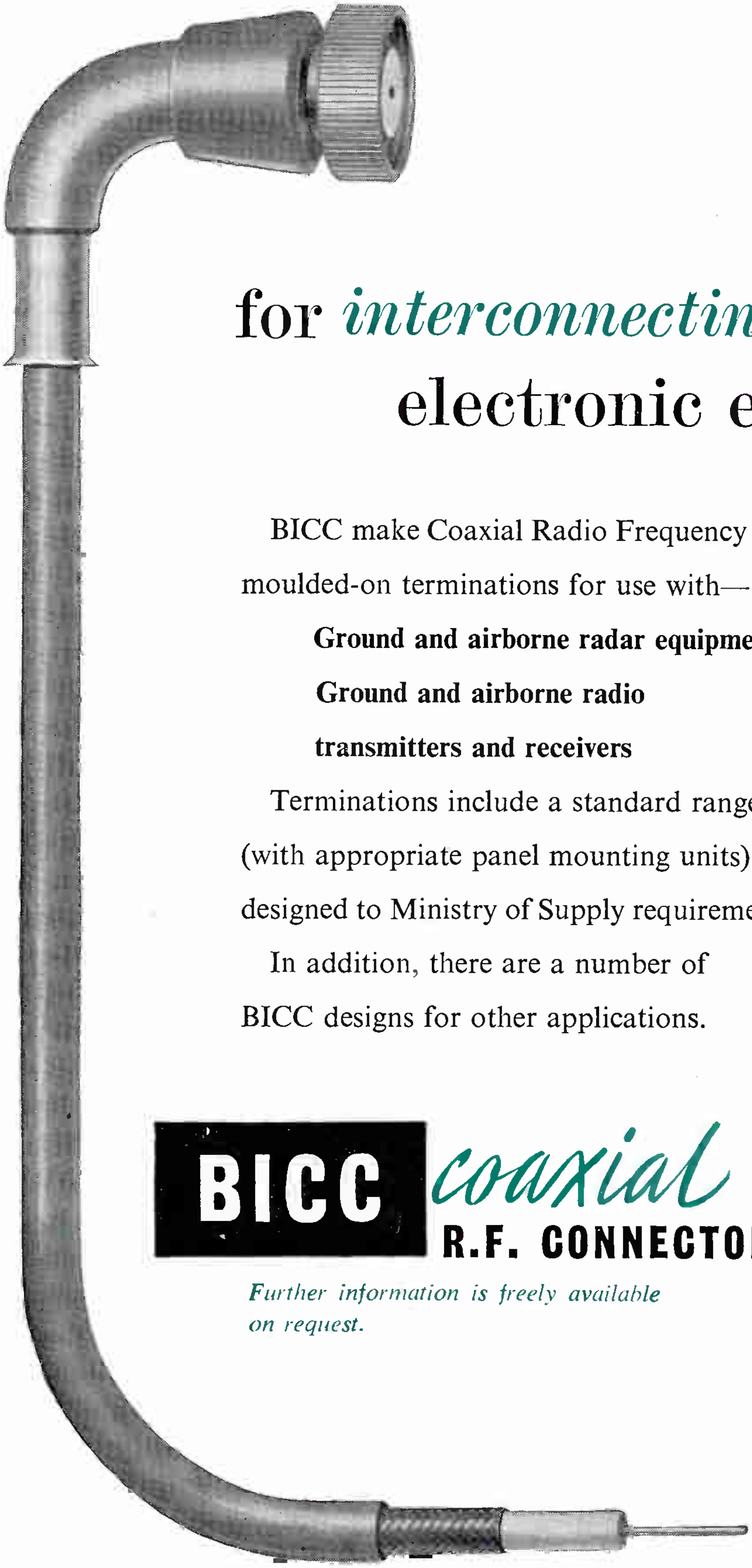
ELECTRONIC & RADIO ENGINEER

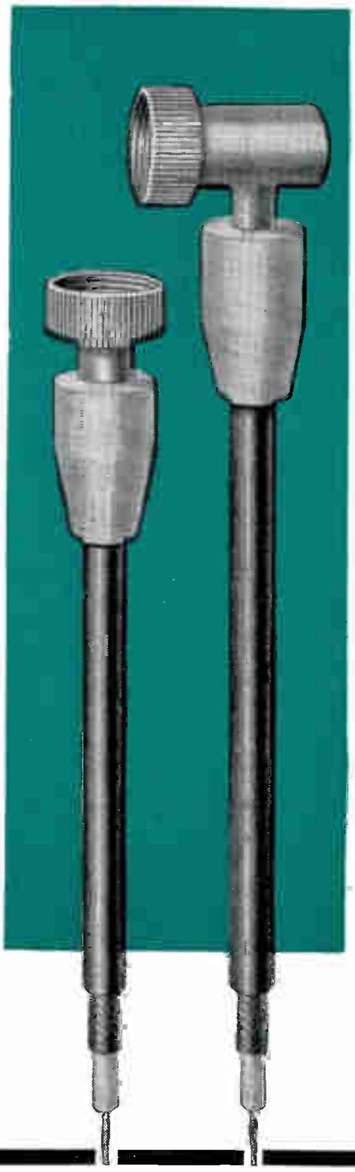
Incorporating **WIRELESS ENGINEER**

In this issue

Three shillings
and sixpence

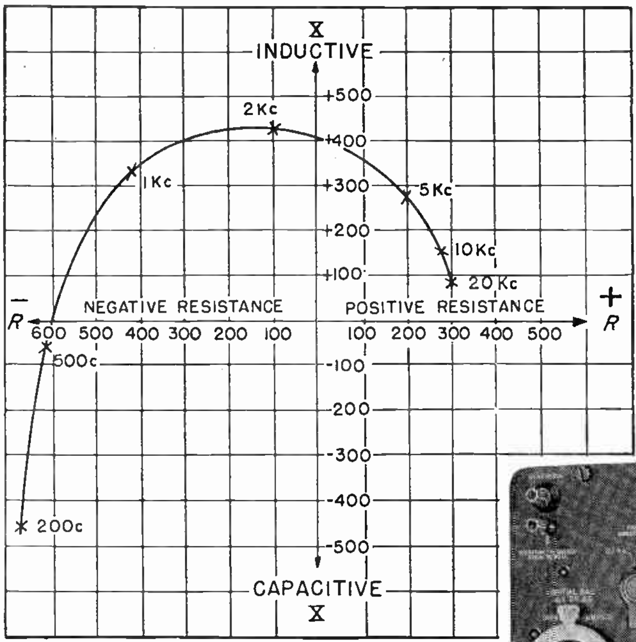
SEPTEMBER 1958 Vol 35 *new series* No 9



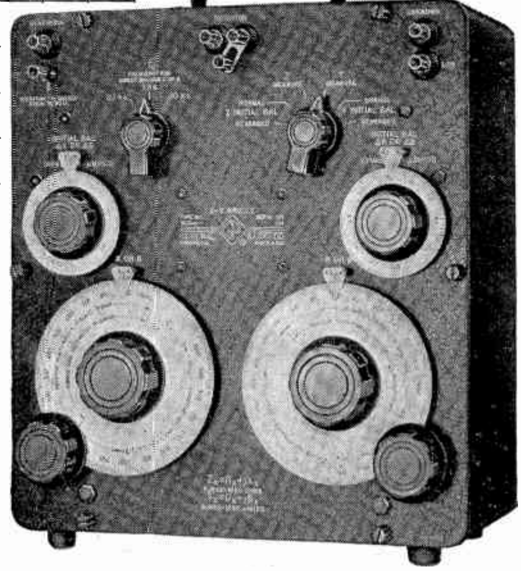


TRADE MARK

Type 1603-A Z-Y Bridge



Impedance of Feedback Circuit ... illustrates ability of the Z-Y Bridge to measure any impedance; quadrature components may be positive or negative, real or imaginary.



Measures Any Impedance...

- From 0 to ∞ ohms
- Positive or Negative
- At Any Phase Angle
- Over 20-c/s to 20-kc Range

The General Radio Z-Y Bridge measures impedances from short circuit to open circuit, at small or large phase angle. Quadrature components, R & X or G & B, are measured directly at calibrated 100 c, 1-kc, and 10-kc bridge positions. Basic accuracy is 1% over most of this range.

The ability to measure impedances of any magnitude accurately with one instrument is an extremely valuable asset in many measurement situations. The Z-Y Bridge can be used for measuring conductivity of liquids, in dielectric cells as readily as it can be used for R-L-C component measurements in the laboratory or production-test department. It will measure open-circuit and short-circuit transformer parameters . . . impedances of batteries and electrolytic capacitors . . . characteristics of audio-transmission networks . . . impedance of electro-acoustic transducers . . . Q and resonant frequency of chokes . . . and impedances of feedback loops, since negative real parameters can be directly measured.

The Bridge also can be used to determine cable-fault locations and circular-arc plots of liquids or solids having lossy polarizations in the audio-frequency range. These are but a few of the countless applications for this unique and versatile device. *You name it — this Z-Y bridge can probably measure it.*

For complete information request a copy of the current "G.R." Catalogue "O" (258 pages), where data is given on pages 34/35.

Impedance and Admittance Range

R: ±1000 ohms	G: ±1000 μmhos
X: ±1000 ohms	B: ±1000 μmhos

Accuracy

R or G: $\pm(1\% + (2 \text{ ohm or } 2\ \mu\text{mho}))$
X or B: $\pm(1\% + (\frac{2f_0}{f} \text{ ohm or } \frac{2f}{f_0}\ \mu\text{mho}))$

f is operating frequency, f_0 is frequency setting of panel selector switch

Impedances of less than 100Ω or (100 μmhos) can be measured on "Initial Balance" dials with considerably greater accuracy—
R or G: $\pm(1\% + (0.2 \text{ ohm or } 0.2\ \mu\text{mho}))$
X or B: $\pm(1\% + (\frac{0.2f_0}{f} \text{ ohm or } \frac{0.2f}{f_0}\ \mu\text{mho}))$

Frequency Range—20 cycles to 20 kc

Maximum Applied Voltage
130 volts, rms on bridge;
less than 32v on unknown

Accessories Recommended
"G.R." Type 1210-B Unit R-C Oscillator and
"G.R." Type 1212-A Unit Null Detector

Accessories Supplied
2 Shielded Cables for generator and detector

Dimensions—12½" x 13½" x 8½"

Net Weight—21½ lbs.

Type 1603-A Z-Y Bridge £222

Claude Lyons Ltd.

76 Oldhall Street Liverpool 3, Lancs.
Telephone: Central 4641/2
Valley Works, Ware Road, Hoddesdon, Herts.
Telephone: HODdesdon 3007-8-9

CL.45

W. BRYAN Savage

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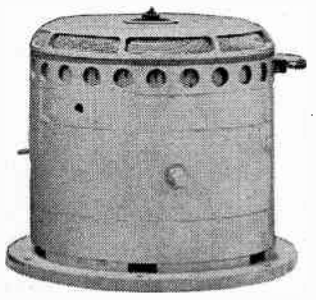
Vibrators



for fatigue testing

Guided missiles, aircraft and all forms of industrial components . . .

TYPE V1001. This vibrator is of the moving coil type with a wound field magnet energised from an external source of direct current. It is continuously rated and therefore suitable for extended fatigue testing as well as for intermittent use in research, development and production work.

The useful frequency range is up to 5 kc/s.

The design of the vibrator is such that the electrical impedance of the "speech coil" shows only a slight rise at the higher frequencies and this obviates the need for frequent output transformer tap changing as the frequency is varied.

SPECIFICATION

Height	20in.
Dia. of base	35in.
Fixing holes, No. and size as required, on a 33¼in. P.C.D.	
Weight	1 ton 15 cwt. 10 lb.
Table	18in. dia.
Thrust	3,000lb. plc.
Excursion	0.500in. max.
Max. permitted acceleration	100 "G"
Direct current force factor	36lb./amp.
Max. input	64 amps (r.m.s.)
Moving coil blocked impedance	1.5 ohms (est.)
D.C. resistance of moving coil	0.875 ohms.
Field coil current	4.6 amps.

Our Technical Dept. is always available to give assistance on any Vibration problems.

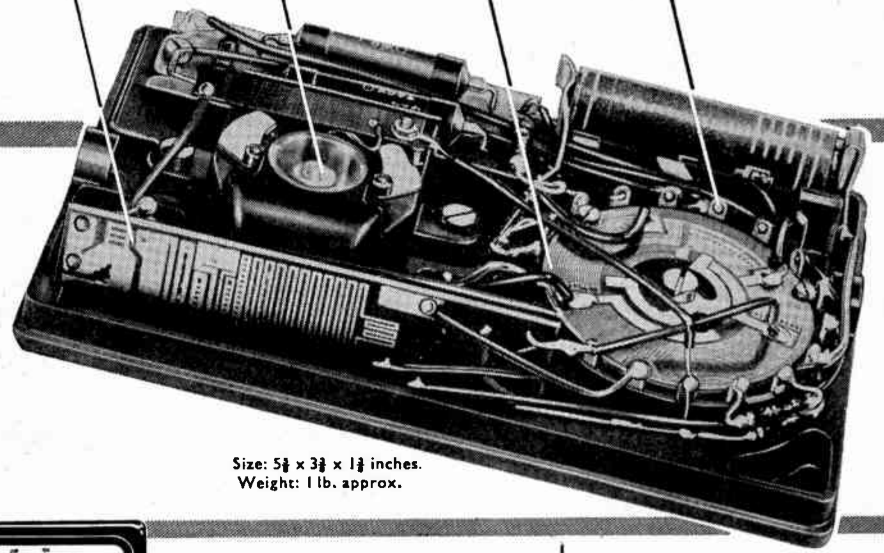
W. BRYAN SAVAGE LTD

17, Stratton Street, London, W.1
Telephone: GROsvenor 1926

DaS958ERE

Pocket size...
pocket wise!

This sturdy multi-range test meter is remarkable for the wide range of test facilities which have been so neatly incorporated. Full advantage has been taken of printed resistor techniques to produce a compact instrument of low weight.



Printed resistance panel for universal meter shunt.

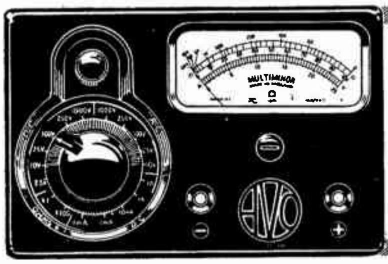
Composite printed resistors and auxiliary switch.

Meter movement is enclosed to give protection against the infiltration of dust.

Robust range switch similar to that used in the famous Avo-Meter. Eighteen fixed silver-plated contacts embedded in a ring of high-grade moulding material are swept by a double contact rotor arm.

Regd. Trade Mark.

Size: 5¾ x 3¾ x 1¾ inches.
Weight: 1 lb. approx.



7 D.C. Voltage Ranges: 0-1,000 V.

5 A.C. Voltage Ranges: 0-1,000 V.

5 D.C. Current Ranges: 0-1A

2 Resistance Ranges: 0-20,000Ω.
0-2MΩ.

Sensitivity:
10,000Ω/V on D.C. voltage ranges.
1,000Ω/V on A.C. voltage ranges.

Accuracy:
3% of full scale value on D.C.
4% of full scale value on A.C.

For a small additional charge, instruments can be supplied to a higher degree of accuracy.

THE
MULTIMINOR

19 Ranges · Single Knob Control · List Price: £9:10s.

Complete with test leads and clips. Leather case if required 32/6.

- **Write for fully descriptive leaflet.**

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VICtoria 3404 (9 lines)

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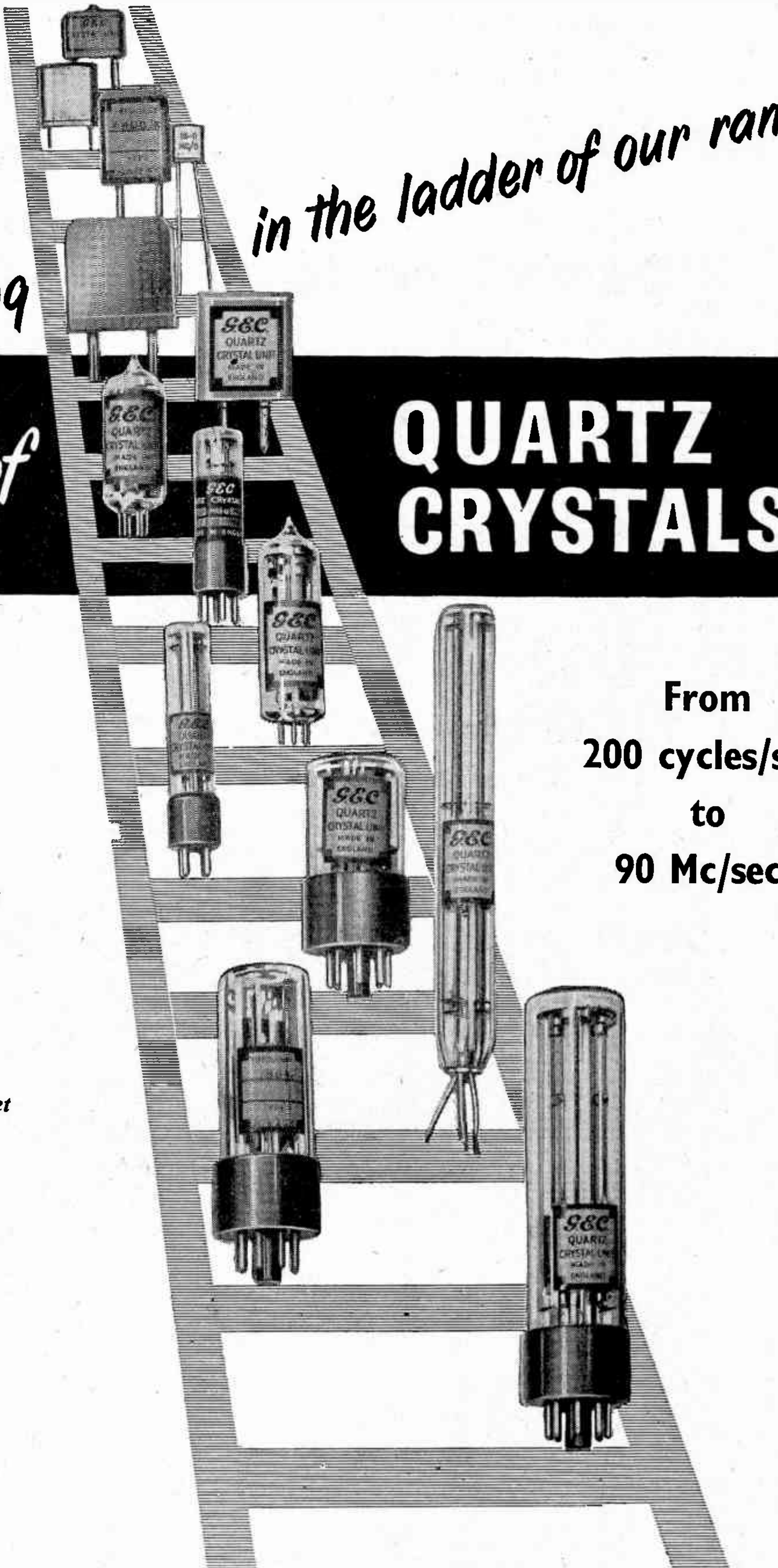

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From 200 cycles/sec. to 90 Mc/sec.

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

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

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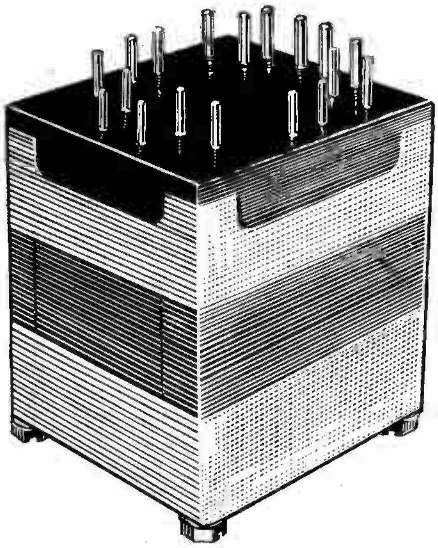


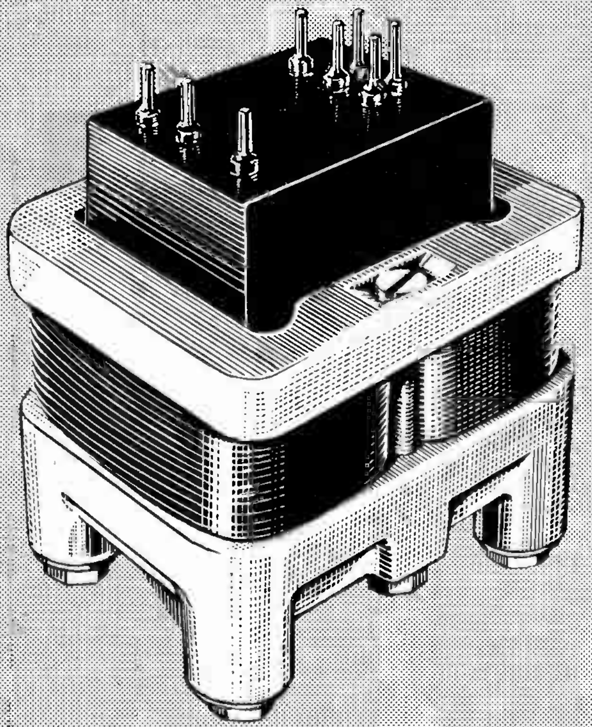
*R*120 *and R*130 *laminated series.*

FORTH SERIES

TRANSFORMERS AND CHOKES

smaller in size— lighter in weight

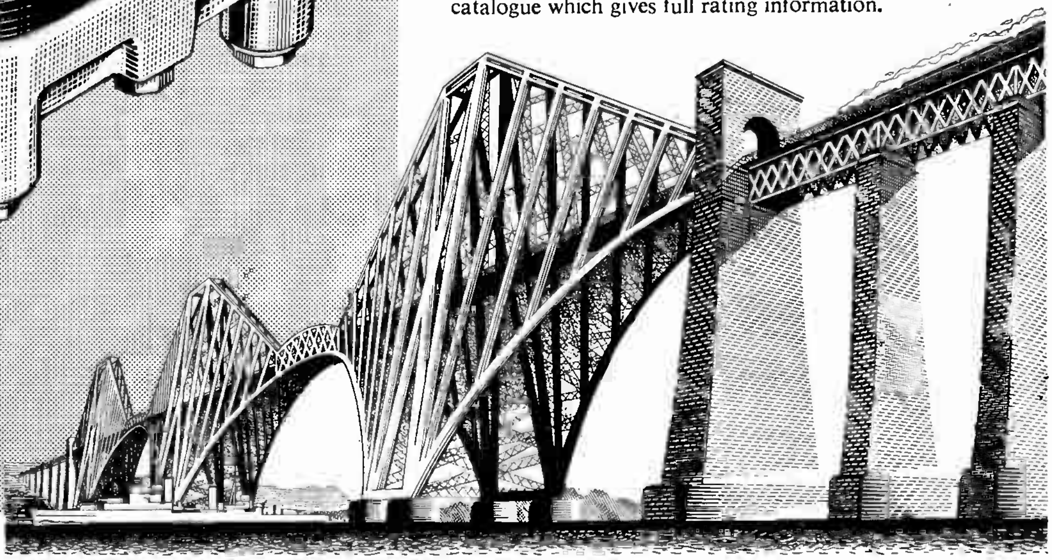
*R*200 *'C' Core series.*



FERRANTI

The new range of Ferranti Resin Cast Transformers and Chokes has been named after this famous Scottish landmark which represented a remarkable advance in engineering design when it was constructed over 60 years ago. To-day, the new techniques in manufacture and construction of 'C' Core Transformers have enabled Ferranti Ltd. to make a significant contribution to Electronic Engineering.

The Forth series components will have particular appeal to designers of airborne equipment since savings in weight and volume of up to one-third can be achieved over the resin cast and oil-filled units now available. Moreover, the quality requirements of the Joint Service Specification RCS.214 are met in every respect. Please write for a catalogue which gives full rating information.



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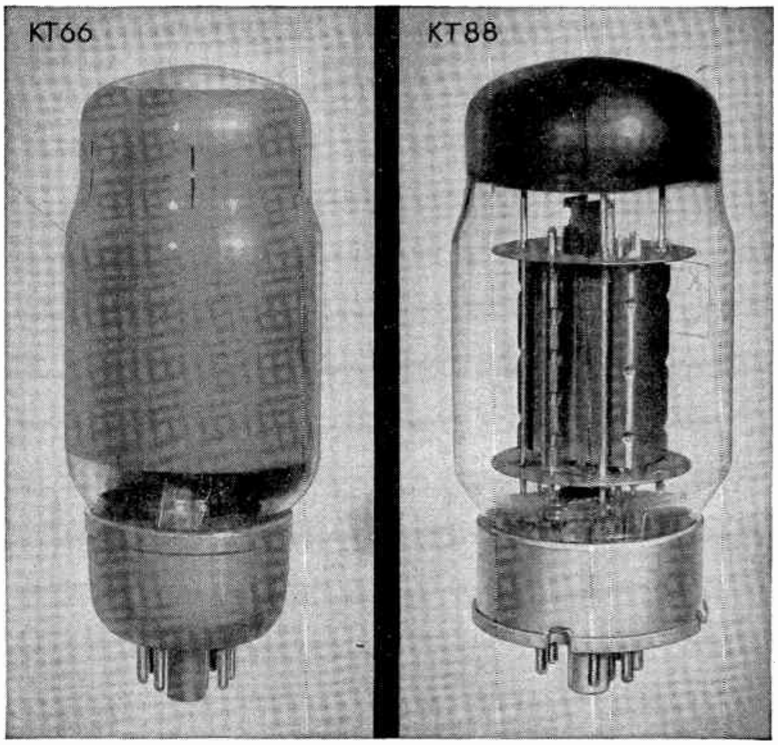


G.E.C.

