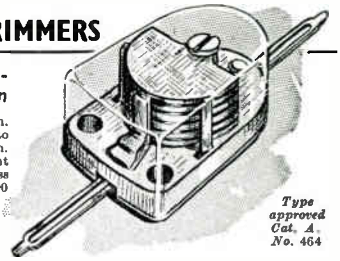

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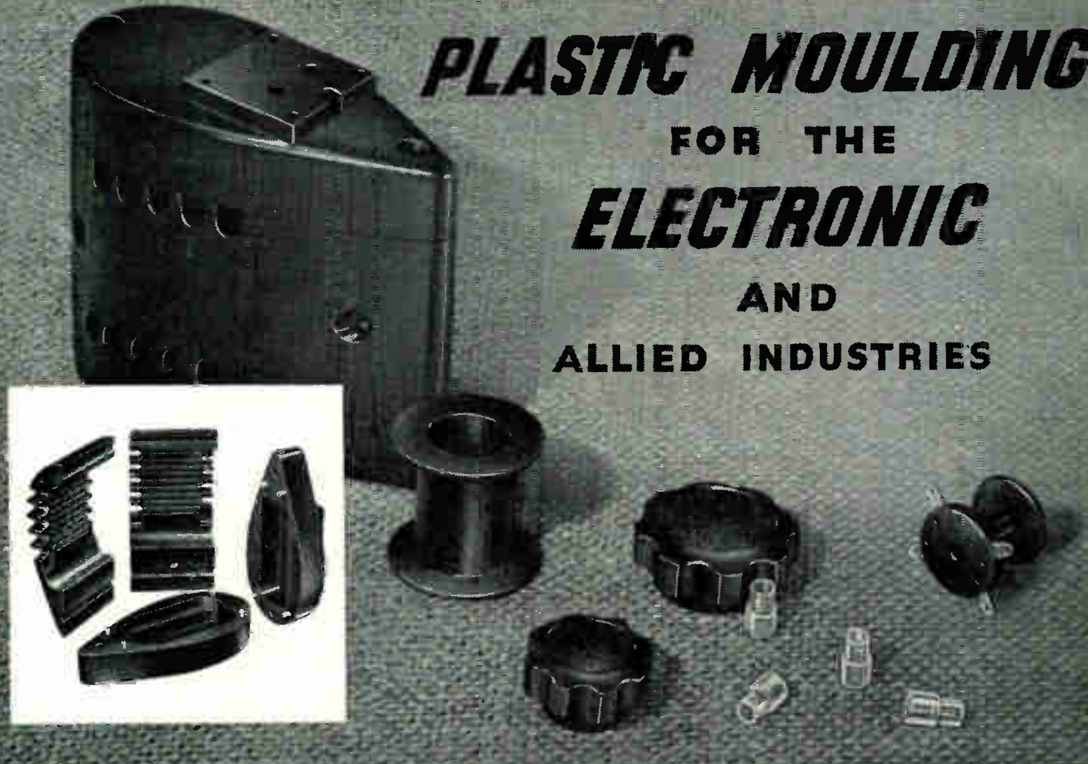

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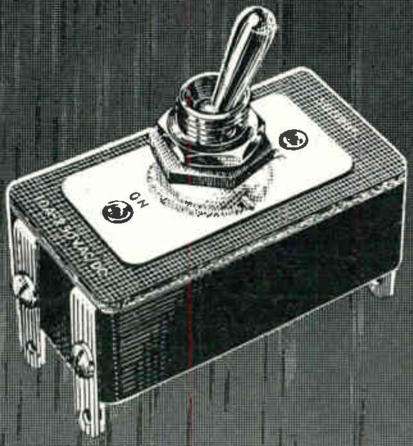
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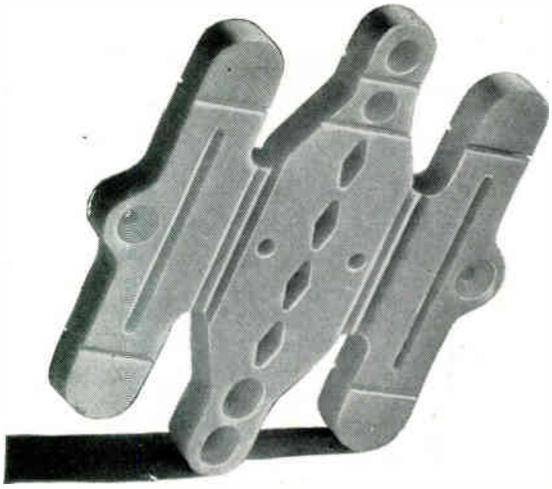