VALVES

for reliability and power at audio frequencies

The KT 66 — still leading the world after 21 years

When the KT 66 was introduced in 1937, it was far ahead of its time. So far ahead that it still leads the world today. Over the years that have passed, many millions of these valves have been manufactured and the excellent design plus the quality of materials used have won a phenomenal reputation for long-lasting reliability. 12 valves, recently installed in multi-channel radio equipment, each completed 32,600 hours without failure. The KT 66 has been used in a number of well-known high quality audio amplifiers including the 'Williamson' and the 'Leak Point 1', designed for outputs of up to 50 watts.

	KT 66	KT 88
V_a (max.)	500	600 volts
V_{g2} (max.)	400	600 volts
p_a (max.)	25	35 watts
g_m	6.3	11 mA/V
Pout (AB1 push-pull fixed bias, U.L.)	50	100 watts
V_h	6.3	6.3 volts
I_h	1.27	1.8 amps
Price ... ... P.T.	17s. 6d. 6s. 10d.	£1 2s. 6d. —

The KT 88 — for even higher powers

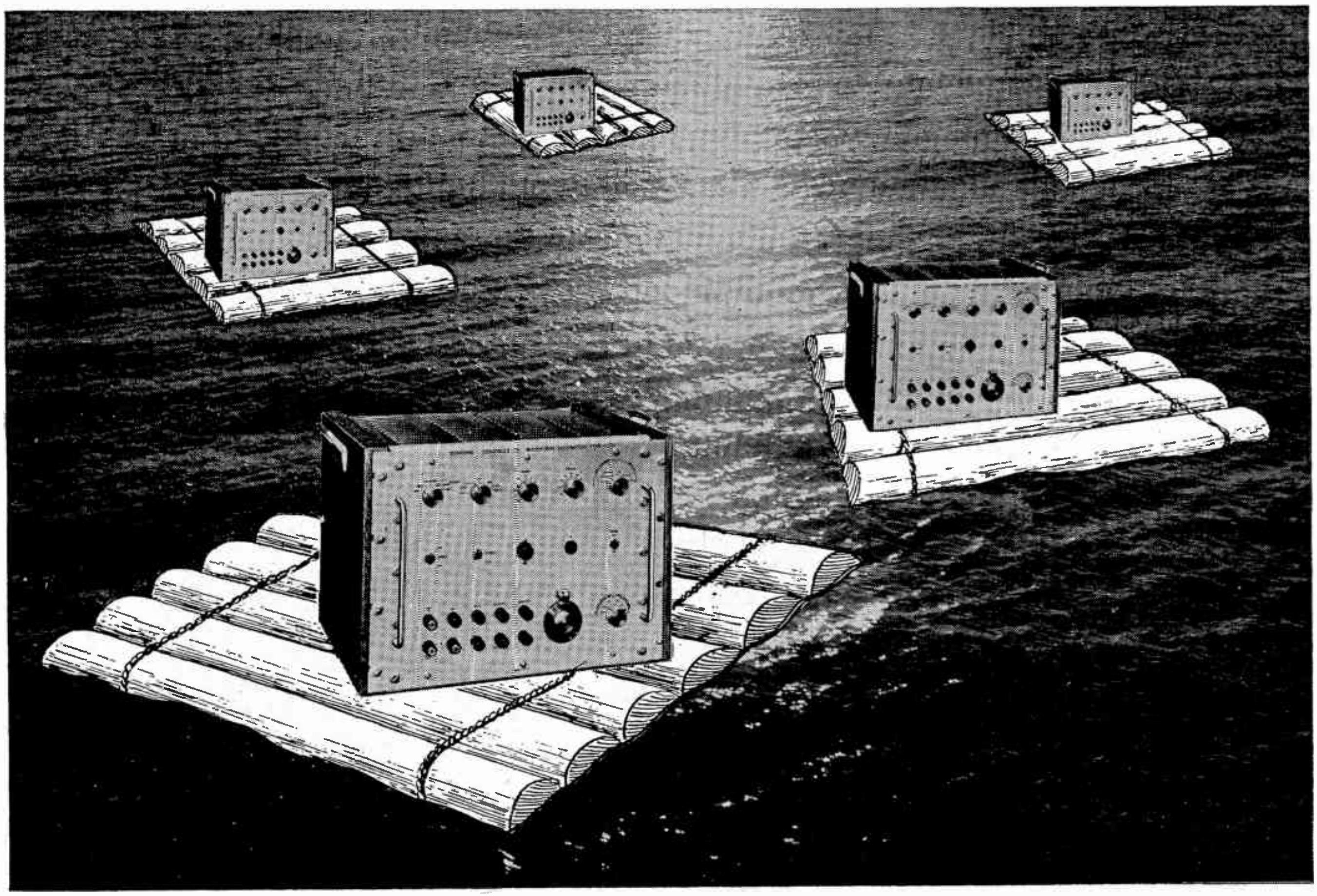
As the need for higher powers of audio frequencies increased, the G.E.C. used the basic design of the KT 66 and experience gained in its manufacture to develop the KT 88. This valve has a maximum anode dissipation of 35 watts as opposed to 25 watts for the KT 66, has a higher g_m and a cathode of larger emissive area. Physically the valve uses a smaller envelope and a pressed glass base and two valves in push-pull can provide 100 watts of audio power. The KT 88 is therefore ideal for high power public address systems in addition to many industrial applications.

For Data Sheets giving full technical descriptions of the KT 66 and KT 88 together with 'circuit supplement' sheets giving typical application details, write to the Valve and Electronics Department.

G.E.C.

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

40% of SERVOMEX Low frequency wave form generators are sold overseas



The LF. 51 is the most flexible supply of test signals for testing servo-mechanisms and automatic controllers

We have always claimed that the LF.51 is unique and we are very proud that it is furthering the reputation of British instruments among control engineers all over the world. With the aid of the LF. 51 and suitable recorders or oscilloscopes one can make exhaustive tests of the dynamic behaviour of almost any system capable of being excited by an electrical signal.

★ *Sine waves from 500 cycles/sec. down to one cycle every 2,000 secs. (33 minutes)*
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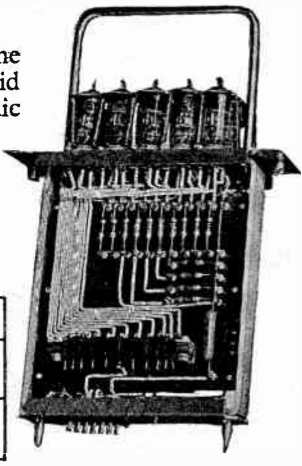
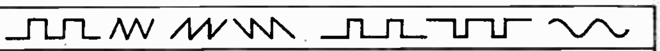
★ *Ramp functions with a linear rise variable from 1 millisecond to 1,000 seconds*
★ *Over 30 different waveforms available*
★ *Voltage variable between 100 microvolts and 150 volts peak-peak*
★ *Load current up to 5 milliamps peak*



1. Selector switch at CONTINUOUS"	
2. Selector switch at "SINGLE" - a step with continuously variable slope.	
3. Selector switch at "DOUBLE" - a single isolated pulse in TEN different forms.	

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

Radar 625

PORTABLE TELEVISION SIGNAL GENERATOR

A new instrument intended for the European market and countries using the 625 line television standards

MODEL 625 allows the development and service engineer to be completely independent of live transmissions and in addition provides him with a wide choice of test patterns of known characteristics for adjusting and testing television receivers.

Vision and Sound R.F. signals are provided *simultaneously* on Bands I and III and over the I.F. band 31.5 Mc/s to 40 Mc/s. In addition, a sound I.F. carrier at 5.5 Mc/s is available and facilities are provided for F.M. modulation of the carrier either internally, or externally from a radio receiver or phonograph.

The comprehensive range of patterns provided allows a complete and rapid checking of synchronising and interlacing; linearity and gradation; bandwidth and definition; width and height; sound-on-vision, etc. Fine vertical rulings allow "ringing" and overshoot to be checked. Fully modulated definition bars at 3.5 Mc/s, 4 Mc/s and 4.5 Mc/s are provided over the whole screen. The provision of a constant signal modulated with the correct synchronising and blanking pulses enables quantitative measurements to be made of the various stages of a receiver and the performance of different receivers may be readily compared with positive conclusions.

The synchronising waveform is

STRICTLY TO C.C.I.R. STANDARDS

with correctly interlaced frame and equalising pulses and the correct front and back "porches".

A steel case of robust construction ensures maximum protection and adequate ventilation is provided.

Model 625 is an instrument that no laboratory or service department concerned with 625 line TV can afford to be without.

BRIEF SPECIFICATION

R.F. Output.	Vision carrier 31.5 Mc/s to 85 Mc/s and 172 Mc/s to 225 Mc/s continuously variable. Intercarrier Sound at 5.5 Mc/s fixed operation. Sound and Vision signals may be used simultaneously. The 5.5 Mc/s carrier is available for I.F. alignment. Vision level: 10μV to 10 mV pp. (Negative modulation) Sound level: 3μV to 3 mV r.m.s. Controlled by continuously variable output control. Output impedance 300 ohms balanced or 75 ohms unbalanced.
Sound.	750 c.p.s. F.M., deviation 20 Kc/s internally or externally modulated.
Video Modulation.	Eight patterns in addition to plain black or white raster, include vertical and horizontal lines, graded bars and definition to 4.5 Mc/s.
Synchronising.	C.C.I.R. standard 625 line complete with interlaced frame synchronising and equalising pulses.
Video Output.	Positive and negative 5 V. pp.
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Size.	13½ in. (34 cm.) long; 10½ in. (27 cm.) high; 7 in. (18 cm.) deep. Weight 18 lb. (8 Kgs.)
Supply.	90/150 volts and 190/250 volts 50 cycles A.C. mains.

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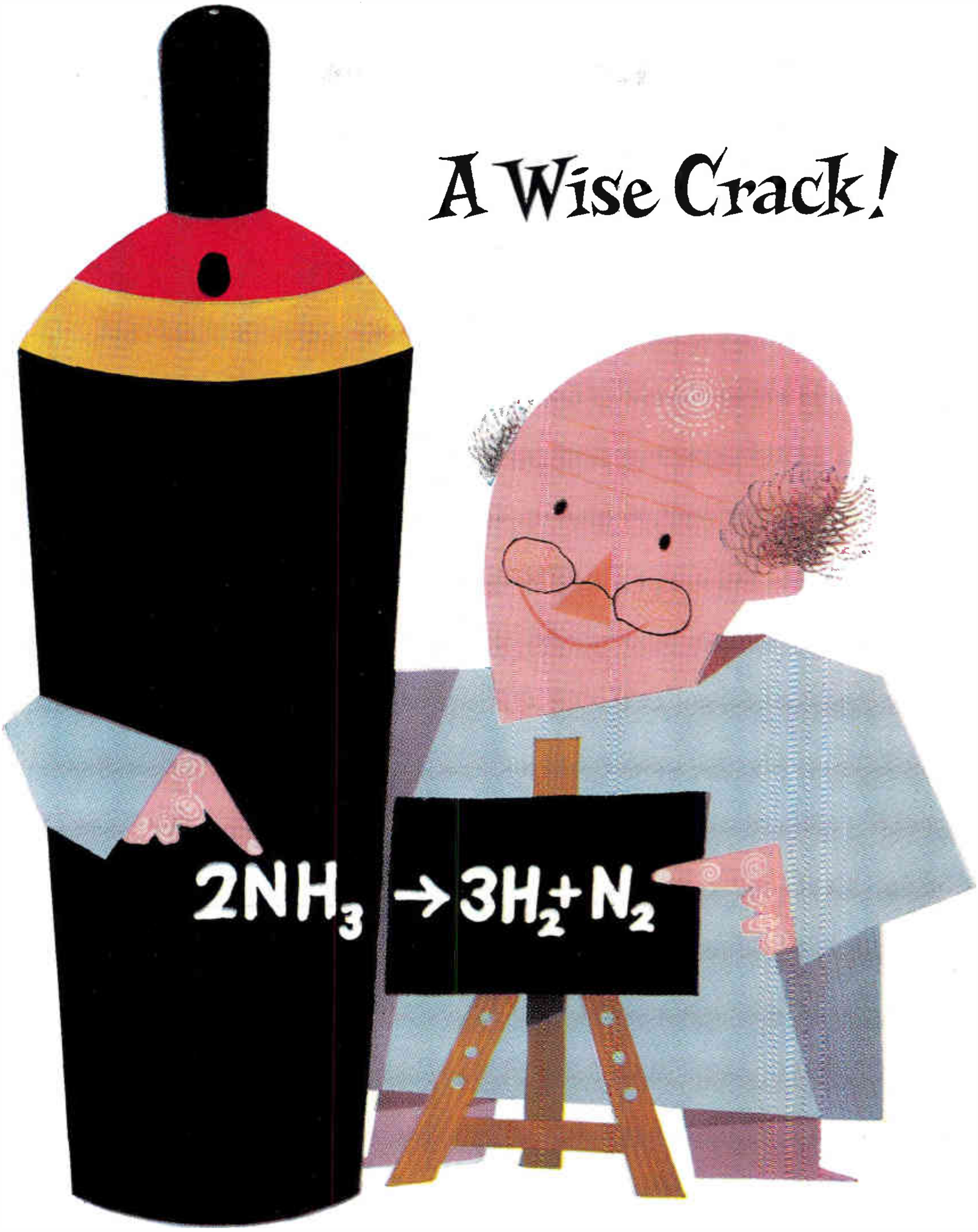

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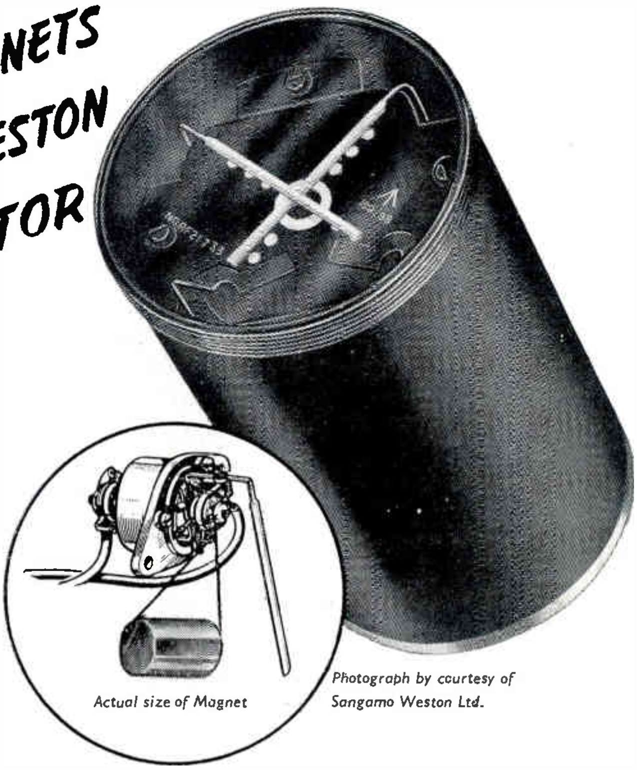
Actual size of Magnet

Photograph by courtesy of Sangamo Weston Ltd.

Four Murex magnets are used in the movement assembly of this hermetically sealed aircraft homing indicator by Sangamo Weston Ltd. In this application, as in many others, Murex Sintered Permanent Magnets provide high flux density and magnetic stability in a small area.

For technical information on the design, performance and characteristics of Murex Permanent Magnets write for a copy of the new Magnets Catalogue. Technical representatives always available for consultation and advice.

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General Circuit Techniques

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vital to engineers, designers and students

The final volume in this comphrehensive survey of modern television principles. Written by members of the BBC Engineering Division, it covers such subjects as counter circuits; frequency dividers; principles and circuitry of d.c. restorer and d.c. planting; gamma control amplifiers; fixed and variable equalisers; electrical characteristics of scanning coils; fixed and line output stages; etc.

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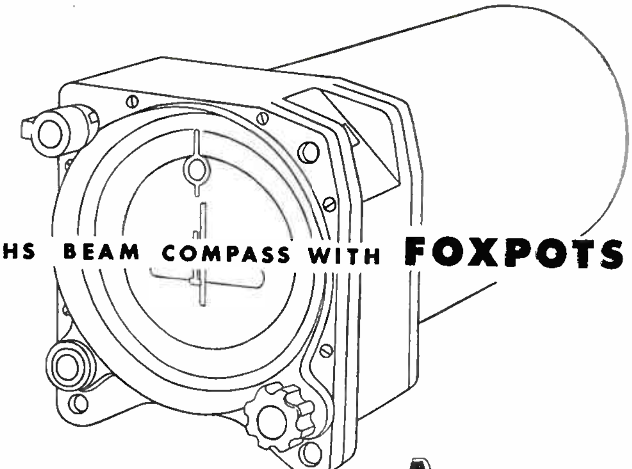

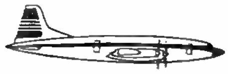

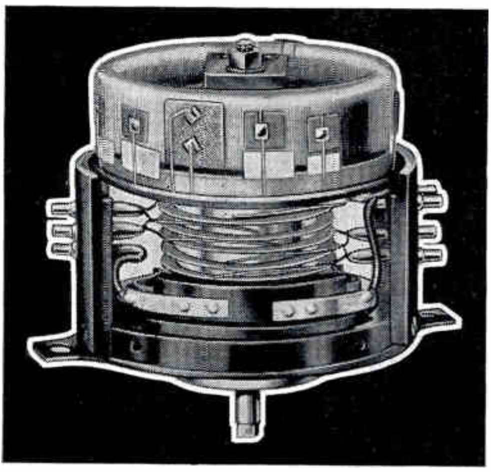

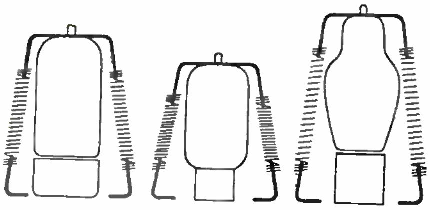

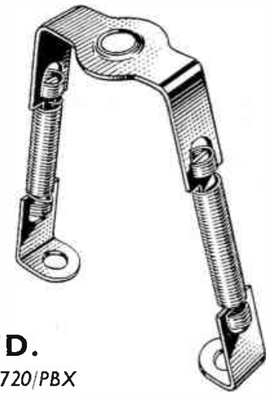

Z303C

Z502S

In production and available now

MULLARD DECADE STEPPING TUBES

Two widely used types of cold cathode stepping tube are now being manufactured by Mullard under the numbers Z303C and Z502S.

The Z303C counting tube, which is equivalent to the CV2271, has cathodes 1 to 9 brought out together and cathode 0 connected to a separate pin on the international octal base. The tube can count in a clockwise or counter-clockwise direction at a maximum speed of approximately 4 kc/s. Applications include nucleonic scalers, industrial counting, frequency division and timers.

The Z502S selector tube, which is equivalent to the CV2325, has the ten main cathodes brought out to separate pins on the B12E 12-pin base. This tube is suitable for industrial batching, welding control, electronic tachometry, telemetry and computing. In both tubes the glow is clearly visible through the dome of the bulb, and its position may be identified by means of an additional numbered escutcheon.

Data on the Z303C and Z502S and a comprehensive Report describing recommended circuits which offer a high degree of operational reliability are available free upon request.

Mullard

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TRANSFORMERS
Exact voltage control
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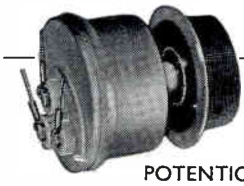
250VA — 2.5 kVA



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Proofed against Dust,
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Accurate — tough — compact.
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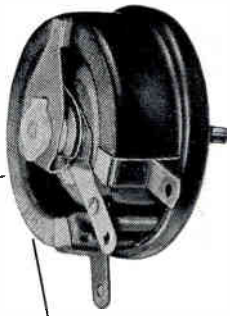

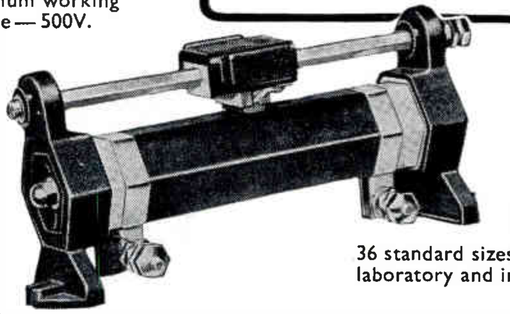

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laboratory and industrial use.

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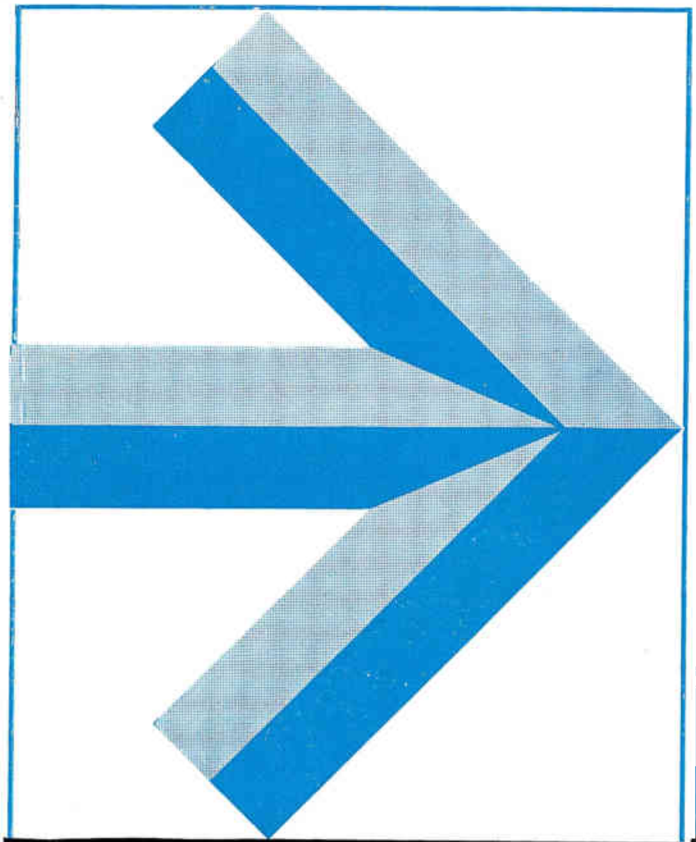

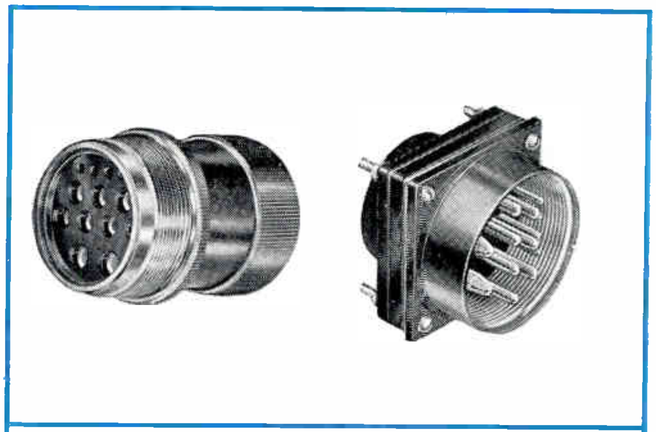

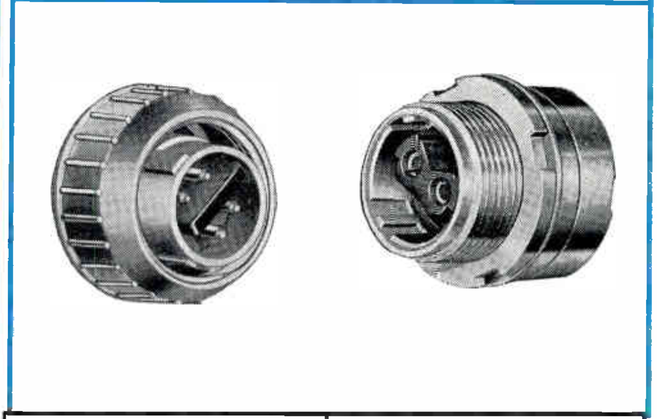

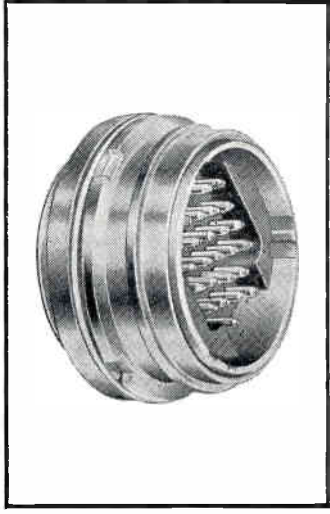

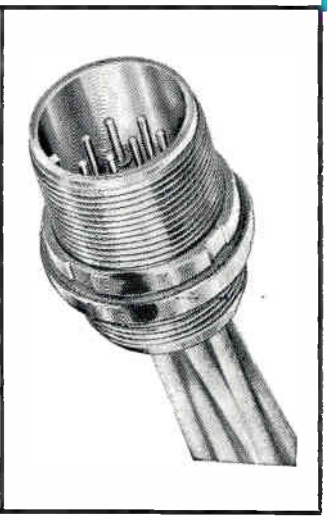

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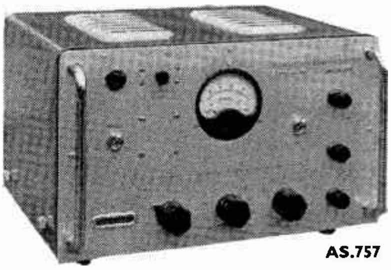
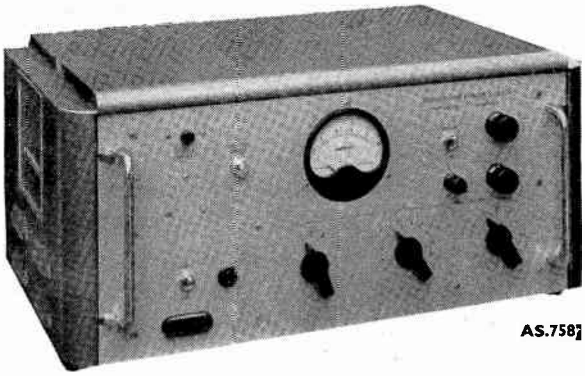


AS.757



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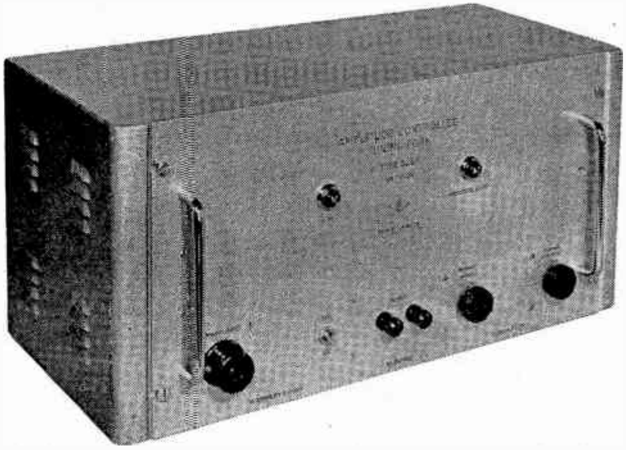

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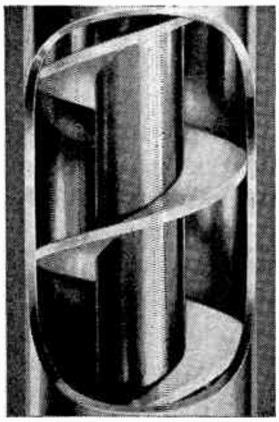

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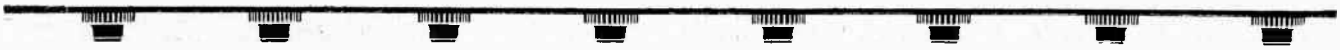
World Radio History



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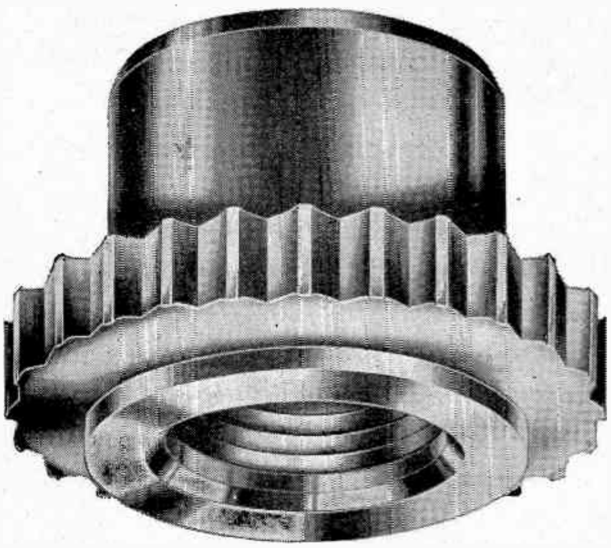

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The main tables first of all classify the valves by function (frequency changers, efficiency diodes, etc.), then by manufacturers' names and finally into current, replacement or obsolete types, as recommended by the makers. Within each section the valves are listed in order of their heater voltages. A useful feature of the layout is that it enables comparisons to be made between the electrical characteristics of valves from different manufacturers.

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World Radio History

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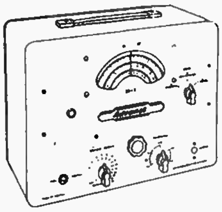

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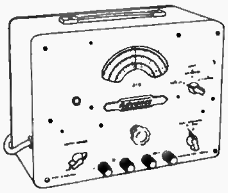

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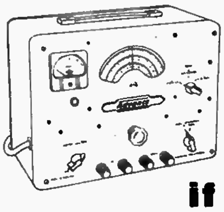

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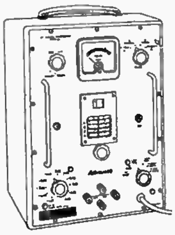

if it's signal generators

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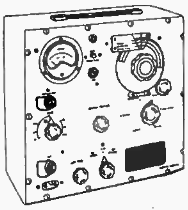

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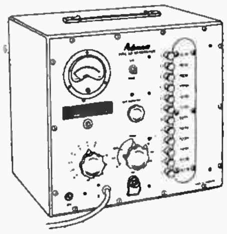

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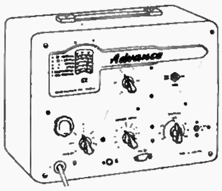

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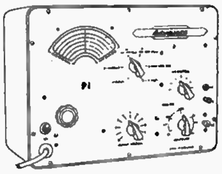

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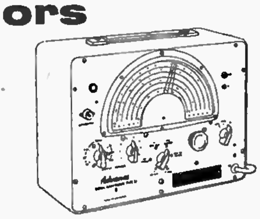

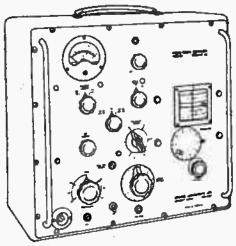

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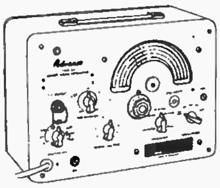

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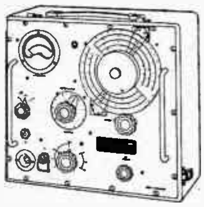

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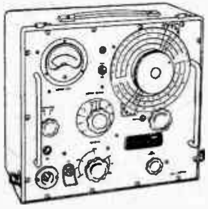

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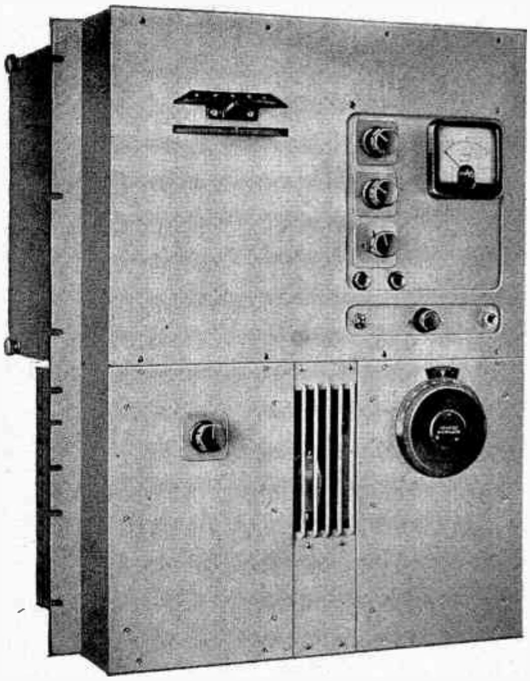
In this redesigned oscillator the high-frequency circuits are contained in a single small unit. Similarly the output amplifier and power supply are contained in a second unit.

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



FREQUENCY RANGE 0–16,000 c/s

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

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

FREQUENCY INTERPOLATING SCALE

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

ALTERNATIVE OUTPUTS ARE PROVIDED:

OUTPUT NO. 1

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

OUTPUT NO. 2

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

OUTPUT LEVEL

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

HARMONIC CONTENT

Output No. 1: – 60 dB total harmonic over whole frequency range.
Output No. 2: The harmonic content will vary somewhat according to the load, e.g. with a load resistance of 50 ohms or greater, and the output adjusted to 8 volts, the harmonic content will be less than – 50 dB over a frequency range 100 c/s–10 kc/s.

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

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CONTENTS

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Permanent Magnets

Design Advisory Service

No.6

Methods of Magnetising Magnets

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

Permanent magnets require a magnetising force proportional to the coercivity of the material. It is extremely important that the magnetising force is not below the specified minimum value, otherwise a reduced performance may be obtained from the magnet.

To assist the designer these values are given below.

MATERIAL	MAGNETISING FORCE	
	c.g.s.	M.K.S.
'Ticonal' C, G, G_X and L	2,500 AT/cm.	0.25×10^6 AT/m.
'Ticonal' K	3,600 AT/cm.	0.36×10^6 AT/m.
'Reco' 3A	2,000 AT/cm.	0.20×10^6 AT/m.
'Magnadur' I	12,000 AT/cm.	1.2×10^6 AT/m.

Modern magnet materials require considerably more magnetising power than earlier materials. It may be as high as 40 times that required by tungsten or chrome steels. In order to obtain the maximum effect from the magnetising current it is recommended to short circuit the magnet during magnetisation by a heavy iron yoke.

The magnetising current may be obtained by several methods; some of which are outlined below.

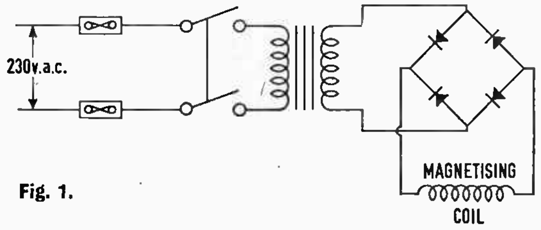

Fig. 1.

a. Metal Rectifiers (Fig. 1.)—generally preferable where it is possible to use a low current and many turns on a pre-wound coil. If an electromagnet is used it should be possible to attach specially shaped pole pieces to magnetise various types of magnets or magnet assemblies.

b. Ignitron Pulse Circuits—the use of ignitrons for controlling half cycle pulses is recommended where repetition magnetisation is required. The current available directly from the mains is usually sufficient for magnetising most magnets, but if a higher current is required, specially designed transformers are available for stepping the current up to higher values, of the order of 100,000 amps. This method is not recommended for magnets of large section as the eddy-current shielding effect produced during pulse magnetisation opposes the magnetising field and incomplete magnetisation may result.

c. Storage Accumulators—these are normally recommended for supplying current to coils for experimental work. When using accumulators adequate precautions should be taken for breaking the highly inductive circuit carrying high current.

d. Capacitor Discharge—the high current usually associated with magnetising can be obtained by discharging the current from a large charged condenser. This method generally uses an ignitron both as a switch and as a means to prevent oscillation with consequent partial demagnetisation of the magnet.

e. Motor-generators—can be particularly useful for supplying the current necessary for magnetising purposes. A wide variety of these are available to suit individual requirements.

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'MAGNADUR' CERAMIC MAGNETS
FERROXCUBE MAGNETIC CORES

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MC266

VOLUME 35 NUMBER 9 **ELECTRONIC & RADIO ENGINEER**

SEPTEMBER 1958 *incorporating WIRELESS ENGINEER*

National Radio Show

THIS year's exhibition is the Silver Jubilee, being the 25th National Radio Show. This does not mean that it started in 1934. There were none held during the war, nor in 1948, which puts back the start to 1926. There were exhibitions of similar character before that, of course, but as they were not 'national' they do not rank in the present series.

The last thirty-two years has seen enormous changes in domestic apparatus. Elsewhere in this issue we have included a few pictures to illustrate the modern styling and, with them, one of a 1926 receiver by way of contrast!

Television still forms a major part of the exhibition. Sound-broadcast receivers, which now very frequently include Band II f.m. reception, are another major section. What is new this year is the stress laid on high-fidelity and on apparatus for reproduction from stereophonic disc recordings.

For many years there has been an enthusiastic band of high-quality devotees who spared neither time nor money to improve the reproduction of sound. Recently, a knowledge of the benefits to be derived from such apparatus has been spreading far outside this coterie. So much is this the case that high-fidelity apparatus now contributes appreciably to British exports.

Curiously, the taste for such apparatus has spread among the general public rather more slowly here than abroad and it is noteworthy that this year's exhibition is the first at which any great stress has been laid upon it. Sound reproduction has now become a section of the show ranking equally with television and sound-broadcast sets.

Transistorized Vehicle Speedmeter

By D. R. Ollington, D.F.H.*

SUMMARY. *Electronic pulse-counting techniques are applied to measure the time taken for a vehicle to traverse a short known distance, and hence to determine the speed. A crystal oscillator provides the basic timing mechanism. It is started and stopped by the vehicle by means of hydraulic detector tubes, and the number of output pulses is displayed on meters. Transistors are used throughout.*

Developed at the request of the Road Research Laboratory of the D.S.I.R. for statistical analysis of vehicle speeds, this instrument is constructed entirely from transistors arranged in standard circuit blocks. The technique of measurement is to determine the time it takes for a vehicle to travel the distance between two rubber tubes spaced just under six feet apart on the road.

General

There are a number of methods available to determine speed on the road, most of these requiring a relatively long distance, or time, over which to make the measurement. Recently, two electronic methods have become available which give a nearly instantaneous reading, or, if not instantaneous, soon enough for a stationary observer to make a recording. The instruments concerned are the radar speedmeter and the time-interval speedmeter. The radar method uses a microwave beam to illuminate the road in the direction of approaching vehicles, and a receiver with a frequency-sensitive detector to record the reflected signals from any object within the beam. The difference frequency between the received signal and that radiated is displayed on a meter calibrated in m.p.h. Difficulties have been experienced when using speedmeters of this type for several reasons, perhaps basically because it is not possible to confine the radiated beam to a closely defined area, say, one side of a road, nor is it possible to say categorically how far its compass extends. There is also the difficulty that the range depends on the size of the reflecting surface.

Measurement of the speed of a vehicle by timing it over a short distance ensures that the velocity is known at a given point, say at a cross-roads or on a bend. There are several advantages with the system for this reason and a comparison of the two meters will be made at the end of the article.

Technique

The measurement of time in the transistorized speedmeter is carried out by totalling the number of pulses derived from a crystal oscillator on a three-decade counter. Fig. 1 is a block diagram of the instrument. The oscillator employed operates at 10 kc/s and the square-wave output is passed to two binary dividers to give a resultant frequency of 2·5 kc/s. This frequency has been chosen so that a convenient tube spacing on the road can be employed and so that the interpretation of the speed of the vehicle is simplified. The 2·5-kc/s signal is fed to the gate which is in turn opened and closed by pulses from pressure switches attached to the ends of the two rubber detector tubes. The gate is interlocked so that a pulse fed to it from the 'stop' detector prior to one being received from the 'start' detector does not affect the operation. Similarly, after a 'start' pulse has been received no further pulses at the input can affect the gate although the 'stop' input is, of course, now operative. After the gate has closed no further vehicles will open it until it has been manually reset. The counting decades which follow the gate each employ four binary stages arranged with feedback to form a scale-of-ten. All three are cascaded and an output is taken from each to a meter fitted on the front panel (see Fig. 2). The left-hand meter totals the hundreds of pulses counted, the centre, the tens, and the right-hand, the units. Since the frequency fed to the gate is 2·5 kc/s the unit of

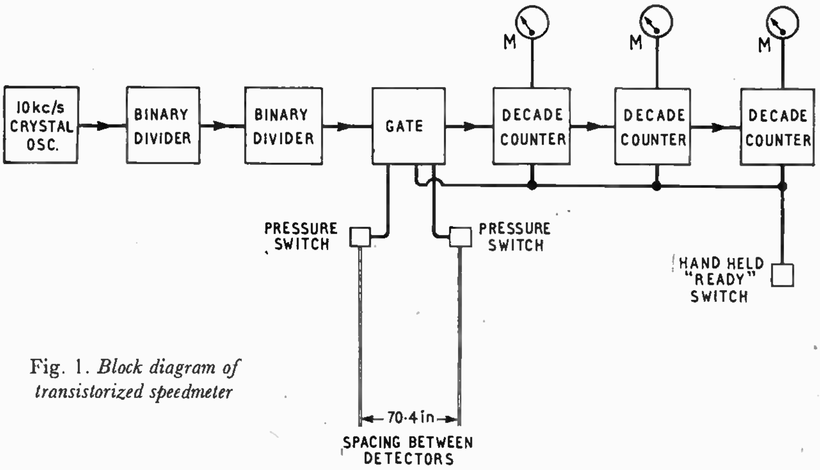

Fig. 1. *Block diagram of transistorized speedmeter*

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* *Venner Electronics Ltd*

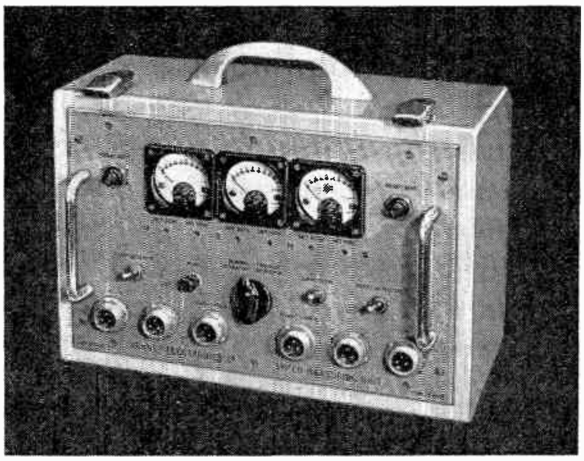

Fig. 2. *Transistorized speedmeter*

measurement is therefore 0·4 msec. It was a prerequisite of the design that the indication was to be a reciprocal of the vehicle speed so that ordinary mathematical tables could be used for conversion. In fact, a direct reciprocal is not possible since the indication is to three significant figures and the vehicle speeds to be obtained will be generally below 99 m.p.h. For example, a meter reading of 333 should correspond to a speed of 30 m.p.h., and the constant must therefore be 10,000. The detector spacing can be determined as follows :—

Let x be the detector spacing, in miles and y be the vehicle speed in m.p.h. Then t, the time taken for the vehicle to cross the detectors is x/y hours. In seconds, the elapsed time is $3600\ x/y$.

Since there are 2500 pulses generated per second, the number of pulses (n) which will pass while the vehicle traverses the detectors is $n = 3600 \times 2500\ x/y$ (1)

By definition $n = 10{,}000/y$ .. .. .. (2)

Equating (1) and (2)

$$x/y \times 3600 \times 2500 = 10{,}000/y$$

$$\text{Therefore } x = \frac{10{,}000}{3600 \times 2500} \text{ miles}$$

$$= \frac{10{,}000 \times 5280 \times 12}{3600 \times 2500} \text{ inches}$$

$$= 70{\cdot}4 \text{ inches.}$$

Potential Errors

There are two main sources of error, the gaining or losing of a pulse during the gating operation and the chance that the two detectors may not be set accurately on the road. In the first case the error is known, since in all gating systems of this type the tolerance is ± 1 count. Thus, at 30 m.p.h. either 332 or 334 may be the indicated count (for a true 333). The percentage error in this case is therefore 0·3%. It is not quite so easy to specify the error due to incorrect installation of the detector tubes, but it can be stated that an error of 1 inch from parallel is visually a very noticeable one, and exceeds that normally found in practice. An error of one part in seventy represents 1·4%. Thus, at 30 m.p.h. if both errors fall the same way the reading obtained may be 1·7% from true. This is equivalent to 0·5 m.p.h. If a jig is used each side of the road a very high order of accuracy is obtained.

It might occur to the reader that there is a source of error due to the detectors being installed at an angle other than a right-angle to the traffic flow. In fact, this is not the case so long as the spacing at the curbs remains at 70·4 ins, since the figure produced is a parallelogram. Traffic running parallel to the curb is still timed over the correct distance. If a vehicle crosses correctly-installed detectors at an angle, then there will be an error since the velocity of the vehicle along the road will be less than the velocity at which the vehicle is travelling. If the car crosses the strips at an angle of, say, 10°, then the error will be 1·5%. The equipment will always read lower than the true speed.

The shock-wave produced when a wheel passes over a detector travels down the rubber tube at 1100 ft. per second (approximately 1 ft. per msec). A difference of 2 ft in the length of the tubes can amount to an error of approximately 5 counts which, although not serious at 30 m.p.h. (1·5%) might be important when measuring speeds of the order of 80 m.p.h. (indication 125). The percentage error at this point is 4%. It is, however, a simple matter to ensure that the tubes are the same length and this error is not, therefore, of any great importance.

Operation

The batteries are included in the instrument, which is fitted in a lidded carrying case making the equipment completely portable. The detectors are, as previously described, rubber tubes which are laid on the road, being secured either by nails, if the surface is tarmac, or by plugs and screws, if concrete. Diaphragm pressure switches are connected to the ends of the tubes nearest the operator, the remote ends being terminated in plugs with a small hole at the centre to ensure that there is little or no reflection of the shock wave produced when a vehicle passes. The contacts of the diaphragm switch are connected via the sockets on the front panel of the instrument to the two Eccles-Jordan stages controlling the gate. A hand switch serves as a 'ready' control, making the instrument 'alert' to the

Fig. 3. *Speedmeter in use*

next vehicle that passes, and also to reset the gating circuitry and counting decades after a reading has been made.

The latest type of instrument (Figs. 2 and 3) has facilities for two sets of detectors and has connections for two 'ready' switches. One pair of detectors may be used on, say, the north–south side of the road and the other on the south–north side. The operator is now able to measure speeds of vehicles going in either direction by pressing either the left-hand or right-hand 'ready' switch. By operating the switch in the centre of the front panel both sets of detectors are paralleled and function so that as soon as a vehicle touches one of them this tube becomes the start detector, the other the stop detector. This arrangement is particularly suitable for country roads where there is little traffic, since it obviates the need for the operator to continually reverse the start and stop detectors.

Comparison with Radar Meter

Although both the time-interval speedmeter and the radar meter determine speed over a relatively short distance there are one or two important differences. As has already been stated, the radar meter has a relatively-wide illuminating beam and because of this it is not possible to use it on a road which has much traffic either approaching or receding from it. One can only be certain of the speed of a vehicle when it approaches the observer with an otherwise clear road. For traffic-speed analysis, this is a bad thing, since those vehicles recorded are to some extent unique, inasmuch as they were not followed or preceded by other traffic. From the point of view of police use the ideal offender is the one with an open road—perhaps the safest conditions for exceeding the limit, although admittedly outside the law. The road detector speedmeter can be used to determine the speed of almost any vehicle with the exception of one overtaking another. From a safety point of view it can be used to show that a driver exceeded the speed limit at dangerous points, such as at crossings or at known accident black spots.

Acknowledgments

This instrument was developed by the joint cooperation of the author with P. D. Whiting B.Sc. and J. A. Hillier B.Sc. of the Road Research Laboratory. The author is indebted to both the directors of Venner Electronics Limited and the Road Research Laboratory D.S.I.R. for permission to publish this article, and to N. L. Ayres, Graduate I.E.E., of Venner Electronics Limited for a good deal of the circuitry.

The Fringe of the Field

By Quantum

ALFVÈN WAVES

Some time ago I tried to tackle the Stanford University Press symposium on Magnetohydrodynamics, edited by R. K. M. Landshoff. This I found, in spite of its unassuming size, to be a very formidable piece of work, ranging from the theory of the dynamics of ionized gases to the generation and control of magnetically-driven shock waves. It seemed that, after all, it would be profitable to start a little closer to the beginnings of the subject, and H. Alfvèn's "Cosmical Electrodynamics" (Oxford, Clarendon Press, 1950), though no less formidable, does give a very clear idea to anyone quite unversed in these matters of what magnetohydrodynamics is about, and where its developments may lead. The real clue to the whole subject is, I suppose, the question of scaling. In extrapolating laboratory experiments to the conditions of the wider world, it is natural to assume that the results may be completely valid as they stand; but ship designers, for example, have known since the days of Reynolds and Rayleigh that a change of scale brings with it quite surprising consequences of its own. That certainly happens here as well.