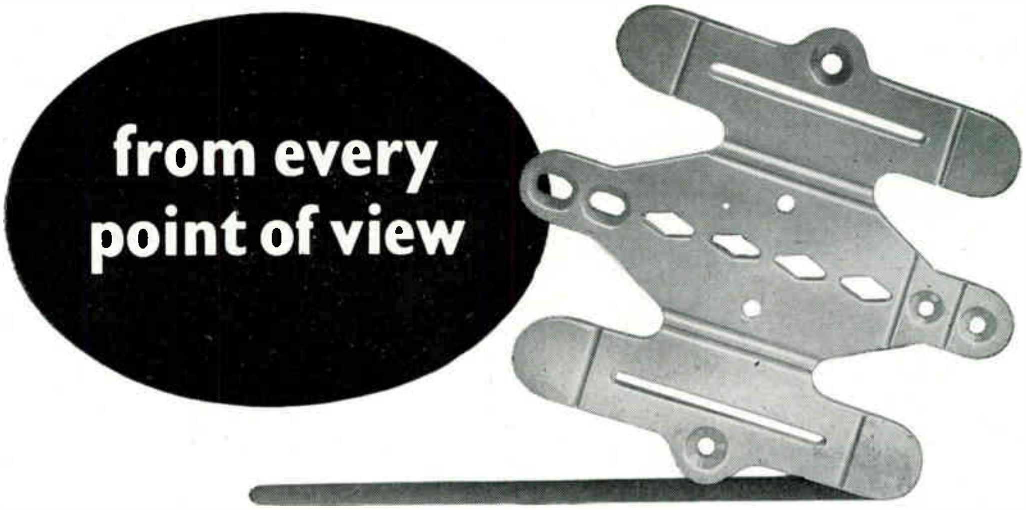
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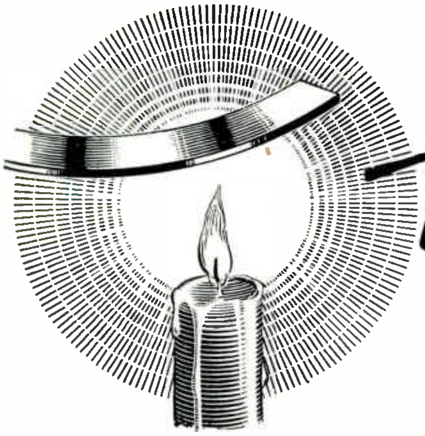

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World Radio History

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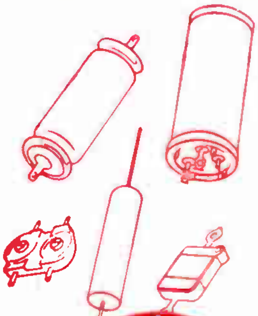

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Due to the unique combination of characteristics inherent in the design of Carpenter Polarized Relays, they are providing practical solutions to many problems in scientific research and electrical engineering.

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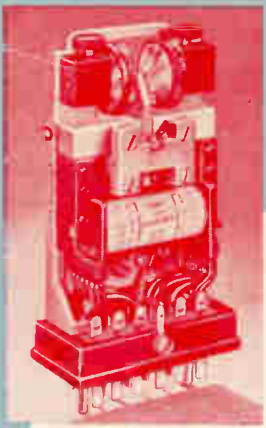
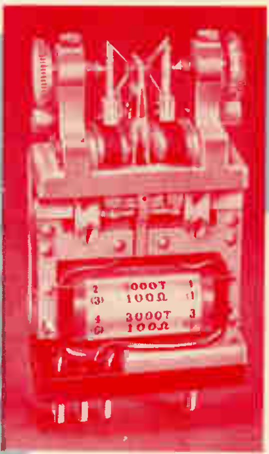
The relay operates successfully direct from valves, the barrier layer type of photo cell, and also from thermo-couples. In the amplification of very weak D.C. signals, the relay's almost perfect contact performance makes it suitable for the inversion of the D.C. into A.C. to permit electronic amplification. *Three Main Types are available.*



TYPE 3

has maximum speed: maximum contact pressure and sensitivity: minimum transit time. Dimensions: 120 mm. × 61 mm. × 33 mm. Weight 625 gm.



TYPE 4

has longer contact travel than the Type 3, with medium speed and sensitivity. Dimensions: 89 mm. × 56 mm. × 25 mm. Weight 368 gm.

TYPE 5

miniature relay with outstanding performance for size. Rugged design of exceptional thermal stability. Dimensions: 56 mm. × 37 mm. × 19mm. Weight 140 gm.

Relays can be supplied with either solder tag or plug base.

Note.—All dimensions quoted exclude connection terminals.

Complete details and performance data will be willingly supplied on request by the Manufacturers:

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