Another feature is the tremendous importance of relatively small magnetic fields when spread over a sufficiently large volume of space, and the pictorial treatment of lines of force, not merely as the strained laterally-repelling etherial lines of Faraday, but as threads loaded with matter like spider-webs on a September morning. I should have added that this has to be a particular kind of matter—totally ionized and perfectly conducting—which again involves an extrapolation to the corona or the Crab nebula if you want to find it already there in a natural state. The Stanford symposium refers in several contributions to dimensional arguments, and the theory still seems to be in the tentative stage at which intuition plays a large part. This is a quite usual and probably essential approach to difficult problems; there is, for example, the classical case of Einstein's investigation of infra-red spectra which he undertook after ascertaining that the various factors involved combined to give a dimensionless product of a reasonable-looking *size*!

One-Dimensional (or Plane) Waves in a Conducting Fluid

The vector notation used in this paragraph is not going to burst out into any highbrow mathematics. As the first point to make is the interaction between current and magnetic field, which leads to a mechanical force, and thence to a mechanical wave, it seems best to dispose of this in the quickest way. The chapter on "Waves in Liquids" in C. A. Coulson's book on Waves (Oliver and Boyd, 1944) deals with the purely hydrodynamical side very clearly, and converts the vectors to Cartesians for you as well.

The equation of motion (that is, the counterpart of "force = mass × acceleration") for a unit mass of liquid of density ρ, where $\mathbf{V}$ is the velocity vector, $\mathbf{F}$ a vector representing the external forces, and p the hydrostatic pressure is

$$\frac{d\mathbf{V}}{dt} = \mathbf{F} - \frac{1}{\rho}\,\mathrm{grad}\,p \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

Maxwell's equations for the purely electromagnetic part are

$$\mathrm{curl}\,\mathbf{H} = \frac{1}{c}\left(4\pi\mathbf{i} + \frac{\partial\mathbf{D}}{\partial t}\right) \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

and

$$\mathrm{curl}\,\mathbf{E} = -\frac{1}{c}\frac{\partial\mathbf{B}}{\partial t} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

where $\mathbf{i}$ is the current-vector, $\mathbf{H}$ and $\mathbf{E}$ the magnetic and electric field vectors, and $\mathbf{B}$ and $\mathbf{D}$ the magnetic induction and electric displacement respectively.

The interweaving between equation (1) and Maxwell's equations arises because there is an *induced* e.m.f. in the liquid once it has started to move, so that this gives a contribution to $\mathbf{E}$; and, of course, a mechanical force on the current-carrying liquid which comes into the pressure-term of (1).

So there are, in addition,

$$\mathbf{i} = \sigma\left[\mathbf{E} + \left(\frac{\mathbf{V}}{c}\,\mathbf{B}\right)\right] \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

where σ is the conductivity, and the conversion of equation (1) to

$$\frac{d\mathbf{V}}{dt} = \mathbf{F} + \frac{1}{\rho}\left[\left(\frac{\mathbf{i}}{c}\,\mathbf{B}\right) - \mathrm{grad}\,p\right] \quad \ldots \quad \ldots \quad (5)$$

Finally, $\mathbf{B} = \mu\mathbf{H}$, where μ is the permeability.

These are the general equations. To adapt them to the one-dimensional (or plane) wave case, suppose the primary magnetic field $\mathbf{H}_0$ is parallel to the z-axis of a Cartesian system, and that the vector field $\mathbf{H}$ is $\mathbf{H}_0 + \mathbf{h}$; the only current flowing is in the x-direction, and the only motion of the liquid in the y-direction (Fig. 1). The displacement current $\partial\mathbf{D}/\partial t$ is negligible compared with $\mathbf{i}$, and after substitution and simplification the conductivity σ, which appears in the denominator of one term, is put equal to infinity. The resulting equation, in terms of the y-component h_y of $\mathbf{h}$ is

$$\frac{\partial^2 h_y}{\partial t^2} = \frac{\mu H_0{}^2}{4\pi\rho}\,\frac{\partial^2 h_y}{\partial z^2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

This is the equation of a plane wave, travelling out in both senses along the z-axis with velocity

$$v = \pm\, H_0\sqrt{(\mu/4\pi\rho)}$$

This expression gives v in cm sec^{-1}, since H_0 and μ are in absolute C.G.S. units. Substitution of reasonable figures for H_0 and ρ gives values for v which seem astonishingly low, as you can see for yourself.

The process represented by all these symbols is pictured in Fig. 1. The pillar of fluid A, displaced in the y-direction, experiences an induced electric field E, current flows in such a direction as to give a mechanical force opposing the change, but the return paths through pillars B and C are in the opposite directions, so both B and C are in their turn displaced to the right and the whole process is in its turn repeated with them, and with each successive pillar above and below. The *equation* is that for a propagation of a change in $\mathbf{h}$; but this is accompanied by a motion of the liquid in phase with the change in $\mathbf{h}$. In other words, it is just as if each line of force due to the original $\mathbf{H}_0$ was propagating a transverse wave, just like a stretched string that had been plucked, and was moving to and fro *carrying liquid attached to it*! I used the term "each line of force"; I hope I may be forgiven, for I don't know how many there are supposed to be, nor how much liquid is supposed to adhere to each. The term "liquid" has to be explained away too—this is more strictly a perfectly conducting incompressible fluid. The only thing there seems no need to fuss about is the attachment itself; for it follows that, unless infinitely great induced currents are to be allowed, there can be no relative motion of a magnetic field and a perfect conductor—if the field moves, the conductor simply has to move with it. In Chapter IV of his book, Alfvèn explores the behaviour of a simple sinusoidal wave of this type thoroughly, and then proceeds to show that the results are also true for oscillations of any arbitrary form.

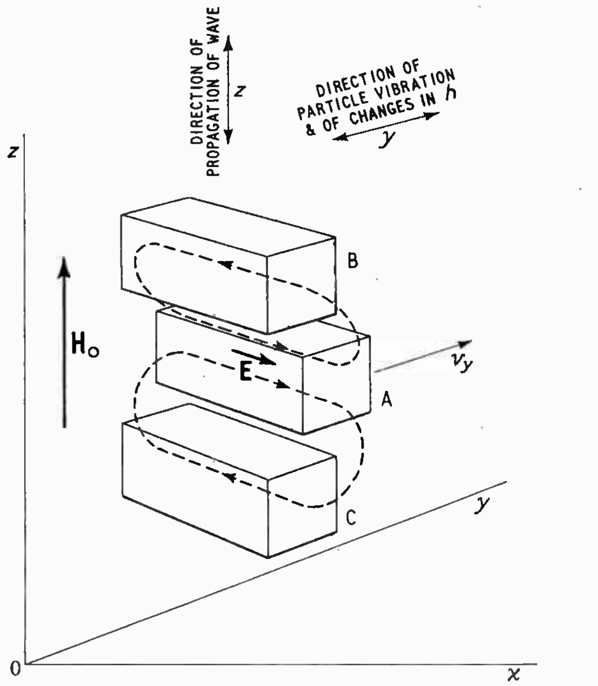

Fig. 1. *A very simple picture illustrating the propagation of an Alfvèn wave. The main field* $\mathbf{H}_0$ *is up the z axis. The dotted lines show the currents circulating while A is moving to the right parallel to the y axis. The resulting mechanical forces tend to restore A to its mean position, while displacing B and C to follow the motion of A. Hence a transverse wave, of liquid displacement in phase with variation in magnetic field, is propagated both ways in the direction of* $\mathbf{H}_0$.

Lundquist's Experiment

Without in any way approaching an infinitely good conductor, mercury or some liquid metal might be expected to offer some evidence for the possibility of such waves, heavily damped though they must be. S. Lundquist (*Nature*, 23rd July, 1949) succeeded in demonstrating this effect, with the apparatus shown in

Fig. 2. The vertical magnetic field was of the order of 10^4 oersteds, and rotational oscillations of a disc at the bottom of a mercury column about 15-cm long were transmitted up to the surface and observed using a little mirror floating on the top; the frequencies were of the order 0.1 to 1.5 c/s.

Further experiments on Alfvèn waves, by Lundquist and by B. Lehnert, in which the arrival of the waves at the top was detected by measuring potential differences there, are mentioned in the Stanford symposium book. An interesting feature of this work is the attempt to relate it to phenomena on the astrophysical scale; the appropriate scale factors were shown to involve "a large number"!

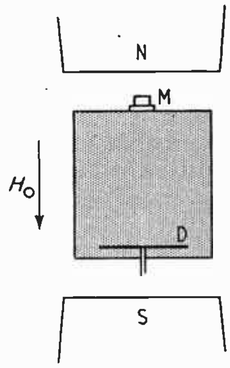

Fig. 2. Lundquist's experiment. The oscillating disc D propagates torsional Alfvèn waves up the mercury column, parallel to the magnetic-field direction. These are detected by the floating mirror M.

Scalesmanship

The gamesmanship quip has long outlived its topicality, and this heading is offered with more apologies; but the simplest analogy I can think of is a sporting one. You enjoy your billiards; but an hour on a quarter-sized table is a penance. You like a game of tennis; table-tennis is an equally diverting and serious sport. The difference? In one case little more has happened than a scaling down of the physical size of the arena and the impedimenta; in the other, the whole outfit has been adapted to give the same kind of game and employ the same kind of skill in miniature. What constitutes a proper change of scale in a physical experiment?

Since the Reynolds' Number, or its magnetic equivalent, plays an important part in the scaling calculations of magnetohydrodynamics, it may be well to take a little time off to consider the standard case of the ship (leaving the gamesman for good). For tank experiments on ship models, a faithful reduction of linear dimensions only would not give dynamically similar results.

The only factors that can possibly be concerned in the experiment are the linear size of the ship, represented by a length a, the velocity v, the viscosity of the liquid η, its density ρ, the force F pushing the boat along, and the acceleration due to gravity, g. When the dimensions of each of these factors in mass (M), length (L) and time (T) are written down, it is found that there are only three possible dimensionless combinations of them—the "reduced force" $F/\rho a^2 v^2$; the Reynolds' Number R which is $av\rho/\eta$; and the Froude Number v^2/ga. The physical relation involving the factors must be expressible in terms of these dimensionless products, in the form

$$\frac{F}{\rho a^2 v^2} = f_1\left(\frac{av\rho}{\eta}\right) + f_2\left(\frac{v^2}{ga}\right),$$

where f_1 and f_2 are some functions.

If this equation is to hold whatever the actual value of the linear size a, then each of the dimensionless products must have the same value always. Suppose for the moment that g and ρ are not alterable. If a is reduced by a factor of n, then the Froude number has the same value if v is reduced by $n^{½}$; when we have done this, the Reynolds' number will have the same value if we can make η decrease by a factor $n^{3/2}$; the propulsive force F must then be reduced by a factor n^3. Having done all these, if we can, then the whole operation may be performed in miniature under dynamically similar conditions. It should be noted that we do *not* have to know the form of the functions f_1 and f_2; this is the information it is hoped to *discover* from the scaled-down experiment.

The general principle behind all this is that, in scaling operations, the different factors must be altered in such a way as to preserve the same value for the dimensionless products of these factors. When this is done, a dynamically similar situation results. The Reynolds' Number R is probably the most important of these dimensionless products in ordinary fluid-flow; it crops up whenever fluids are being pushed about, and its size is important as well as its shape, for its value may often decide which of two alternative kinds of motion shall occur. It is also, as a dimensionless ratio, expressible in a number of ways, such as "the ratio of the dynamical term—involving v and ρ—to the dissipative term involving η", for example.

Now, when magnetic fields are changing in the neighbourhood of conductors, the dissipative effect depends on the conductivity, of course; and the idea of *magnetic viscosity* dates back to Ewing and Rayleigh, as a sort of second-order hysteresis effect due to the time it takes for magnetic lines of force to seep into the specimen. My edition of Bates's "Modern Magnetism" is nearly twenty years old, and refers to some work being undertaken in 1938 on magnetic viscosity in ferromagnetics; but in solids it is a small effect. In a conducting liquid, on the other hand, with lines of force and liquid waving about in phase (or trying to) magnetic viscosity plays as important a part as fluid viscosity. The dissipative effect is here represented by the magnetic viscosity $\nu_m = 1/\mu\sigma$; and the magnetic Reynolds' Number, R_m, is the dimensionless product obtained by substituting ν_m for η/ρ in the ordinary one. That is,

$$R_m = \frac{av}{\nu_m} = av\mu\sigma.$$

This is one of the quantities to be preserved unchanged in scaling experiments of the Lundquist type up to cosmic size. It will be noted that, for example, the fact that σ must necessarily be small with things like liquid metals is covered by its appearance in R_m. The ratio R_m/R has been estimated for ionized hydrogen as $2 \times 10^{-7}\ (\alpha/\rho)$, where α is the degree of ionization and ρ the density in grams per c.c., so that the extrapolation can in fact be done.

One final point. The magnetic viscosity is obviously a measure of the departure of a fluid from the 'perfect conductor' ideal. The lower its value, the more firmly is the 'liquid' attached to the magnetic lines of force. Only if R_m is very high do the conditions for the propagation of Alfvèn waves really hold.

Ringing Amplifier

USE IN CRYSTAL-VIDEO TRANSPONDER RECEIVERS

By S. Rozenstein* and E. Gross*

SUMMARY. *In certain secondary radar applications, radar transmitters operating in the microwave bands using pulse widths of the order* 0·1 *microsecond are used both as primary radars and as interrogators. The design of a transponder receiver on novel lines, for applications where small size, simplicity and low power consumption are desirable, is described.*

The input signals are r.f. pulses of 0·2 *μsec duration and the receiver is characterized by low noise level, nearly constant internal delay for large variations in signal strength, and extremely low power consumption.*

Operating on the tuned crystal detector–video amplifier principle, it is based on a novel method of pulse amplification to be called the 'ringing amplifier'. Use is made of the transient response of a tuned circuit to a pulse. The resultant damped oscillations are amplified in cascaded tuned stages. The second half-cycle of the wave is used to trigger the responder, the following oscillations being removed by a damping diode.

A gain of 110 *dB is obtained for a* 0·2 *μsec pulse in a miniaturized circuit of compact construction using five valves with h.t. consumption of only* 1·8 *W. The fixed internal delay is* 0·4 *μsec, and the delay variation is less than* 0·1 *μsec for a* 70-*dB variation in signal strength.*

The purpose of a radar transponder is either the amplification of the weak echo of a small or distant radar target, or the location, identification and tracking of one particular target to the exclusion of unwanted echoes. This permits the tracking of aircraft, small vessels and missiles beyond their radar range and the positive identification of certain ground locations.

The transponder, which is carried by the target, receives and amplifies the interrogating r.f. pulse of the distant radar and sends back either a single r.f. pulse or a series of coded pulses. This reply may be sent either on the frequency of interrogation or on a different frequency, depending on the system used.

The transponder contains a receiver which converts the received r.f. pulses into trigger pulses of suitable shape and amplitude in order to trigger its associated pulse transmitter (the responder). For most applications small size and low power consumption are of primary importance in transponder design. This article is concerned only with the design of the transponder receiver. The writers have been concerned with transponder applications where no separate interrogating transmitter is used, but where the primary microwave radar itself has also to serve as interrogator. In this case the transponder receiver must necessarily be pretuned to the frequency of the radar it is to be used with. Different types of receiver can be used for this purpose, the superheterodyne type giving the highest sensitivity[1]. However, the problem of securing local-oscillator stability necessitates in most cases the use of a.f.c. circuits which add to the bulk and power consumption of the receiver. Of greater practical value is the crystal-video type of receiver, which, though of a lower order of sensitivity[1], is distinguished by its stability, reliability and small size. These receivers consist of a pretuned r.f. circuit and a crystal detector followed by a high gain video amplifier. This means the elimination of the stability problems of a local oscillator.

For practical applications in the microwave region the r.f. circuit consists of a resonant cavity of suitable bandwidth matched to the detector crystal. Any number of cavities can be designed and pretuned for all the wavebands and frequencies of different radars and fitted to the transponder without necessitating any alignment in the field.

The design of the video amplifier depends on the pulse width of the radar interrogator. Certain types of modern microwave radars operate with extremely short pulses of the order of 0·1 μsec, which would necessitate a video bandwidth of the order of 10 Mc/s. In ordinary video amplifier circuits, even using the most suitable valves, wide bandwidth means low gain per stage and high overall power consumption, as in practice the valves operate with low anode-load resistances at high anode currents.

The overall gain of the amplifier must be sufficient to bring the input noise—mostly caused by the crystal detector—up to a level of about 10 V, which requires an amplification of 110–120 dB depending on the bandwidth[2]. The trigger circuit of the responder is so biased that the noise does not trigger the transmitter but video pulses of a signal-to-noise ratio of at least 2 : 1 cause a response to be transmitted. This occurs as soon as the leading edge of the trigger pulse reaches a certain level,

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* *Scientific Department, Ministry of Defence, Israel.*

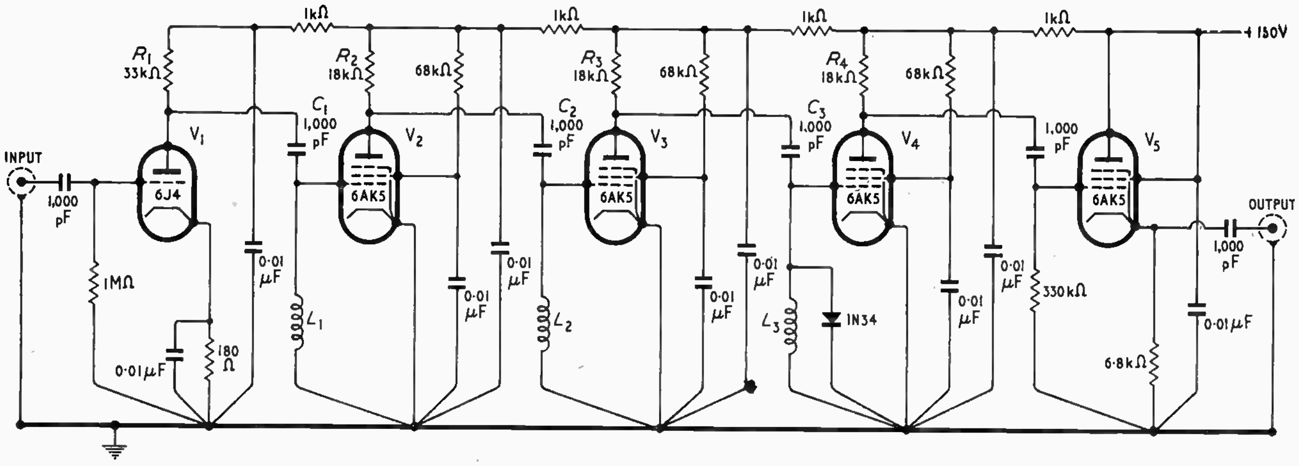

Fig. 1. *Schematic diagram of the ringing amplifier*

say 10 V above noise. The slope of the leading edge up to the triggering level depends on the amplitude of the pulse caused by variations of signal strength. Changes of slope cause a variation in the delay between received and transmitted pulse and therefore introduce inaccuracies in range measurement. Changes of the trigger slope should therefore be kept as small as possible within the working range of 70 dB variation in signal strength[3].

Principles of the Ringing Amplifier

Instead of amplifying a narrow pulse by a series of wide-band video amplifier stages, the pulse from the video detector is used to initiate a train of damped oscillations (ringing) in the first stage. These oscillations are amplified by cascaded tuned stages until they are of an amplitude sufficient to trigger the transmitter. As only a single trigger pulse is desired for every pulse received, all unwanted oscillations must be eliminated in the final stages of the amplifier. The single pulse thus obtained is passed via a cathode-follower as a trigger to the transmitter. Because of the transformation of the pulse into damped oscillations, the output pulse cannot be a faithful replica of the input pulse. This, however, is of no importance, as the purpose of this receiver is only to provide a fast trigger for its responder.

Description of the Practical Circuit

The practical amplifier (Fig. 1) consists of a triode input stage, three pentode amplifying stages and a cathode-follower output stage. The video pulse at the grid of the input stage V_1 causes ringing of the resonant circuit consisting of L_1 and the interstage stray capacitance, which is damped by the anode-load resistance R_1. A 6J4 triode was used here because of its low noise figure. The three 6AK5 pentodes operate with zero bias and at low screen voltages (40 V). Under these conditions their mutual conductance is 4 mA/V.

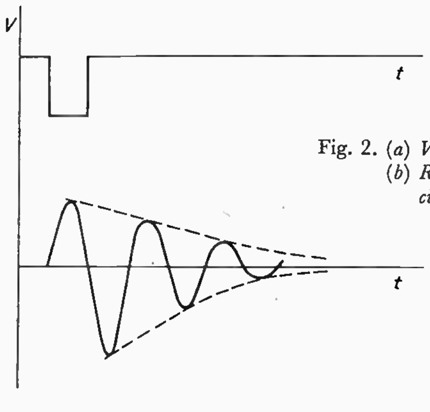

Fig. 2. (*a*) *Video pulse at grid of* V_1
(*b*) *Response of first resonant circuit in optimal case*

A 1N34 crystal diode is connected across the tuned circuit at the grid of the third amplifying stage V_4 which has no tuned circuit in its anode load. It therefore operates as a video amplifier. This stage is *RC*-coupled to a 6AK5 cathode-follower V_5. All these stages operate with low anode currents and the total current consumption is only 12 mA at 150 V.

Detailed Circuit Operation

The pulse from the video detector is negative at the grid of the input stage. The transient response of the tuned circuit at the anode due to the leading edge of the pulse is a train of damped oscillations whose first swing is of positive polarity. The pulse trailing edge causes another train of damped oscillations but of opposite polarity. The amplitude will be greatest if the oscillations caused by both the leading and the trailing edges of the pulse coincide. This is the optimal condition and will occur if the input pulse-width equals half the wave period (Fig. 2). Pulses shorter or longer than optimum will cause oscillations of a smaller amplitude. If the pulse is much longer than the period, two distinct trains of oscillations will be produced. The amplifier is designed to make use of the optimal condition.

The damped oscillations are amplified by two tuned amplifying stages. After amplification a change in the waveform takes place as the wave-train envelope now takes a finite time to rise to maximum (Fig. 3). This rise time depends on the Q of the resonant circuits and the number of cascaded single-tuned stages in the amplifier.

A compromise value of Q which gives sufficient gain per stage while allowing the envelope to rise fast enough to a maximum must be chosen. The condition must be avoided whereby with signals of decreasing strength triggering could be effected by a later swing of the wave than under strong-signal conditions. For consistent operation, only the first two half-cycles of the wave can be considered as trigger pulses, the second half-cycle, which is initially the strongest swing and which also receives higher amplification than the first because of the envelope rise-time, is therefore selected. It will be shown in the Appendix that the optimum value for Q permits the envelope maximum to occur coincident with third half-cycle of the wave.

Any positive signal above the noise will draw grid current in the last amplifier stage. The first swing at the grid of V_4, being positive, is therefore limited by grid current. The second half-cycle, which is of negative polarity at this grid, will appear as the first strong positive pulse at the output.

It should be pointed out that the coupling capacitor C_3, which is negatively charged because of the first swing, will add its potential to the amplitude of the second, negative, half-cycle and thus increase the trigger amplitude. This action is intensified by the crystal diode which reduces the charging time-constant of the coupling capacitor.

The next positive half-cycle, which is actually the strongest swing, drives V_4 into saturation and causes both the grid of V_4 and the crystal rectifier to pass current. C_3 then charges, and subsequently discharges through R_3 and L_3 in series. Until the end of this discharge, which is effective for about 2·5 μsec for the strongest signals, the receiver is saturated.

The waveform across L_3 (shown in Fig. 4) is non-oscillatory. An exact mathematical solution is difficult because of the non-linear elements in the circuit, but the following simplified analysis shows the general features of operation.

The equivalent circuit diagram is given in Fig. 5.

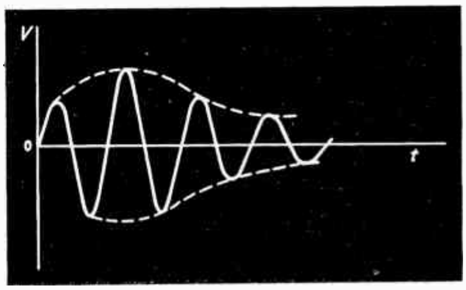

Fig. 3. *Train of oscillations with a finite rise time*

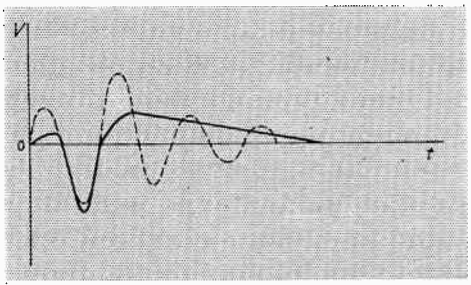

Fig. 4. *Waveforms at the grid of* V_4
Dotted line: under linear conditions (valve and diode removed)
Solid line: under actual non-linear conditions

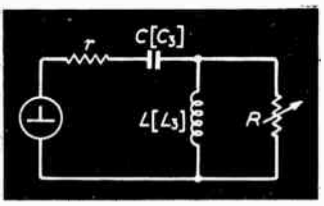

Fig. 5. *The equivalent circuit of* V_4 *grid*

R denotes the diode resistance which will vary according to the signal amplitude from a few kilohms down to 100 ohms. The stray capacitance across L_3 will be disregarded in the calculation. Its effect will be seen by observing the actual output waveform.

The input to the circuit is a wave of damped oscillations whose frequency is high compared to that of $1/2\pi\sqrt{L_3C_3}$ and may therefore be simulated as a number of successive positive and negative impulses. The output response to a positive impulse expressed in Laplace notation will be:

$$E_0(p) = \frac{\dfrac{pRL}{R+pL}}{\dfrac{pRL}{R+pL}+r+\dfrac{1}{pC}}$$

$$= \frac{p^2}{p^2\left(1+\dfrac{r}{R}\right)+p\left(\dfrac{r}{L}+\dfrac{1}{RC}\right)+\dfrac{1}{LC}}$$

$$= \frac{R}{R+r}\cdot\frac{p^2}{p^2+2\alpha p+\omega_0^2}$$

$$\text{where } \alpha = \frac{1}{2}\,\frac{r/L+1/RC}{1+r/R},$$

$$\omega_0 = \sqrt{\frac{1}{LC\,(1+r/R)}}$$

The response will be

$$E_0(t) = \mathcal{L}^{-1}E_0(p)$$

and will take the general form

$$e^{-\alpha t}\,[\cos\omega_1 t - (\alpha/\omega_1)\sin\omega_1 t] \qquad \text{for } \omega_0 > \alpha$$

$$e^{-\alpha t}\,(1-\alpha t) \qquad \text{for } \omega_0 = \alpha$$

$$e^{-\alpha t}\,[\cosh\gamma t - (\alpha/\gamma)\sinh\gamma t] \qquad \text{for } \omega_0 < \alpha$$

where $\omega_1 = \sqrt{(\omega_0^2 - \alpha^2)}$ and $\gamma = \sqrt{(\alpha^2 - \omega_0^2)}$.

In the actual circuit,
$r = 18\ \text{k}\Omega;\ C_3 = 1000\ \text{pF};\ L_3 = 270\ \mu\text{H};\ R_{crit.} = 2{\cdot}2\ \text{k}\Omega.$

As the amplitude of the output waveform is decreasing, R simultaneously increases and will eventually exceed $R_{crit.}$, so that the waveform will decay with a positive polarity without change of phase.

The combined effects of the oscillatory input and the oscillations of L_3 and its stray capacitance are responsible for the superimposed damped oscillations as they appear in Fig. 6 (f).

The effect of grid-current damping in the various stages on the ringing waveform may be explained in the following manner: If all stages in the amplifier are operating in the linear mode, the output from V_1, V_2, V_3, V_5 will be shown in Fig. 6 (a), (c), (e) and (g) respectively. The linear mode can exist because the grid–cathode resistance is fairly high for low inputs and becomes 1 kΩ only for positive peaks above 1 V. The practical damping is therefore performed only in the fourth stage and by the additional crystal diode.

If the input pulse from the video detector increases markedly, so that grid current damping in the first tuned circuit becomes noticeable, the latter will cause a change of the relative amplitudes between the first, second and third half-cycles [Fig. 6 (d)]. But as the triggering pulse is developed from the second swing and as its amplitude is sufficient to develop a satisfactory trigger pulse in the linear mode, no deterioration of the desired waveform will occur due to non-linear operation of the initial stages.

The outputs of stages V_1, V_2, V_3, V_5 under non-linear conditions are shown in Fig. 6 (b), (d), (f) and (h) respectively.

To illustrate the behaviour of the tuned circuit under linear conditions, Fig. 6 (i) shows the response to a 1·5-μsec negative input pulse, while (j) shows its response to the optimal input pulse of 0·2 μsec.

In order to avoid the creation of new oscillations, the last stage operates as an over-driven video amplifier with a resistive load. The positive trigger from the last stage is fed to the transmitter through a cathode-follower.

The non-linear operation of the amplifier stages effectively limits large signals by grid-current damping so that the input to the last video amplifier stage remains largely constant over a wide range of variations in signal strength. The selection of the second half-cycle of the damped oscillations as the trigger pulse inevitably introduces a certain amount of fixed delay which remains constant, however, over the whole operating range. The variable delay caused by the variation in amplitude at the grid of the last amplifying stage is very small and only of the order of 0·1 μsec over a 70-dB variation in signal strength.

Design Notes

The circuit diagram of the first single-tuned stage is shown in Fig. 7. The resonant frequency of the tuned circuit is first determined. If τ is the input pulse duration then the resonant frequency for the optimum case will be

$$f_0 = 1/2\tau \qquad (1)$$

The coil inductance may be approximated from

$$L = 1/\omega_0^2 C = \tau^2/\pi C \qquad (2)$$

where C denotes the total stray capacitance. The value of C is the sum of the output and input capacitances of the adjacent stages and the coil and wiring stray capacitances. It will be shown in the Appendix that the Q factor should be 2. The equivalent resistance in parallel with the tuned circuit is then calculated from

$$R = Q\omega_0 L = 6{\cdot}28\, L/\tau \qquad (3)$$

Assuming the a.c. resistance of the valve under zero bias conditions to be r_a, then the resistance will be

$$R_L = r_a R/(r_a - R) \qquad (4)$$

With the resonant frequency f_0 as the base, the design of the other tuned stages follows the same lines. The overall amplification in the ringing amplifier is obtained in two steps. The first step is due to the tuned stages and the second due to a conventional video stage.

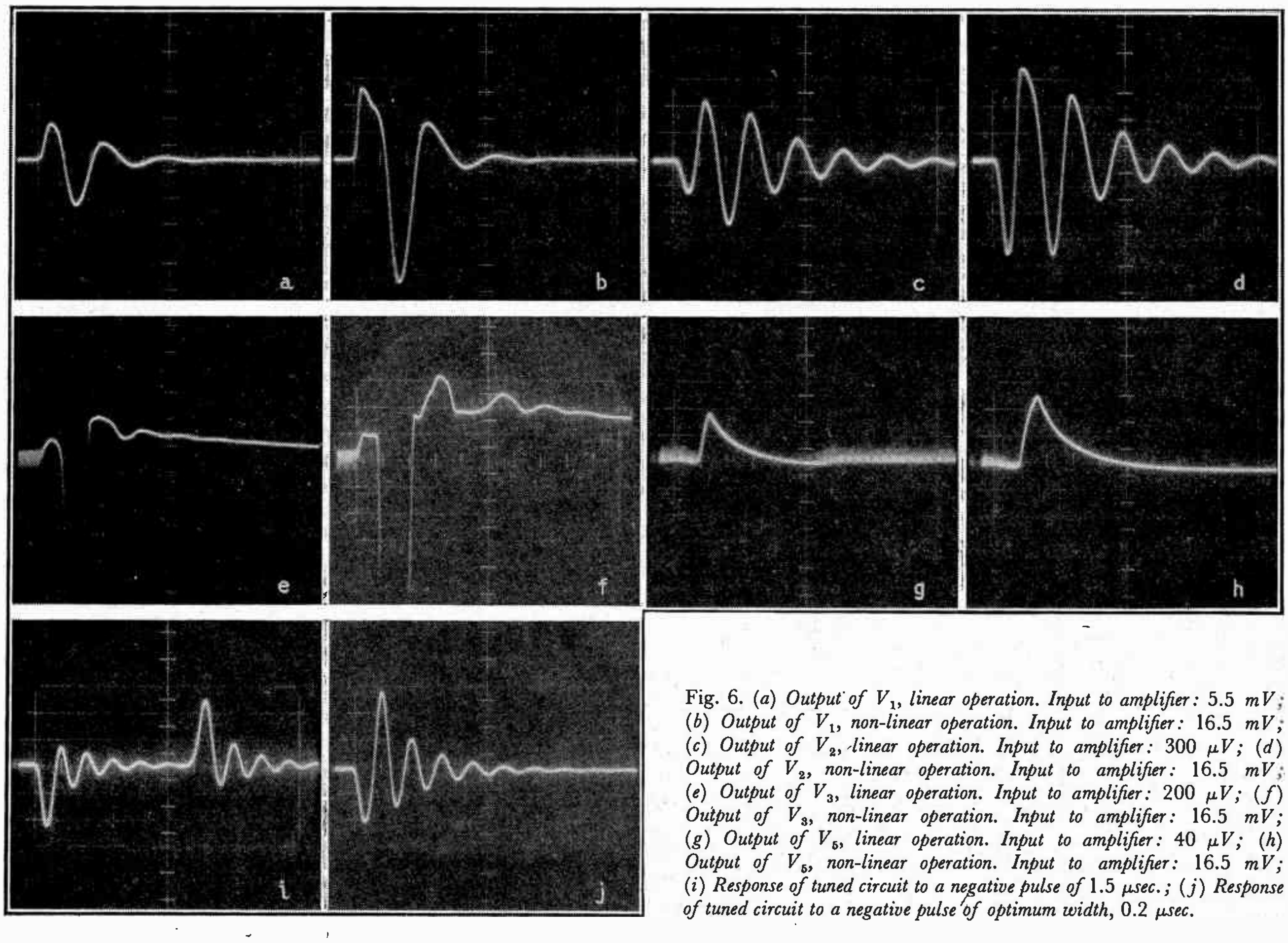

Fig. 6. (*a*) *Output of* V_1, *linear operation. Input to amplifier:* 5.5 *mV;* (*b*) *Output of* V_1, *non-linear operation. Input to amplifier:* 16.5 *mV;* (*c*) *Output of* V_2, *linear operation. Input to amplifier:* 300 μ*V;* (*d*) *Output of* V_2, *non-linear operation. Input to amplifier:* 16.5 *mV;* (*e*) *Output of* V_3, *linear operation. Input to amplifier:* 200 μ*V;* (*f*) *Output of* V_3, *non-linear operation. Input to amplifier:* 16.5 *mV;* (*g*) *Output of* V_5, *linear operation. Input to amplifier:* 40 μ*V;* (*h*) *Output of* V_5, *non-linear operation. Input to amplifier:* 16.5 *mV;* (*i*) *Response of tuned circuit to a negative pulse of* 1.5 μ*sec.;* (*j*) *Response of tuned circuit to a negative pulse of optimum width,* 0.2 μ*sec.*

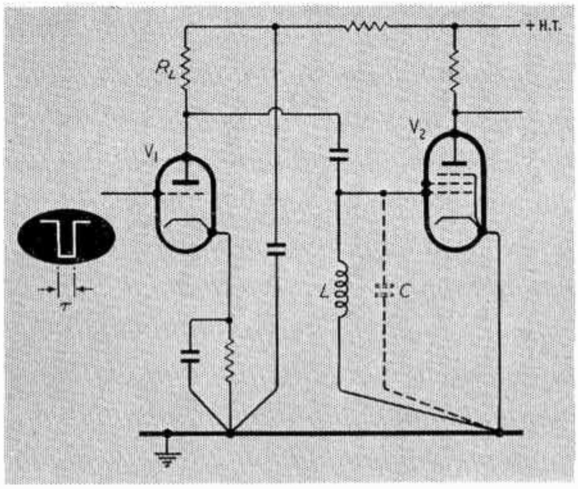

Fig. 7. *First single-tuned stage*

The overall tuned circuit amplifier gain is shown in the Appendix in case of three identical tuned stages to be

$$A_1 = (g_m R)^3/8 \quad \ldots \quad (5)$$

The gain of the final stage is

$$A_2 = g_m R_0 \quad \ldots \quad (6)$$

where R_0 is the load resistance of the output stage. The overall amplification of the amplifier is then

$$A = A_1 A_2 = g_m{}^4 R^3 R_0/8 \quad \ldots \quad (7)$$

Construction

The prototype model was constructed with miniature valves but sub-miniature valves could easily be used instead. Care must be taken in the mechanical construction of the amplifier to avoid feedback, as otherwise it is liable to oscillate. Every stage must be completely screened and all its earth connections made to a single point as in high-gain h.f. applications. The circuit diagram (Fig. 1) shows that there is no common h.t. line but that series filtering is used. The 1-kΩ filter resistors serve as feed-through insulators in the screening partitions. This kind of decoupling provides the highest attenuation between input and output.

Conclusions

The ringing amplifier makes use of a well-known principle for a radically new design of transponder receiver. For use with interrogating radars transmitting pulses of 0·1–0·2 μsec width it has a sensitivity equal to crystal-video receivers using wide-band video amplifiers, but it has appreciably lower power consumption. Because of its non-linear operation it has a large dynamic range with negligible variation of trigger delay. The number of valves and components is smaller than that of a video amplifier of similar specifications, increasing the reliability of the unit, while the lower power consumption permits easier miniaturization with lower heat dissipation. This should make it a useful component in the design of miniature radar transponders.

Acknowledgments

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APPENDIX

The following analysis is based on linear operation of the amplifier. In the practical amplifier non-linear operation starts when the output signal approaches the amplitude desired for triggering the responder. The linear analysis will therefore serve only to determine the circuit parameters for maximum sensitivity.

1. *Response of* n *identical single-tuned circuits to an input pulse*

The equivalent circuit diagram of one single-tuned stage is given in Fig. 8. The impedance of the tuned circuit expressed in Laplace notation is

$$Z(p) = \frac{p/C}{p^2 + p/RC + 1/LC} = \frac{p/C}{(p+\alpha)^2 + \beta^2} \quad \ldots \quad \ldots \quad (1)$$

where

$$\alpha = \frac{1}{2RC} = \frac{\omega_0}{2Q}; \quad \beta = \omega_0 \sqrt{1 - \frac{1}{4Q^2}}$$

The value of Q is defined as $Q = \omega_0 RC = R/\omega_0 L$ where $R = r_a R_L/(r_a + R_L)$ and ω_0 is the resonant angular frequency of the tuned circuit. The frequency f_0 is so chosen that half its period will equal the pulse duration τ. This case is recognized as the previously mentioned optimum case. The angular frequency ω_0 can then be written as $\omega_0 = \pi/\tau$.

The discussion on the transient response to an input pulse will be in two steps. The transient response to the leading edge of the input pulse will first be analysed. The transient response to the trailing edge will then be deduced from the previous analysis as it will be of the same nature except for a shift of τ seconds in time accompanied by the reversal of the waveform phase. The final output response is the sum of the above mentioned responses. Assuming the use of n identical tuned circuits and an input pulse having an ideal waveform, the response to an input step of amplitude E_{in} is

$$E_n(p) = E_{in}\,(1/p)\,(-1)^n\,[g_m Z(p)^n]$$

$$= E_{in}\,(-1)^n g_m{}^n \frac{p^{n-1}}{[(p+\alpha)^2 + \beta^2]^n} \quad \ldots \quad \ldots \quad (2)$$

If we put $a = \alpha + j\beta$; $b = \alpha - j\beta$ then

$$E_n(p) = E_{in}\,(-1)^n\, g_m{}^n \frac{p^{n-1}}{(p+a)^n\,(p+b)^n} \quad \ldots \quad \ldots \quad (3)$$

and the output response as a function of time

$$E_n(t) = \mathcal{L}^{-1} E_n(p) = E_{in}\,(-1)^n \frac{g_m{}^n}{(n-1)!} \frac{d^{n-1}}{dp^{n-1}}$$

$$\left[\frac{p^n e^{pt}}{(p+b)^n\,(_{p\,=\,-a})} + \frac{p^n e^{pt}}{(p+a)^n (_{p\,=\,-b})}\right] \quad \ldots \quad \ldots \quad (4)$$

2. *The solution for the case of three single-tuned stages*

Assuming a negative input pulse the response to the negative-going leading edge for the case of three stages ($n = 3$) is

$$E_{3L}(t) = E_{in} \frac{g_m{}^3}{8C^3} \exp.\left(-\frac{\omega_0}{2Q}t\right)\left\{\left[\frac{1}{\omega_0}\left(1 - \frac{1}{4R^2}\right)t^2 - \frac{1}{Q\omega_0{}^2}t + \frac{1}{\omega_0{}^3}\left(1 + \frac{3}{4Q^2}\right)\right] \sin \omega_0 t + \left[\frac{1}{Q\omega_0}t^2 - \frac{1}{\omega_0{}^2}\left(1 + \frac{3}{4Q^2}\right)\right] \cos \omega_0 t\right\} \quad \ldots \quad (5)$$

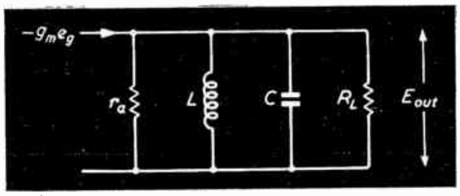
The response E_{3T} to the positive going trailing edge is

$$E_{3r}(t) = -E_{3L}(t - \tau) \quad \ldots \quad (6)$$

The total response is

$$E_3(t) = E_{3L}(t) - E_{3L}(t - \tau) \quad \ldots \quad (7)$$

Fig. 8. *Equivalent diagram of single-tuned stage*



3. *Determination of the Q factor*

It was previously mentioned that for high overall gain a high-Q circuit is desirable, while in order to obtain the maximum of the envelope at the peak of the second swing a low-Q circuit is necessary.

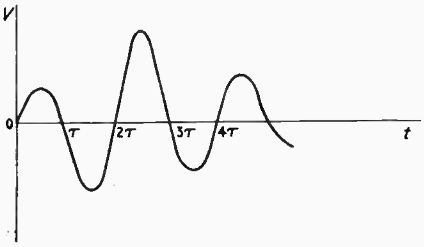


Fig. 9. *Response of the third stage*

It was felt that a good compromise would be to let the maximum of the envelope occur at the peak of the third swing. The time at that instant is $t = (5/2)\tau$ (Fig. 9). The value of $\cos \omega_0 t$ in Equ. (5) in the neighbourhood of that time is negligible while the value of $\sin \omega_0 t$ is practically unity. The amplitude of the third swing in the third stage will then be

$$E_3(t) = -\frac{g_m{}^3}{8C^3}\exp.\left(-\frac{\omega_0}{2Q}t\right)\left\{\frac{1}{\omega_0}\left(1-\frac{1}{4Q^2}\right)t^2-\frac{1}{Q\omega_0{}^2}t + \frac{1}{\omega_0{}^3}\left(1+\frac{3}{4Q^2}\right)+\exp.\left(\frac{\pi}{2Q}\right)\left[\frac{1}{\omega_0}\left(1-\frac{1}{4Q^2}\right)(t-\tau)^2 - \frac{1}{Q\omega_0{}^2}(t-\tau)+\frac{1}{\omega_0{}^3}\left(1+\frac{3}{4Q^2}\right)\right]\right\} \quad \ldots \ldots \ldots (8)$$

diffentiating E_3 (t) in respect of time and equating to zero gives

$$0 = -\frac{\pi}{2Q\tau}\left(1-\frac{1}{4Q^2}\right)t^2 + 2t - \frac{3\tau}{2Q\pi}\left(1+\frac{1}{4Q^2}\right) + \exp.\left(\frac{\pi}{2Q}\right)\left[-\frac{\pi}{2Q\tau}\left(1-\frac{1}{4Q^2}\right)(t-\tau)^2+2(t-\tau)-\frac{3\tau}{2Q\pi}\left(1+\frac{1}{4Q^2}\right)\right] \quad \ldots \ldots (9)$$

putting $t = (5/2)\tau$ the value of Q is obtained. The calculated Q factor is equal to 2.

4. *Determination of the overall tuned-circuit gain*

The overall tuned-circuit gain is defined as the ratio between the peak amplitude of the second swing in the final tuned circuit to the input pulse E_{in}.

The absolute value of the second swing is evaluated by putting $t = (3/2)\tau$ and $Q = 2$ in Equations (5), (6) and (7). The gain equation will then be

$$A = \frac{|E_{s.s.}|}{E_{in}} = \frac{1}{30}\left(\frac{g_m\tau}{C}\right)^3 \quad \ldots \ldots \ldots \ldots \ldots (10)$$

The gain equation may be written in another form by putting

$$Q = \omega_0 RC = \frac{\pi}{\tau}RC$$

$$A = \left(\frac{g_m R}{Q}\right)^3 \quad \ldots \ldots \ldots \ldots \ldots \ldots (11)$$

In our case where $Q = 2$ the gain is

$$A = \frac{1}{8}(g_m R)^3 \quad \ldots \ldots \ldots \ldots \ldots \ldots (12)$$

REFERENCES

[1] A. Roberts: "Radar Beacons", M.I.T. Radiation Laboratory Series No. 3. McGraw-Hill, 1947, p. 151.

[2] Ibid., p. 177.

[3] Ibid., p. 178.

Millimicrosecond Pulse Generator

EQUIPMENT FOR TRANSIENT TESTS ON WIDE-BAND NETWORKS

By O. H. Davie, M.I.E.E.*

SUMMARY. *The generator uses a length of high-frequency cable whose distributed capacitance is charged from a known d.c. potential. The cable is then discharged into the load by means of a magnetically-operated mercury switch. Pulses of* 1*-millimicrosecond rise time are produced at frequencies up to* 120 *per second. The essential features of rise time, waveform and duration of the pulse are all controlled by the relatively stable characteristics of the cable and pulse amplitude is continuously variable, without the use of a signal attenuator.*

The type of pulse required for transient tests of aperiodic networks whose frequency range extends to tens of megacycles must have a rise time of only a few millimicroseconds, have minimum distortion and be variable in amplitude over a very wide range.

Suitable pulses may be generated by the simple arrangement shown in Fig. 1. This consists of a length of high-frequency cable; e.g., Uniradio 67,† which is used as a uniform transmission line comprising series inductance with distributed capacitance to earth. This capacitance is charged through the resistor R_1 to a potential V determined by the setting of the pulse amplitude control. When fully charged, the cable is connected to the output load by means of a special co-axial switch having mercury-wetted contacts and driven, in the simple case, at mains-supply frequency.

When the switch contacts close, the energy stored in the cable begins to transfer to the output load resistor. The potential to earth at the input to the cable (point A, Fig. 1) then *falls* to a value $V/2$, while the output potential *rises* to $V/2$ to form the leading edge of the output pulse ($t_0 - t_1$, Fig. 2). The rise time of this leading

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* *Technical Director, Cossor Instruments Ltd.*

† Equivalent Cables: U.S.A., RG8/U; France, KX50MI.

edge is determined by the high-frequency characteristics of the cable and the presence of stray inductance and capacitance in the circuit. To reduce the effect of these stray parameters, the complete system comprising the charging cable, switch, output lead and termination are screened and designed, as near as possible, as a continuous 52-Ω line. By the use of the most suitable cable and by careful design of the system as a complete coaxial unit it is possible to obtain a pulse rise time (10%–90%) of $0{\cdot}8 \times 10^{-9}$ sec. With a rise time of 10^{-9} sec. the overshoot can be reduced to less than 3%.

The duration of the pulse is determined by the length and propagation constants of the charging cable. The negative step waveform, at A in Fig. 1, travels along the cable with a velocity proportional to $1/\sqrt{LC}$ which, in the case of Uniradio 67, is approximately 8 inches per millimicrosecond. Since there is an open circuit at the far end of the cable, the step waveform is reflected in the same phase. The reflected negative step travels back along the cable until it reaches the point A, when it reduces the output potential to zero and terminates the pulse ($t_2 - t_3$, Fig. 2). The duration of the pulse is thus twice the transit time of the charging cable. At the end of the pulse the cable is not fully discharged, since the output will assume a potential determined by the value of the resistors R_1 and R_2. The output during this time ($t_3 - t_4$) is approximately 1% of the pulse amplitude but it returns to zero when the switch contacts open and the cable begins to recharge.

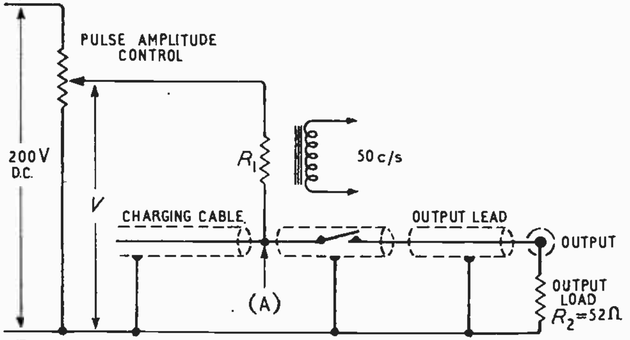
It will be noted that the pulse amplitude is controlled by variation of the direct potential applied to the resistor R_1, so that no variable attenuator network is required in the high-frequency parts of the circuit. The pulse amplitude is varied without change in waveform and positive or negative-going pulses can be generated by reversing the polarity of the d.c. supply.

The Mercury Switch

This simple form of pulse generation is made possible by the use of a unique form of relay, developed originally for telephone work. It was required to provide a low-resistance contact, free from electrical and mechanical wear, and capable of operating consistently without attention for long periods of time. It is manufactured by C. P. Clare & Co., Pratt Blvd., Chicago, Illinois, and a life of 10^{10} operations is claimed.

The switch is illustrated in Fig. 3 and consists of a glass capsule containing an atmosphere of hydrogen at high pressure. Operating in this atmosphere is a single armature swinging between two sets of contacts mounted at the opposite end of the capsule. Platinum contact surfaces are used and these are wetted with mercury.

Fig. 1. *Basic circuit of pulse generator*



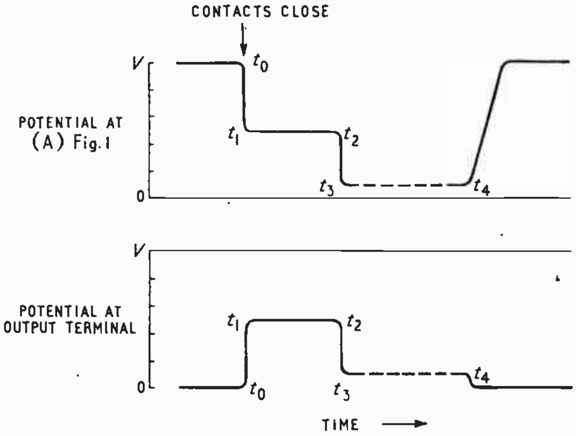


Fig. 2. *Circuit waveforms*

Mercury reaches the contacts by means of two parallel wires which form a capillary connection along the armature to a reservoir at the base. The mercury is replenished by the wick action along the capillary and returns to the reservoir when the contacts open. The procedure by which it returns can be studied from stroboscopic photographs. As the contacts separate, the mercury joining the two surfaces extends to form a filament. This filament becomes narrower in cross-section and finally parts at two points leaving a globule of mercury to fall under gravity to the reservoir below. Because of the mercury flow, it is necessary to operate the switch in the vertical position as shown. Further information on the construction of this type of switch is given by Brown & Pollard[1].

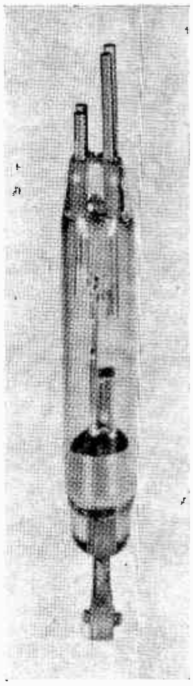


Fig. 3. *Hermetically sealed mercury switch element*

(Courtesy of C. P. Clare & Co., Chicago)

Supply Circuits

A convenient method of providing the magnetic drive for the switch uses a solenoid wound with its axis in line with that of the switch, the switch being effectively inserted into the winding in a manner similar to an iron core (Fig. 5). The drive required with this type of winding is approximately 200 ampere-turns and this can be derived from the mains-supply frequency. It is, however, preferable to drive from a source whose frequency can be adjusted, thereby avoiding the effects of mechanical resonances in the system. The effects of these resonances can be reduced by magnetic damping, some measure of which is obtained automatically when the solenoid is driven from a low-impedance source. Perhaps not the most efficient, but

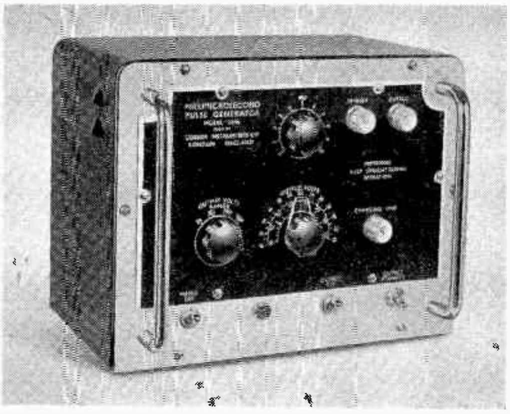


Fig. 4. *A commercial form of the generator*
(Courtesy of Cossor Instruments Ltd., London)

certainly a convenient source of low impedance uses a cathode follower whose input signal is derived from a multivibrator of variable frequency. Satisfactory operation is possible at frequencies up to 120 c/s.

Amplitude calibration of the output pulse can be obtained from a meter indicating the potential V applied to R_1. This potential is twice the amplitude of the output pulse [Fig. 2 (b)]. In a commercial form of this generator (Fig. 4), pulse amplitude is obtained directly from two calibrated controls. One of these is a stepped attenuator which reduces the direct potential by known amounts according to the range of amplitude required, and the other is continuously variable. By this means, it is possible to set the amplitude to any desired level from 1 to 100 volts, but lower outputs can be obtained if required.

Trigger Pulse

When the generator is used in conjunction with a high-speed oscillograph it is usually necessary to delay the signal pulse until the time base is operating. Sometimes the time taken for the pulse to pass through the network under test is sufficient for this purpose but usually an additional delay is necessary.

In Fig. 4 the trigger pulse is derived from a 6 : 1 potential divider connected directly to the output so that both pulses arrive at their respective terminals at the same time. The signal pulse may then be delayed by passage through a longer output cable than that connected to the trigger circuit. The delay obtained should be kept to the minimum to preserve the pulse rise time and with most transient oscillographs the length of cable rarely exceeds 5 feet.

Impulse Waveform

For very many tests the charging cable can be replaced by a non-inductive capacitor contained in a completely screened unit. The output waveform then takes the form of a fast rise followed by an exponential decay. The rise time is still in the region of a few millimicroseconds, depending upon the high-frequency performance of the capacitor. A resistor of 52 Ω is assembled in series with the capacitor in the screened unit and provides the correct source impedance to feed the output circuit. Since the decay time constant is relatively long, the output can be used as a step waveform for high-speed transient tests.

Low-Frequency Applications

When the charging cable is removed the circuit can be used as a low-frequency square-wave generator of extremely good performance. In one position of the switch the output terminal is connected to earth via the load resistor R_2, while in the other position it is connected to the tap of the potential divider formed by the resistors R_1 and R_2. With this arrangement of the circuit, the output waveform can be of very low frequency, have a perfectly flat top and be free from the effects of contact bounce often associated with electro-mechanical generators. When R_2 is made equal to R_1; e.g., 47 kΩ, the pulse amplitude is restored to its former value, $V/2$.

Performance

The great advantage of this type of generator is that the essential characteristics of the pulse—rise time, waveform and duration—are all controlled by the relatively stable constants of a high-frequency cable. The pulse is therefore reliable and not likely to change its characteristics unbeknown to the operator.

It is usually of sufficient amplitude for direct connection to the deflectors of a cathode-ray tube so that its characteristics can be examined before application to the network under test. The pulse amplitude can then be reduced, or even reversed in polarity, without change in waveform or the introduction of uncertainties often associated with the transient performance of attenuators. Because of this feature it is particularly suited to the testing of wide-band amplifiers[2].

The maximum amplitude which can be obtained is limited by the potential across the switch contacts. For most applications this potential is limited to 500 V but above 120 V a minute arc may occur a few millimicroseconds before the contacts close. When this occurs the pulse rises to approximately 80% of its peak amplitude, stays at this level while the arc is conducting and rises to its peak when the contacts close.

In many applications the leading edge of the pulse may be too fast for the network under test, involving frequency components which the network would not normally be expected to receive. In these circumstances, the rise time may be increased by the use of a simple *RC* integrating circuit at the output of the generator. The generator can then be used for transient tests on almost any aperiodic network. On the other hand, a network

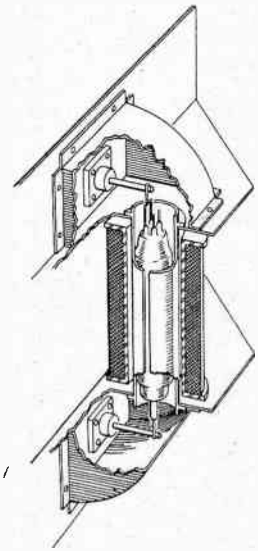


Fig. 5. *A typical switch unit showing the relative positions of the capsule, coaxial copper screen, and driving solenoid*

which has a good response to a one-millimicrosecond transient will not add anything to the waveform of slower transients, even if their frequency spectrum extends beyond the range for which the network was designed. By the use of an integrating circuit of this type the generator can provide a high-speed exponential time base and will display in detail the waveform of events which can be triggered from the leading edge of the output pulse.

The disadvantage of the generator is the very low duty cycle of the pulses produced and this is limited by the operating frequency of the switch. With commercial cathode-ray tubes operating at 2 kV it is necessary to exclude all extraneous light to observe the trace. Since many oscillographs now make use of tubes with post-deflection acceleration, it is often convenient to remove the connection to that electrode and apply an external source of higher potential when transient tests are being conducted.

By suitable choice of the value of the capacitor used in place of the charging line, the fast rise and exponential decay of the output waveform can be made to simulate the pulse produced by a scintillation counter. This enables the generator to be used for certain tests on counting equipment. It will produce a steady 3,000 counts per minute when the switch is driven at mains frequency and a maximum of 7,000 per minute using the multivibrator circuit. By the judicious use of cables and by the technique of 'bouncing' it is possible to obtain various combinations of pulses, both upright and inverted, which will meet the test requirements of specialized equipment. The main use, however, lies in its ability to generate reliable pulses of suitable characteristics for the transient testing of wide-band aperiodic networks.

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MATHEMATICAL TOOLS

By Computer

Conditions for Minimum Variation in a Function—2

In last month's article we considered how to minimize variations in a polynomial over a given range, and found two alternative approaches. We discovered that we could make the polynomial "maximally flat", so that there was no appreciable variation until a high power of the independent variable x became important, or we could follow Tchebycheff and reduce to a minimum the total variation over the given range, without restricting the part of the range at which the greatest and least variations occur. Here, and in next month's article, we seek to extend these ideas to a function which is not a polynomial. The particular functions we have chosen are rational functions (the quotient of two polynomials) in which the denominator is always positive, and the numerator has lower degree than the denominator. The methods used do not depend upon these properties of the particular functions chosen, and can be applied to many other types of function also. The rational function arises in connection with the problem of group-delay correction.

Gouriet[1] has shown that the transfer function associated with the bridged-T network of Fig. 1, for a steady input frequency $\omega/2\pi$, is

$$\frac{\phi_2(-p)}{\phi_2(p)} = \frac{1-ap+bp^2}{1+ap+bp^2} = \frac{(1-b\omega^2)-ja\omega}{(1-b\omega^2)+ja\omega} = G_2(\omega) \quad (1)$$

where

$$L = aR, \quad C = a/R, \quad q = a/\sqrt{b} \quad .. \quad .. \quad (2)$$

and R is the termination resistance. If $q > 1$, so that the shunt inductance in Fig. 1 is negative, Gouriet also points out that an effectively equivalent circuit is obtained by having coupling between the two series inductances with coupling factor $(q^2-1)/(q^2+1)$. The third member of Equ. (1) shows that the gain is unity at all frequencies, since the numerator and denominator differ only in the sign of their imaginary parts, and they therefore have the same modulus

$$D_2 = \{(1-b\omega^2)^2 + a^2\omega^2\}^{1/2} \quad .. \quad .. \quad .. \quad (3)$$

We are concerned with the way in which the phase θ_2 of $G_2(\omega)$ in Equ. (1) varies with frequency. Our aim is to make θ_2 vary with ω (over a given range) as linearly as possible (or rather, ultimately, to make the corresponding phase-angle θ_4 obtained when two of the bridged-T networks of Fig. 1 are connected in cascade vary as linearly as possible with ω, as explained

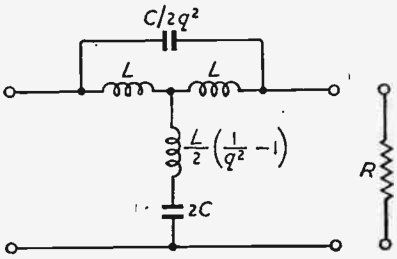


Fig. 1. *Bridged-T all-pass filter with transfer function of Equs.* (1) *and* (2)

below). If θ_2 is to vary as linearly as possible with ω, then $d\theta_2/d\omega$ must be kept as constant as possible, so that we have to deal with the same type of problem as in last month's article, but with the difference that $d\theta_2/d\omega$ is, as we shall see, not a polynomial in ω nor the reciprocal of such a polynomial.

In order to evaluate the phase θ_2 of $G_2(\omega)$, it is convenient to define ψ_2 by the equation

$$\tan\psi_2 = \frac{a\omega}{1-b\omega^2} \quad (0 \leqslant \psi_2 < \pi) \quad .. \quad .. \quad .. \quad (4)$$

in which case, from Equ. (3),

$$\sin\psi_2 = a\omega/D_2\ ; \quad \cos\psi_2 = (1-b\omega^2)/D_2 \quad .. \quad (5)$$

From Equ. (4) or Equ. (5) combined with Equ. (1),

$$G_2(\omega) = \frac{1-j\tan\psi_2}{1+j\tan\psi_2} = \frac{\cos\psi_2 - j\sin\psi_2}{\cos\psi_2 + j\sin\psi_2}$$
$$= \frac{e^{-j\psi_2}}{e^{j\psi_2}} = e^{-2j\psi_2} \quad .. \quad .. \quad .. \quad .. \quad (6)$$

so that the phase θ_2 of $G_2(\omega)$ is given by

$$\theta_2 = -2\psi_2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7)$$

The group delay is defined as $-\, d\theta_2/d\omega$ or $2d\psi_2/d\omega$, so we require $d\psi_2/d\omega$. This is most easily obtained by differentiating both sides of Equ. (4) with respect to ω as they stand, thus from the right-hand side

$$\frac{d}{d\omega}\left[\frac{a\omega}{1-b\omega^2}\right] = \frac{(1-b\omega^2)a-(-2b\omega)a\omega}{(1-b\omega^2)^2} = \frac{a(1+b\omega^2)}{(1-b\omega^2)^2} \quad (8a)$$

and from the left-hand side

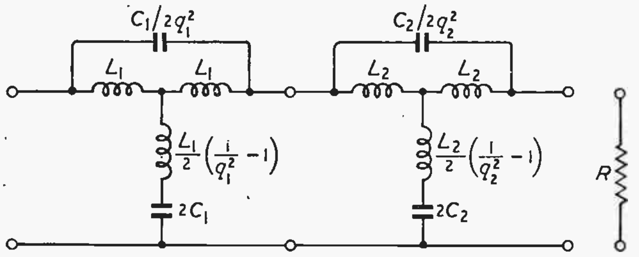
$$\frac{d}{d\omega}\left[\tan\psi_2\right] = \frac{d}{d\psi_2}\left[\tan\psi_2\right]\cdot\frac{d\psi_2}{d\omega} = \sec^2\psi_2\frac{d\psi_2}{d\omega}$$
$$= \frac{D_2^2}{(1-b\omega^2)^2}\frac{d\psi_2}{d\omega} \quad .. \quad .. \quad .. \quad (8b)$$

Equating the last member of Equ. (8a) with the last member of Equ. (8b), we have

$$\text{Group delay} = -\frac{d\theta_2}{d\omega} = 2\frac{d\psi_2}{d\omega} = \frac{2a(1+b\omega^2)}{D_2^2} \quad (9)$$

First, we have to consider how to keep this as constant as possible at low frequencies. In last month's article, where the expression whose variations were to be minimized was a polynomial, we achieved this by equating to zero the coefficients of as many powers of the independent variable as we could, in ascending order. In the case of the group delay in Equ. (9), however, we cannot just put b equal to zero, because that leaves out of account the variations of the denominator D_2^2. We need to express the group delay as a series (finite or infinite) of ascending powers of ω^2, and then equate to zero as many of the coefficients of powers of ω^2 as we can, in ascending order. It is possible to write down such a series if we know a sufficient number of the successive derivatives of the group delay when $\omega = 0$, but the evaluation of these derivatives is awkward because of the denominator D_2^2 in Equ. (9). Instead, we shall obtain our series by dividing the denominator D_2^2 into the numerator, cancelling the lowest power of ω^2 remaining at each stage. If D_2^2 is divided into $(1+b\omega^2)$ in this way, the first three terms of the quotient are

Fig. 2. *Two bridged-T all-pass filters in cascade*



$$1 + (3b-a^2)\omega^2 + (5b^2-5a^2b+a^4)\omega^4 \ .. \quad .. \ (10)$$

and this process of division is simply a way of finding out that

$$D_2^2\{1+(3b-a^2)\omega^2+(5b^2-5a^2b+a^4)\omega^4\}$$
$$= 1+b\omega^2+(\text{terms in } \omega^6 \text{ and higher powers}) \quad .. \ (11)$$

For sufficiently small ω, so that ω^6 is negligible, the expression (10) is an adequate representation of $(1+b\omega^2)/D_2^2$. If we want the expression (10) to be as constant as possible at low frequencies, we clearly require

$$a^2 = 3b \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \ (12)$$

and in this case the coefficient of ω^4 in the expression (10) reduces to $-a^4/9$. Thus to keep the group delay [Equ. (9)] as constant as possible at low frequencies, a, b must satisfy Equ. (12) and must be as small as possible, and the group delay will fall from its zero-frequency value.

If we now consider the alternative requirement that there shall be as little variation of the group delay as possible for $0 \leqslant \omega \leqslant 1$, we first want to consider the difference between the group delay [Equ. (9)] and its value $2a$ when $\omega = 0$. This is

$$2a\left[\frac{1+b\omega^2}{D_2^2} - 1\right] = \frac{2a}{D_2^2}\,\omega^2\{(3b-a^2)-b^2\omega^2\} \quad (13)$$

In the absence of the denominator D_2^2, we should expect, by analogy with last month's article, that the correct procedure was to make

$$a^2 = b(3-b) \quad .. \quad .. \quad .. \quad .. \quad .. \ (14)$$

so that the group delay would return to its initial value when $\omega = 1$. Substituting from Equ. (14) into Equ. (9), the group delay reduces to

$$\frac{2\{b(3-b)\}^{1/2}(1+b\omega^2)}{1+(b-b^2)\omega^2+b^2\omega^4} \quad .. \quad .. \quad .. \ (15)$$

This is found by straightforward differentiation to have a maximum when

$$b\omega^4 + 2\omega^2 - 1 = 0 \text{ or } b\omega^2 = (1+b)^{1/2} - 1 \quad (16)$$

and the maximum value is

$$\frac{2\{b(3-b)\}^{1/2}}{2(1+b)^{1/2}-1-b} \quad .. \quad .. \quad .. \quad .. \ (17)$$

For small b, the expression (17) differs little from the value $2a$ or $2\{b(3-b)\}^{1/2}$ of the group delay for $\omega = 0$ and for $\omega = 1$, but the values of b near 3, the expression (17) is large. For $b = 1\cdot5$ its value is $3/(\sqrt{10}-2\cdot5)$ which is about $4\cdot53$, and thus the group-delay varies between 3 and $4\cdot53$ for $0<\omega<1$; for $b=1$ the corresponding variation is between $2\sqrt{2}$ when $\omega=0$ or $\omega=1$ and $2+\sqrt{2}$ when $\omega = \sqrt{2}-1$, a range of

$2-\sqrt{2}=0{\cdot}586$. Thus for small and moderate values of b, we have obtained a reasonably satisfactory 'maximally-flat' curve and another curve with satisfactory overall variation in the range $0<\omega<1$, by techniques analogous to those used in last month's article for minimizing variations of a polynomial. For sufficiently large values of b, these curves cease to be satisfactory. Unfortunately, these are only two variable parameters a and b in Equ. (1), and this does not give us very much room for manoeuvre, so we now pass on to the case of two of the bridged-T networks of Fig. 1 in cascade. We shall adopt the same general procedure, but the equations are more complicated.

The transfer function for the two bridged-T networks for Fig. 1, arranged as in Fig. 2, is

$$\frac{\phi_4(-p)}{\phi_4(p)} = \frac{1-a_1p+b_1p^2}{1+a_1p+b_1p^2} \cdot \frac{1-a_2p+b_2p^2}{1+a_2p+b_2p^2}$$
$$= \frac{(1-c_2\omega^2+c_4\omega^4) - j\omega(c_1-c_3\omega^2)}{(1-c_2\omega^2+c_4\omega^4) + j\omega(c_1-c_3\omega^2)} = G_4(\omega) \quad \ldots \quad \ldots (18)$$

where

$$L_1 = a_1R\,;\; C_1 = a_1/R\,;\; q_1 = a_1/\sqrt{b_1}\,;\; L_2 = a_2R\,;$$
$$C_2 = a_2/R\,;\; q_2 = a_2/\sqrt{b_2} \quad \ldots \quad \ldots \quad \ldots (19)$$

and

$$c_1 = a_1+a_2\,;\; c_2 = b_1+b_2+a_1a_2\,;\; c_3 = a_1b_2+a_2b_1\,;$$
$$c_4 = b_1b_2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (20)$$

Normally we should expect L_1, C_1, L_2, C_2 and R to be given, and we should then deduce a_1, b_1, q_1, a_2, b_2 and q_2 from Equ. (19) and c_1, c_2, c_3 and c_4 from Equ. (20). Here, however, we once again 'work backwards'. we define an angle ψ_4 [analogous to ψ_2 in Equ. (4)] by

$$\tan\psi_4 = \frac{\omega(c_1-c_3\omega^2)}{1-c_2\omega^2+c_4\omega^4} \quad \ldots \quad \ldots \quad \ldots (23)$$

so that

$$\sin\psi_4 = \omega(c_1-c_3\omega^2)/D_4\,;$$
$$\cos\psi_4 = (1-c_2\omega^2+c_4\omega^4)/D_4 \quad \ldots \quad \ldots (24)$$

where

$$D_4{}^2 = (1-c_2\omega^2+c_4\omega^4)^2 + \omega^2(c_1-c_3\omega^2)^2 \quad (25)$$

Thus D_4 is the modulus of the numerator and denominator of $G_4(\omega)$ in Equ. (18), and $G_4(\omega)$, like $G_2(\omega)$ in Equ. (1), always has unit modulus. The equation analogous to Equ. (6) is

$$G_4(\omega) = \frac{1-j\tan\psi_4}{1+j\tan\psi_4} = \frac{\cos\psi_4 - j\sin\psi_4}{\cos\psi_4 + j\sin\psi_4} = e^{-2j\psi_4} \quad (26)$$

and therefore the phase θ_4 of $G_4(\omega)$ is given by

$$\theta_4 = -2\psi_4 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (27)$$

Now Equ. (23), like Equ. (4), can be differentiated with respect to ω on both sides to give

$$\frac{d}{d\omega}\left[\frac{\omega(c_1-c_3\omega^2)}{1-c_2\omega^2+c_4\omega^4}\right]$$
$$= \frac{(1-c_2\omega^2+c_4\omega^4)(c_1-3c_3\omega^2)+\omega^2(c_1-c_3\omega^2)(2c_2-4c_4\omega^2)}{(1-c_2\omega^2+c_4\omega^4)^2} \quad \ldots \quad \ldots (28a)$$

and

$$\frac{d}{d\omega}[\tan\psi_4] = \frac{d}{d\psi_4}[\tan\psi_4]\frac{d\psi_4}{d\omega}$$
$$= \sec^2\psi_4\frac{d\psi_4}{dw} = \frac{D_4{}^2}{(1-c_2\omega^2+c_4\omega^4)^2}\frac{d\psi_4}{d\omega} \quad \ldots (28b)$$

Equating the last members of Equs (28a) and (28b), we have

$$\text{Group delay} = -\frac{d\theta_4}{d\omega} = 2\frac{d\psi_4}{d\omega} = \frac{2}{D_4{}^2}\{(1-c_2\omega^2+c_4\omega^4)(c_1-3c_3\omega^2)+\omega^2(c_1-c_3\omega^2)(2c_2-4c_4\omega^2)\} \quad (29)$$

We shall endeavour to find algebraically the most appropriate values for c_1, c_2, c_3 and c_4 so that the group delay $G_4(\omega)$ in Equ. (18) shall be maximally flat or shall have as little total variation as possible for $0<\omega<1$. Then we shall deduce from Equ. (20) the values of a_1, a_2, b_1, and b_2; this simply means that we have to factorize the expression

$$\phi_4(p) = 1 + c_1p + c_2p^2 + c_3p^3 + c_4p^4 \quad \ldots (21)$$

into two real quadratic factors

$$\phi_4(p) = (1 + a_1p + b_1p^2)(1 + a_2p + b_2p^2) \quad \ldots (22)$$

and this has been the subject of earlier "Mathematical Tools" (February and March 1957; May and July 1958), so that it need not be considered further here. Knowing a_1, a_2, b_1 and b_2 from Equ. (22), the values of L_1, C_1, L_2, C_2, q_1, q_2 are obtainable from Equ. (19) and therefore the circuit of Fig. 2 is determined. For applying algebraic techniques, Equ. (21) is a more convenient form for $\phi_4(p)$ than Equ. (22), and we shall therefore work entirely with Equ. (21) and the corresponding expression in terms of ω given in Equ. (18).

In order to evaluate the phase θ_4 of $G_4(\omega)$ in Equ. (18),

This is the expression whose variations we wish to minimize, but first it is convenient to change our notation and write

$$c_1 = x_0/\omega_0\,;\; c_2 = Kx_0{}^2/\omega_0{}^2\,;\; c_3 = Lx_0{}^3/\omega_0{}^3\,;$$
$$c_4 = Mx_0{}^4/\omega_0{}^4\,;\; X = \omega/\omega_0 \quad \ldots (30)$$

since x_0 then becomes virtually a scale factor; X varies from 0 to 1 over the range of interest, and K, L and M are the significant parameters controlling the way in which the group delay varies for $0\leqslant X\leqslant 1$. The group delay τ_4 given by Equ. (29) now reduces to

$$\tau_4 = \frac{2x_0}{\omega_0}\left[\frac{(1-Kx_0{}^2X^2+Mx_0{}^4X^4)(1-3Lx_0{}^2X^2)+x_0{}^2X^2(1-Lx_0{}^2X^2)(2K-4Mx_0{}^2X^2)}{(1-Kx_0{}^2X^2+Mx_0{}^4X^4)^2+x_0{}^2X^2(1-Lx_0{}^2X^2)^2}\right] \quad (31)$$

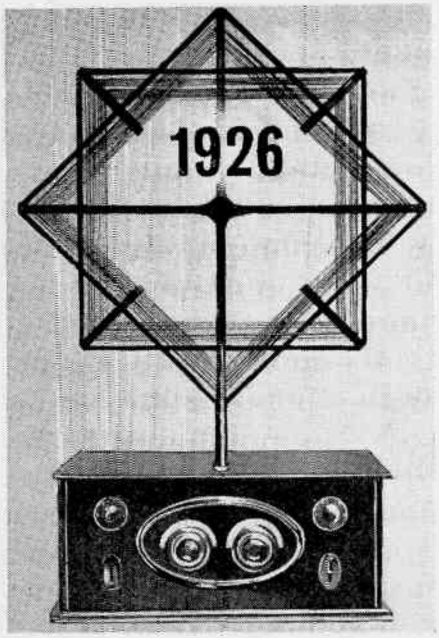
In next month's article we shall consider how to mould the group delay to be maximally flat or to have low total variation in the range $0<X<1$ by purely algebraic means analogous to those used for the group delay (for the bridged-T network of Fig. 1) given by Equ. (9).

REFERENCE

[1] G. G. Gouriet, "Two Theorems concerning Group Delay with Practical Application to Delay Correction," I.E.E. Monograph No. 275 R, December 1957.

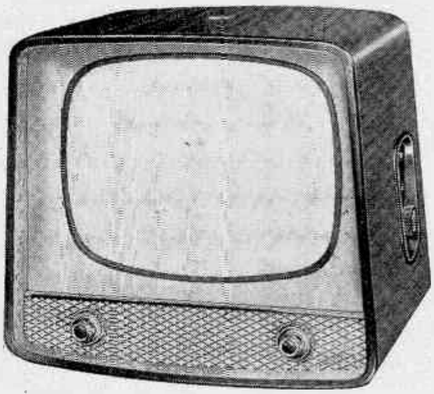
National Radio Show 1958

27th AUGUST—6th SEPTEMBER

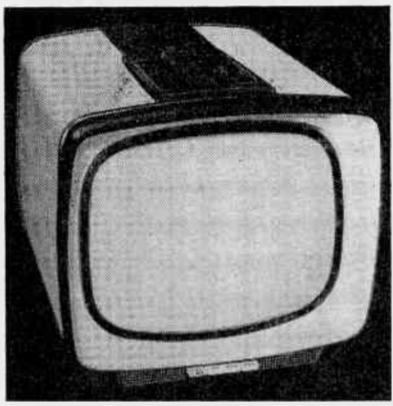


This year's exhibition marks the Silver Jubilee of the National Radio Show, the first being held in 1926. In these pages are illustrated typical examples of the current styling in domestic apparatus and, by way of contrast, we reproduce the photograph of an early superheterodyne taken from the show report in our October 1926 issue.

The contrast extends as much to the interior as to the appearance. Circuit techniques and methods of construction have changed as much as performance, reliability and convenience in 32 years of development.

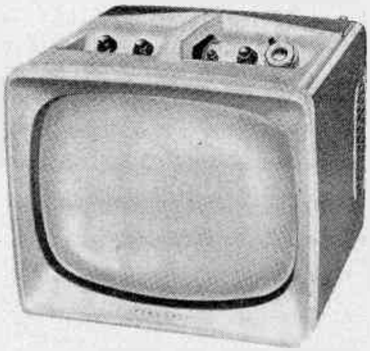


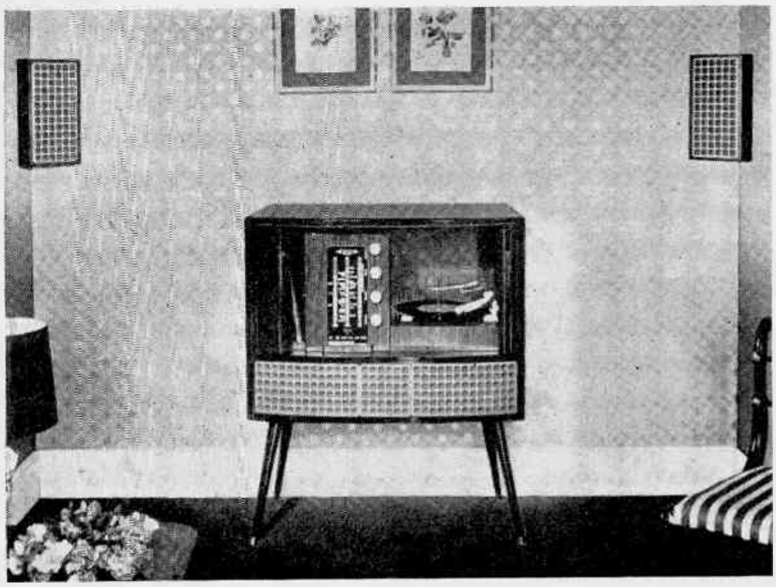
H.M.V. 1870 with 17-in. electrostatically-focused 90° tube. Printed circuits are used for the i.f. amplifier and for the line output stage



Bush TV80 transportable with 14-in. tube. Permeability tuning is used with push-button station selection. The four buttons enable the receiver to be set up for two Band I and two Band III stations. Each button gives full band coverage and can be set for any channel without tools

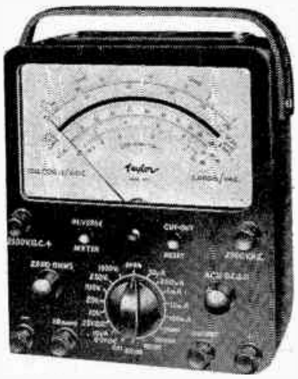
Below: Ferranti transportable television receiver with v.h.f. sound radio. A 14-in. tube is used and the total weight is 33 lb. The controls are recessed into the top of the case



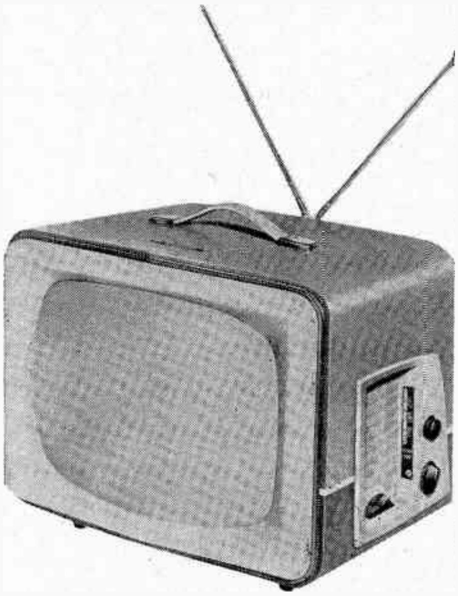


Decca SRG300 radiogramophone with extra loudspeakers for reproduction from stereophonic records. The cabinet contains a bass loudspeaker and two high-frequency units which form the speaker system for radio and single-channel records. For stereo, the internal h.f. units are cut out and replaced by external spaced units working in conjunction with the internal bass unit. The equipment includes twin push-pull amplifiers and v.h.f. as well as long and medium wavebands on radio

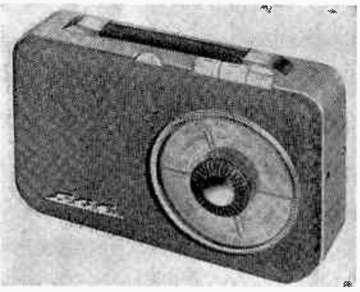
Taylor model 100 A multi-range meter with a resistance of $100{,}000\,\Omega/V$ *on the voltage ranges. It incorporates a* $9\text{-}\mu A$ *meter movement and the ranges cover* $0{\cdot}2\,\mu A$–$10\,A$, $0{\cdot}5\,\Omega$–$200\,M\Omega$, *and* $10\,mV$–$25\,kV$ *d.c. and* $2{\cdot}5\,kV$ *a.c. It has a 5-in. scale with an anti-parallax mirror*



Below: McMichael MP17 portable television receiver with v.h.f. sound receiver. It includes 'plated circuits' in which both sides of the base insulator are used with through connection where necessary. A special feature is made of accessibility for servicing, a hinged chassis member being employed



Left: Philco 3764 radiogram with ultra-linear push-pull output stage. The radio side covers the v.h.f. band only, but provision is made for the connection of the output of an a.m. tuner. Bass and treble speakers are used and the cabinet is designed to form a two-stage loaded enclosure



G.E.C. portable receiver of the mains/battery type weighing $5\frac{3}{4}$ *lb. Push-buttons are used for on/off and band selection*

Right: Bush SRG71 radiogramophone. The receiver covers medium and long waves as well as the v.h.f. band. Provision is made for recording and playing back through a tape recorder and also for the use of stereophonic equipment

Left: Sobell radiogramophone FMG57 covering the v.h.f., medium and long bands. Internal aerials are fitted. The record changer is of the four-speed type and has a crystal pickup

Coils for Magnetic Fields

PART 2: WEIGHT ASPECTS

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

SUMMARY. *The minimization of the total weight of a coil and power supply is studied both for a linear relationship between power and supply weight and also for the law obtained when the power-supply weight is principally a transformer. It is found that aluminium-foil solenoid systems have a weight advantage over copper systems not greater than 35% when minimization calls for small radius-ratio coils, but this disappears when larger coils are required. For large radius-ratio coils, copper coil systems weigh some 60% of the corresponding aluminium coil systems. Only when coil weight alone is important can aluminium coils be considered to have an advantage independent of radius-ratio.*

Microwave valves frequently require the production of long straight electron beams whose current density is such that external focusing means must be provided to prevent spread of the beam due to space charge forces. Although considerable work has been carried out recently on the use of electrostatic containment, this requires the presence of a set of electrodes which, with some notable exceptions[1,2], conflict with the wave-supporting structure with which the beam interacts.

Thus it seems likely that magnetic fields will remain an important means for focusing electron beams. Permanent magnets with suitable pole pieces or field straighteners are used where possible owing to the simplicity, stability, and absence of power supply. However, it is frequently found that sufficient field cannot be obtained from present permanent magnet materials with the demagnetizing forces set up in the required configuration, and resort to a solenoid supplied by direct current is necessary.

The presence of the power supply inevitably increases the weight of the system and it is important to design so that the weight is a minimum[3].

Solenoid Weight and Power

Normally the requirements are for a nearly constant field B, over a certain length L', near the axis of cylinder of radius r_1, which defines a volume from which the coil must be excluded for physical reasons. If the field is required to be uniform to within 5% over this length L', then it is necessary to make the solenoid of length L such that

$$L = L' + 4r \qquad \ldots \quad (1)$$

where r, assumed smaller than r_1, is the radius of the pole-piece hole at the ends of the solenoid. The power P required by the solenoid will then be[4]

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

where B is in kilogauss, ρ is the conductor resistivity in microhm/cm, L is in cm, R is the ratio of the outer radius of the coil to the inner radius, and s is the fraction of the coil section which is filled by conductor carrying current. In this form the power is seen to be independent of the type of construction (e.g., foil, wire or coaxial) except in so far as the filling factor s is affected, and independent of absolute radial dimensions.

Equ. (2) may be written

$$P = \Omega\left(\frac{R+1}{R-1}\right) \qquad \ldots \quad (3)$$

where

$$\Omega = (0{\cdot}796B)^2 \cdot L\left(\frac{\pi\rho}{s}\right) \qquad \ldots \quad (4)$$

and Ω is the minimum power required set up the field B over the required length. It is reached in the limit when the radius ratio R tends to infinity.

The weight of the coil W_c is given by

$$W_c = \pi \cdot r_1^2 L(R^2 - 1) \cdot [s(d_c - d_i) + d_i] \qquad \ldots \quad (5)$$

where d_c is the density of the conductor and d_i the density of the insulant. As this neglects the use of supports or polepieces it is practical, though not essential, to allow for this by assuming that the weight is the same as that obtained by taking the whole coil volume to have the density of the conductor so that we can write

$$W_c = d(R^2 - 1) \qquad \ldots \quad (6)$$

where

$$d = \pi r_1^2 L \cdot d_c \qquad \ldots \quad (7)$$

Power-Supply Weight

A reasonable assumption for the power-supply weight is to assume that it is proportional to the power required[3]; i.e.,

$$W = \alpha P \qquad \ldots \quad (8)$$

However this often gives a higher penalty for large power supplies than occurs in practice. For instance, a frequently occurring case is where the power supply consists of a transformer and rectifier system with

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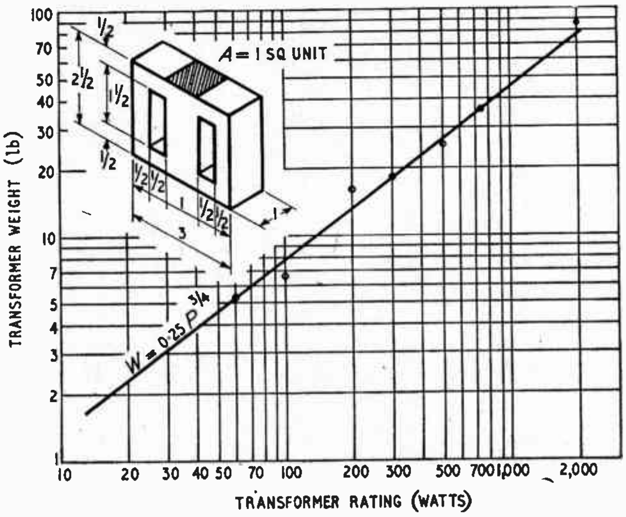

Fig. 1. *Theoretical and practical transformer weight versus power rating. The points marked are data from Gardners Radio Ltd., Catalogue NF47* (1957)

smoothing capacitor. Most of the weight is in the transformer.

Transformer design starts with a calculation of the core area, A (see Fig. 1), from the expression[5]

$$A = \frac{(P)^{\frac{1}{2}}}{C} \text{ square inches} \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

where C is a constant depending on the mains supply frequency. For 50 cycles it is 5·58.

For normal applications the proportions of the lamination stack are those given in Fig. 1.

$$\left.\begin{aligned} &\text{Thus if core area } A = l^2 \\ &\quad\text{then volume of laminations} = 6l^2 \\ &\qquad\text{volume of windings} = 4l^2 \text{ (square coil)} \\ &\qquad\text{weight of transformer } W = (6d_1 + 4d_2)l^3 \end{aligned}\right\} \quad (10)$$

where d_1 and d_2 are the densities in lb/cu. in. of the laminations and windings respectively.

Using Equ. (9) and Equ. (10) to eliminate A and l leads to

$$W = (6d_1 + 4d_2) \cdot C^{-3/2} \cdot P^{3/4} \text{ lb} \quad \ldots \quad (11)$$

$$\text{or } W = \alpha P^{3/4} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

The constant α is 0·25 for W in lb, or 113 for W in grams if both densities are taken as 9 and this gives excellent agreement with data taken from a transformer manufacturer's catalogue[6] (Fig. 1). However, if W is to be taken for the total power-supply weight α should be increased to take account of extra components and mountings. The index of 3/4 is of course obtained from the functional dependence in Equ. (9) and does not depend on any assumptions about the density.

Minimization

Taking the more general supply weight function

$$W = \alpha \cdot P^{\beta} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (13)$$

and using Equs (2), (3) and (6) we have to minimize the total weight

$$W = \alpha\Omega^{\beta}\left(\frac{R+1}{R-1}\right)^{\beta} + d(R^2 - 1) \quad \ldots \quad \ldots (14)$$

with respect to R.

Differentiating and equating to zero,

$$\frac{\alpha\beta\Omega^{\beta}}{d} = R_o(R_o - 1)^2\left[\frac{R_o + 1}{R_o - 1}\right]^{1-\beta} = f(R_o) \quad \ldots (15)$$

This determines the optimum radius-ratio R_o of the solenoid in terms of the known requirements of field, dimensions, power supply parameters and coil resistivity and density. $f(R_o)$ is shown in Fig. 2. The optimum radius-ratio increases with the power supply proportionality constant α, the field and the resistivity of the coil material. It decreases with increase of the internal diameter of the coil, and its density. It is independent of the coil length if $\beta = 1$, and decreases slightly with increase of L for $\beta = 3/4$.

The function $f(R_o)$ of Equ. (14) enables the relationship to be obtained between the optimum radius ratios R_{o1} and R_{o2} which apply when the same field is obtained from coils wound with different metals.

$$\frac{f(R_{o1})}{f(R_{o2})} = (\rho_1/\rho_2)^{\beta} \cdot (d_1/d_2)^{-1} \quad \ldots \quad \ldots \quad \ldots (16)$$

the important feature being the independence from the power supply penalty α, which sets the values of R_{o1} and R_{o2} without altering this relationship between them. If suffix 1 refers to copper and suffix 2 to aluminium this number is 5·94 for $\beta = 1$, and 5·13 for $\beta = 3/4$, the values being virtually independent of the temperatures reached by the coils as metals have nearly the same temperature coefficients of resistivity. The resulting graph of $(R_o)_{Cu}$ against $(R_o)_{Al}$ is shown in Fig. 3. It can be seen that the different values of β give insignificant differences in the relationship between the radius-ratios.

The normal procedure now would be to eliminate

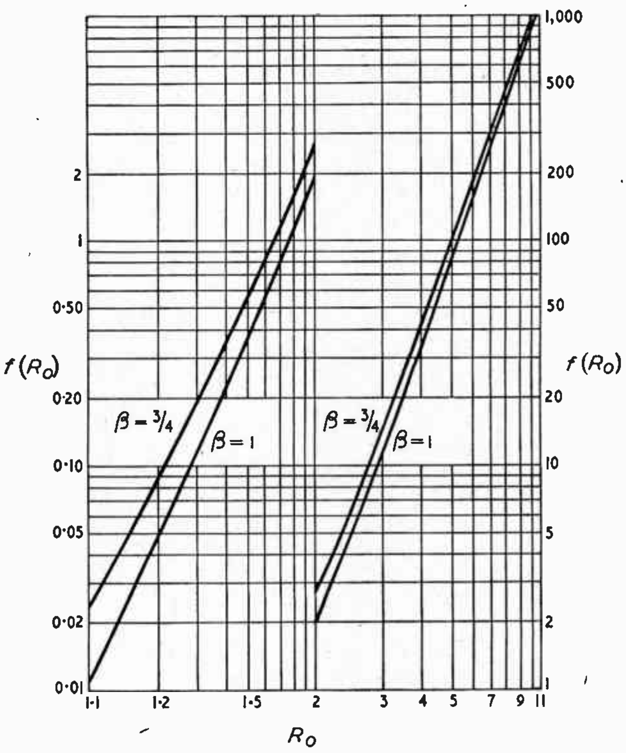

Fig. 2. *Function for determination of optimum radius-ratio for minimum total weight*

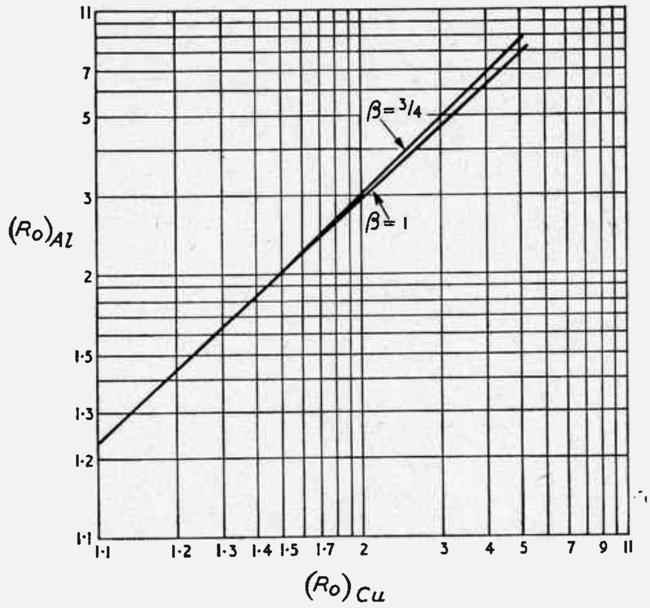

Fig. 3. *Radius-ratio of aluminium coil against radius-ratio of copper coil for minimum-weight systems*

R from Equs (13) and (14) to obtain the minimum weight but this is not possible analytically, as R cannot be obtained from either expression explicitly. The elimination can however be carried out in the limiting cases of R_o large or near unity, and gives

$$W_{min} = \alpha\Omega^{\beta} \text{ for } R_o \text{ large} \quad \ldots \quad \ldots \quad \ldots (17)$$

and $W_{min} = 2^{2\beta/(\beta+1)}\left(\dfrac{\beta+1}{\beta}\right)\cdot(\alpha\beta)^{1/(\beta+1)}\,(\Omega d)^{\beta/(\beta+1)}$ (18)

for R_o near unity.

The absence of the coil density from the first expression is a consequence of the fact that minimum weight systems will have large radius-ratio coils only when power-supply weight is very great in comparison with the coil, making it all important to keep the power-supply weight down via the last factor in Equ. (3). (The ratio of the power-supply weight to the coil weight in a minimum weight system is in fact just R_o/β).

The ratio of the minimum weights for a given field requirement with different metals is thus, in the two limits.

$$\frac{W_{1min}}{W_{2min}} = (\rho_1/\rho_2)^{\beta},\ R_o \text{ large} \quad \ldots \quad \ldots \quad \ldots (19)$$

$$\frac{W_{1min}}{W_{2min}} = \left(\frac{\rho_1 d_1}{\rho_2 d_2}\right)^{\beta/(\beta+1)} R_o \text{ near unity} \quad \ldots \quad \ldots (20)$$

For $\beta = 1$ the weight ratio of copper coil systems to aluminium coil systems is then 1·35 for R_o small, and 0·55 for R_o large, while for $\beta = 3/4$ the corresponding ratios are 1·295 and 0·64. Further investigation is required to show that the ratios remain within these limits for intermediate values of R_o and to determine where the advantage passes from the aluminium system to the copper system.

General Case

Although the minimum weight-ratio expression in the general case cannot be written down, it can be computed for any given pair of metals by using the relationship between the optimum radius ratios in Equ. (16).

Each of the corresponding pair of radius ratios is substituted into Equ. (14) with the appropriate density and resistivity constants d and Ω assigned and the ratio taken. In fact it is simpler first to eliminate $\alpha\Omega^{\beta}$ between Equs (14) and (15) to give

$$W_{min} = d(R_o^2 - 1)\left(\frac{R_o}{\beta} + 1\right) \quad \ldots \quad \ldots \quad \ldots (21)$$

and then to calculate the ratio

$$\frac{W_{1min}}{W_{2min}} = (d_1/d_2)\cdot\frac{(R_{o1}^2 - 1)(R_{o1} + \beta)}{(R_{o2}^2 - 1)(R_{o2} + \beta)} \quad \ldots (22)$$

The answer may equally well be plotted against either radius ratio R_{o1} or R_{o2}

For copper and aluminium the result has been plotted against the optimum radius-ratio for copper in Fig. 4, and the corresponding radius-ratio for aluminium written along the curves.

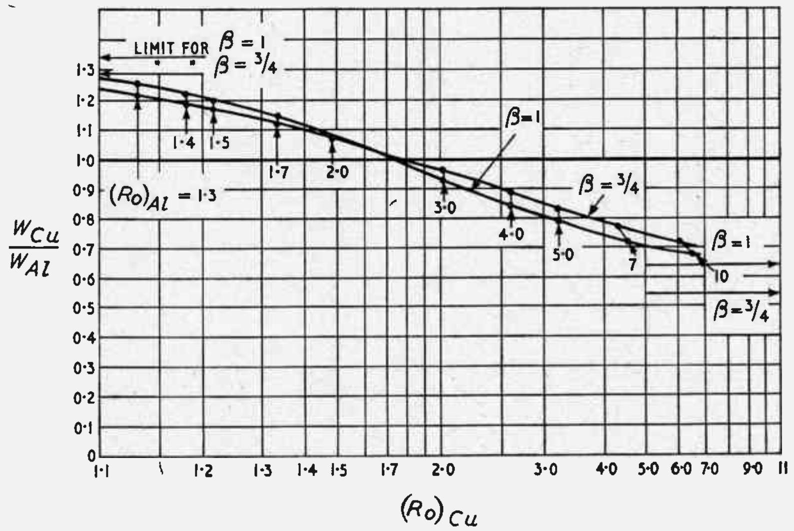

Fig. 4. *Ratio of minimum weights of copper and aluminium coil systems*

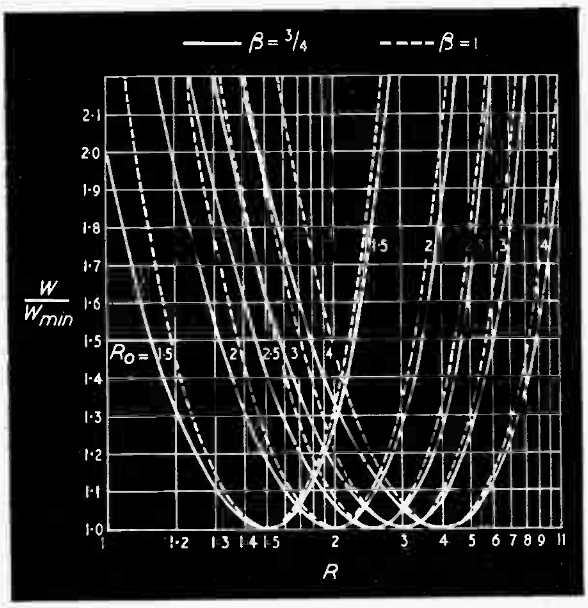

Fig. 5. *Weight increase for non-optimum coil radius-ratio*

The weight ratio of copper to aluminium is seen to decrease steadily between the limits previously derived and the aluminium-coil advantage disappears above about $(R_o)_{Cu} = 1 \cdot 7$, $(R_o)_{Al} = 2 \cdot 5$.

Design Latitude

How important is it to use the optimum radius-ratio for minimum weight? This may be assessed quite generally by dividing the general weight expression of Equ. (14) by the minimum weight expression Equ. (21) and eliminating $\alpha\Omega^\beta/d$ by using Equ. (15).

$$\frac{W}{W_{min}} = \frac{X\left(\dfrac{R+1}{R-1}\right)^\beta + (R^2-1)}{Y} \quad \ldots\ldots (23)$$

where the constants X and Y are given by

$$X = \frac{R_o}{\beta}\,(R_o-1)^2\left(\frac{R_o+1}{R_o-1}\right)^{1-\beta} \quad \ldots\ldots (24)$$

$$Y = (R_o^2-1)\left(\frac{R_o}{\beta}+1\right) \quad \ldots\ldots\ldots (25)$$

For a range of R_o Equ. (23) is shown in Fig. 5 for $\beta = 1$ and 3/4. Choice of the radius-ratio is seen to be fairly wide. It can be seen that if one were never to use any radius-ratio other than 2 : 1 the system weight would not exceed by 30% the minimum attainable if the optimum radius-ratio which should have been used was between about $1 \cdot 5$ and $3 \cdot 5$. This latter range corresponds (see Fig. 2) to a 40 : 1 variation in the coil and power-supply parameter $\alpha\Omega^\beta/d$ and must cover a large proportion of the cases arising in practice.

Coil and Power-Supply Weights in Minimum Total-Weight Systems

Having obtained the conditions for weight minimization one can calculate how the weight in such systems is divided between the coil and supply. As mentioned above in the section on minimization the ratio of the power-supply weight to the coil weight is R_o/β. This follows from taking the ratio of the two terms in the denominator of Equ. (23) and placing $R = R_o$. However it is perhaps of more interest to obtain the ratio of the coil weights and power-supply weights separately for the different metals. The results are given in Fig. 6.

It can be seen that the aluminium coil is always lighter whatever the radius-ratio required. For small radius-ratios the limiting values are the same as those for the total weight, as the weight divides between coil and supply in the constant ratio of $1 : 1/\beta$ for $R_o \approx 1$. The limits for large R_o are given by

$$\left(\frac{W_1}{W_2}\right)_{Coil} = (d_1/d_2)^{1/3} \cdot (\rho_1/\rho_2)^{2/3} \quad \ldots\ldots (26)$$

which gives $1 \cdot 004$ for $\beta = 1$, and $1 \cdot 11$ for $\beta = \frac{3}{4}$.

The power-supply weight ratio is in favour of copper coils beyond $(R_o)_{Cu} = 1 \cdot 35$, the limiting ratios being those of the total weight as given in Equs (20) and (21).

All-Important Coil Weight

This case sometimes occurs when it is necessary to mount a solenoid with a valve on a radar aerial, with

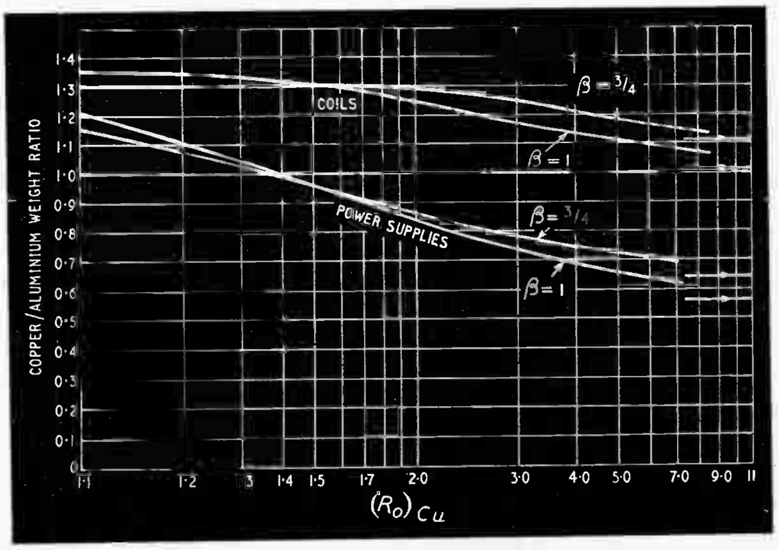

Fig. 6. *Ratio of coil and power-supply weight of copper and aluminium coil systems*

the power-supply distant and its weight unimportant. It is obvious that with no restriction on the power-supply weight, either coil could be made indefinitely small in radius-ratio and thus in weight. To choose between them we must apply some criterion and there are two to choose from. On the basis of coil size for the same weight the copper coil will be smaller, but as both will be near to unit radius-ratio anyway for small weight, this criterion is hardly significant. The remaining criterion is power consumption, or power-supply weight which amounts to the same thing. For equal coil weights

$$(R_1^2 - 1)d_1 = (R_2^2 - 1)d_2 \qquad \ldots \quad \ldots \quad \ldots (27)$$

$$\text{and } W_1/W_2 = \left[\frac{(R_1 + 1)(R_2 - 1)}{(R_1 - 1)(R_2 + 1)}\right]^\beta \cdot (\rho_1/\rho_2) \quad \ldots (28)$$

We only need these expressions in the limit of R near unity which leads to

$$W_1/W_2 = \left(\frac{\rho_1 d_1}{\rho_2 d_2}\right)^\beta \qquad \ldots \quad \ldots \quad \ldots \quad \ldots (29)$$

This gives aluminium the advantage, the copper power-supply being heavier by a factor of 1·83 for $\beta = 1$, and 1·57 for $\beta = \frac{3}{4}$.

All-Important Power-Supply Weight, or Consumption

This rather unlikely case clearly calls for the use of the largest possible radius-ratio coils, when the power consumptions tend to the Ω of Equ. (4). This leaves the powers in the ratio of the resistivities giving the copper coil the advantage with 0·55 times the consumption of the aluminium coil and the power-supply weights related by this factor raised to the power β, which is 0·64 for $\beta = \frac{3}{4}$.

Conclusions

As total weight minimization is not critical the designer can often ignore it, and apply other criteria such as coil size or power consumption. These two factors conflict for a given coil metal but it is clear that a high-resistivity metal such as aluminium is at a severe disadvantage compared with copper, as it is necessary to use a bulky coil to keep the power consumption nearer its minimum for the field required, and this minimum is 1·8 times that of copper anyway.

The possibility of achieving a better filling factor with aluminium by the use of alumina-film insulation would offset this disadvantage only slightly for with 0·00025 in. polyester film readily available, an increase of the filling factor to 100% would not be significant unless one were forced for some reason such as the necessity for a high impedance level to use foil thinner than 0·001 in. With the advent of semiconductor rectifiers capable of carrying high currents with negligible voltage drop this eventuality is remote. Copper coil systems are thus superior unless small coil weight alone is necessary.

Acknowledgements

This study was made as part of a programme sponsored by the Admiralty.

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[3] "Lightweight Aluminium Foil Solenoids for Travelling-wave Tubes". W. G. Worcester et al, *Trans. Inst. Radio Engrs*, ED-3, p. 70, Jan., 1956.

[4] See Part 1.

[5] "Radio Designers Handbook", Langford-Smith, Iliffe & Sons, 4th Ed., p. 235, 1953.

[6] "Transformers and Chokes". Gardners Radio Ltd., Christchurch, Hants, NF-47, 1957.

ERRATA

In Part 1 of this article, the following corrections should be made:

$\Delta^2\theta$ in Equ. (4) should be $\nabla^2\theta$

$L/2$ in Equ. (21) should be a subscript

In the substitutions for use in Equ. (28), in the right-hand column on p. 302, $W_n/2_n$ should be $W_n/2n$.

In Equ. (40), $\theta_1 + \theta_0$ should be $\theta_1 - \theta_0$.

In the footnote on p. 303, percentage current should be fractional current.

The reference to Equ. (45) immediately preceding Equ. (49) should be to Equ. (47).

The expression under the root sign in Equ. (51) should be

$$\frac{\Delta\theta_c}{1 + \alpha\Delta 0_c} \cdot \frac{1}{\Delta\theta_w}$$

In Equ. (53), B_{we} should be B_e.

On p. 306, line 1, column 2, "above" should be "about". In this same column 0·54/0·8 should be $\sqrt{(0{\cdot}54/0{\cdot}8)}$.

U.S. PROGRESS IN GUIDED MISSILES

The United States Air Force TM-76 Mace shown in the photograph has two interchangeable guidance systems. One employs inertial guidance (dead reckoning) and the other map-matching.

The map-matching system relates a film strip, actual or synthetic, to the terrain over which the Mace is flying. If there is any deviation from the programmed route, the missile's flight is adjusted to re-match the terrain with the film. The system permits the Mace to fly at extremely low altitudes, thus hindering enemy radar defences. The altitude can be varied during a flight.

The inertial guidance system gives complete invulnerability to enemy jamming. The missile can be flown over any course, including wide expanses of water, since no ground check points are necessary. The system compensates for changes in wind, temperature and barometric pressures, and is not limited by foul weather or unusual climatic conditions.

The missile can penetrate enemy territory at altitudes over 40,000 ft. or come in low at high speed with pre-selected detours. Another typical flight pattern is a combination of both high and low altitudes—the high to increase range, the low to avoid detection when near a heavily-guarded target. In addition to altitude changes, the inertial system can be pre-set to provide deviations from the true course.

(Courtesy the Martin Company, U.S.A.)

Bootstrap Circuit Technique

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

SUMMARY. *The normal amplifier, the bootstrapped amplifier, the cathode-follower and the anode-follower are shown to comprise a set of four circuits related to one another by simple circuit transformations. Three methods of excitation are distinguished. Each circuit may be put into feedback form, and the four basic feedback configurations applicable to bootstrap amplifiers are given. A number of practical examples are described.*

Although bootstrapping of amplifier valves is frequently resorted to in electronic circuits, and is an established design artifice, no comprehensive account of this practice appears to have been published; as a result, its wide range of usefulness may not be generally appreciated. There are numerous examples of the bootstrap connection in the literature, in addition to its well-known applications to integration, sawtooth generation and pulse amplification in radar systems, but in descriptions of these its characteristic effect has often either passed unnoticed or been attributed to cathode-follower action.

1. BASIC CIRCUIT FORMS

Notation

In specifying the forms of basic electronic circuits and their inter-relationships it has been found convenient to represent each electrode by a single letter, and to denote the method of connection of the valve between source and load by writing these letters in the order in which the corresponding electrodes occur in the intervening signal path[1]. Usually, one of the electrodes is earthed, and this connection is indicated here by using a capital letter for the symbol representing the earthed electrode. In the case of the thermionic triode *a*, *c*, *g* denote anode, cathode and grid, respectively; *b*, *c*, *e* represent base, collector and emitter, respectively, of the triode transistor. In this notation the normal, i.e., common-cathode (earthed-cathode), triode amplifier is specified as *gCa*; the corresponding transistor amplifier (earthed-emitter, common-emitter) connection is written *bEc*. Any one circuit form may be transformed into any other form by making the appropriate changes of connections; correspondingly, the letter sequence denoting any one connection may be transformed into any other by making the appropriate letter substitutions. For example, transposition of anode and cathode transforms the normal amplifier into the cathode-follower. Following mathematical convention such a transposition will be written *ac*, implying that $a \rightarrow c$ and $c \rightarrow a$. Thus, under the transposition $ac: gca \rightarrow gac$.

Ideal Voltage-Operated Forms

Derivation of the Bootstrap Amplifier

The basic bootstrap amplifier in its voltage-operated form may be derived from the normal amplifier (*gCa*) simply by shifting the earth connection from the cathode side of the load impedance to the anode side; i.e., from cathode to anode: it may, therefore, be described as the earthed-anode form of the common-cathode amplifier to distinguish it from the more familiar earthed-cathode form and denoted *gcA* [cf. Figs. 1 (a) and (b)]. This shift of the earth connection will be called the bootstrap transformation and represented as *Aa.Cc.* It does not alter the voltage-gain magnitude, or change the impedance properties of the circuit, but results in an output potential signal which appears, in phase with the input, at the cathode, rather than in anti-phase at the anode; i.e., the sign of the voltage gain expression is reversed from

$$A_{gCa} = \frac{-\mu Z}{r_a + Z} \rightarrow -\mu \text{ as } Z \rightarrow \infty$$

to

$$A_{gcA} = \frac{+\mu Z}{r_a + Z} \rightarrow +\mu \text{ as } Z \rightarrow \infty$$

'Follower' Circuits

Since shift of the earth connection results in fluctuation of the cathode, rather than of the anode, potential, it is necessary to consider the relationship of the bootstrap amplifier to the cathode-follower. The term follower is generally used to express the fact that the output signal of a circuit follows the input signal in magnitude, with little error. It is particularly appropriate to the cathode-follower circuit, in which the output electrode (the cathode) is also in phase with the input electrode (the grid), but is also applied to the earthed-cathode variant (the 'anode follower') because of its similar properties despite the fact that input and output are in anti-phase, the relative phase relationship generally being of little significance compared with voltage gain and output impedance.

Derivation of the Cathode-Follower

The cathode-follower may be derived from the earthed-cathode (common-cathode) amplifier by changing over the connections between the anode and cathode of the valve and the external circuit, as may be seen by comparing Figs. 1 (a) and (c). This is a fundamental change since it completely alters the gain and output impedance properties of the circuit; as is well known, the magnitude of the open-circuit voltage gain, i.e. μ, and the output impedance r_a are both reduced in

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* Coventry Technical College.

the ratio $1/(1 + \mu)$; moreover, the sign of the gain is reversed.

For earthed-anode (cathode-follower):

$$A_{gAc} = \frac{(+)\dfrac{+\mu Z}{r_a + Z}}{1 + \dfrac{+\mu Z}{r_a + Z}}$$

$$= \frac{+\mu Z}{r_a + (\mu + 1) Z} \rightarrow +1 \text{ as } \mu \rightarrow \infty$$

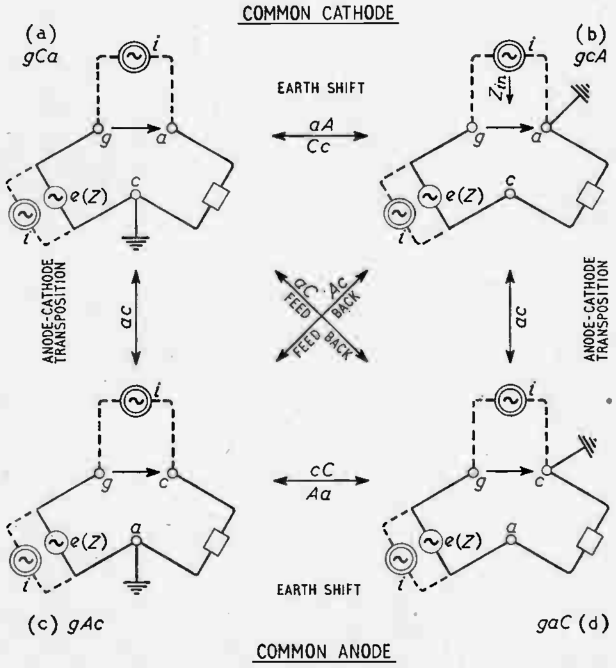
Similarly, for earthed-cathode (anode-follower):

$$A_{gaC} = \frac{(-)\dfrac{+\mu Z}{r_a + Z}}{1 + \dfrac{+\mu Z}{r_a + Z}}$$

$$= \frac{-\mu Z}{r_a + (\mu + 1) Z} \rightarrow -1 \text{ as } \mu \rightarrow \infty$$

The reduction by $1/(1 + \mu)$ in both gain and output impedance is characteristic of the effect of 100% negative feedback on an amplifier of gain $+\mu$; i.e., of the bootstrap form of common-cathode amplifier. Thus, the cathode-follower may also be derived by applying 100% negative series-shunt feedback to the bootstrap amplifier, which may be achieved by returning the lower side of the input source to the output electrode rather than to the common electrode; i.e., to the anode side rather than to the cathode side of the load[2, 3]. In this case, therefore, shift of earth connection from cathode to anode, plus application of 100% negative series-shunt feedback, is equivalent to transposition of the anode and cathode leads of the valve.

Fig. 1. *The basic forms of the normal amplifier (a), bootstrap amplifier (b), cathode-follower (c), and bootstrapped cathode-follower (or anode-follower) (d). The two alternative methods of current operation, in which the impedance Z replaces the voltage source e, are shown dashed*



Distinction between Bootstrap Amplifier and Cathode-Follower

It will be clear from the foregoing, and from a comparison of Figs. 1 (b) and (c), that, provided the name cathode-follower is restricted to the form in which the cathode follows the input in magnitude as well as in phase, there is a clear-cut distinction between it and the voltage-operated bootstrap amplifier; the latter might well be called a cathode-output or cathode-loaded amplifier but cannot be called a cathode-follower amplifier unless the term follower implies only an in-phase relationship between input and cathode potentials. In the literature the 100% feedback form of cathode-output stage is often referred to as the cathode-follower amplifier, which is unfortunate from either point of view, not only because the feedback form has an external gain which is less than unity but also because the non-feedback form, which does provide external gain, has come to be called the bootstrap amplifier and is a variant of the normal amplifier.

Anode-Follower (Bootstrap Cathode-Follower)

Since the present paper is concerned primarily with the bootstrap transformation it is of interest to note that this may equally well be applied to the cathode-follower, with the result shown in Fig. 1 (d), which will be called the voltage-operated form of the anode-follower circuit, so-called because its gain and impedance properties are the same as those of the circuit from which it is derived, except for reversal of the phase of the output signal. This simple relationship does not appear to have been widely noticed, possibly because the anode-follower is usually current-operated, as will be explained in the next section[4].

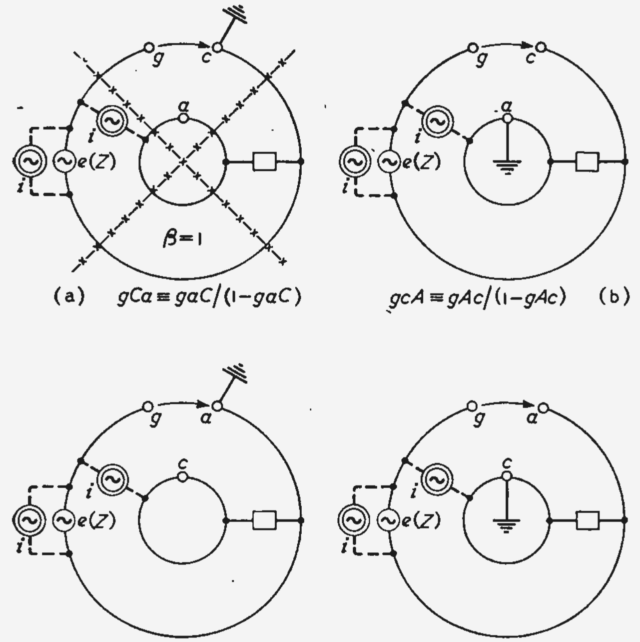
Correlation of Voltage-Operated Forms

The four circuits contained in Fig. 1 comprise a complete set which are related to each other, in the manner indicated by the arrows placed between them, by the three following transformations:

1. Shift of earth lead from the common to the output electrode (the bootstrap transformation),
2. Application of 100% series-shunt type negative feedback, by shifting the lower side of the input from the common to the output electrode (the feedback transformation), and
3. Anode-cathode transposition, which is the product of (1) and (2).

Ideal Current-Operated Forms

The voltage-operated forms of the bootstrap amplifier and bootstrap cathode-follower circuits both suffer from the practical disadvantage of having a floating input; accordingly, single-ended-input forms have been developed, the essential feature of which is the use of a high-impedance (ideally infinite impedance; i.e., constant-current) source. These current-operated forms are obtained by substituting a passive impedance (Z), preferably a pure resistance, for the ideal voltage generator, and supplying it with an input current from the constant-current source. It is not necessary to connect the source directly across the impedance; provided the latter alone is directly connected to one side of the source, the input current may be allowed to return to the other source terminal indirectly through the remainder of the network. In the basic circuit form given in Fig. 1 the



(a) $gCa \equiv gaC/(1-gaC)$ $gcA \equiv gAc/(1-gAc)$ (b)

(c) $gAc \equiv gcA/(1+gcA)$ $gaC = gca/(1+gCa)$ (d)

Fig. 2. *The loop forms of the basic circuits given in Fig.* 1. *The parts (a)–(d) of the two figures correspond*

only alternative indirect path of this kind is through the load impedance. In the single-ended current-fed form of bootstrap amplifier or anode-follower the input current traverses the input impedance to develop the input voltage and then returns through the load across which the output voltage is developed. If the input voltage is taken as $v_i \equiv i \,.\, Z$ the voltage gain will be v_o/iZ and will be greater than for the voltage-operated form because of the additional output voltage produced by the input current in its passage through the load.

A corresponding superposition of input and output may, however, be obtained in the voltage-operated form by taking output from the upper input terminal so that the input voltage adds to the output.

These alternative current-operated forms are indicated by dashed connections in Fig. 1. It will be noted that when the current sources are ideal (i.e., of infinite impedance) the impedance properties of these forms are identical with those of the voltage-operated forms. The single-ended current-operated bootstrapped form of the cathode-follower will be recognized as the usual form of the anode follower.

Impedance Multiplication

The bootstrap connection of the common-cathode amplifier with single-ended current input is generally employed not so much because of its signal polarity properties (such as lack of signal inversion in the bootstrap amplifier) as for its impedance-multiplication property. It is a matter of straightforward circuit analysis to show that, for a resistive load R, the input impedance with this method of operation is

$$Z_{in} = \frac{rR + (\mu + 1)\,RZ + Zr}{r + R} \approx \left(\frac{\mu R}{r + R}\right) Z \text{ for } \mu \gg 1$$

from which it will be clear that the input impedance is approximately equal to the product of the grid–cathode impedance element Z into the voltage gain of the bootstrap amplifier. Thus, instead of thinking of the current-operated bootstrap amplifier as having a grid–cathode impedance element to develop an input voltage, prior place may be given to the impedance element itself, and the amplifier regarded as having been added to it in order to magnify its effective impedance. When the current-operated bootstrap is used for this purpose it is connected as a two-terminal impedance, its two terminals being grid and anode (earthy), as shown in Fig. 1 (b). In a simplified form, where the input current is passed through the valve rather than through an external load impedance (with which the anode–cathode path of the valve is in parallel), the input impedance expression reduces to:

$$Z_{in} = r + (\mu + 1)\, Z$$

since in this case R is infinite.

In this application the bootstrapped valve is usually employed as the load impedance of a pentode amplifier, as will be described in Part 3.

Feedback Forms

Since feedback considerations enter into the comparison of practical forms of the circuits given in Fig. 1 it is of interest to consider first the feedback forms of the ideal circuits; these are given in Fig. 2. It will be seen that the feedback forms of the normal and bootstrapped amplifiers involve *gac*; i.e., common-anode transmission through the valve; conversely, the normal and bootstrapped forms of the cathode-follower which use the *gac* path in direct transmission involve *gca* transmission in the feedback form. Thus, direct *gca* transmission may be regarded as loop *gac* transmission with 100% positive feedback, while direct *gac* transmission is loop *gca* transmission with 100% negative feedback[2, 3]. Algebraically,

$$\text{(a)} \quad A_{gCa} = \frac{A_{gaC}}{1 - A_{gaC}}$$

$$\text{(b)} \quad A_{gcA} = \frac{A_{gAc}}{1 - A_{gAc}}$$

$$\text{(c)} \quad A_{gAc} = \frac{A_{gcA}}{1 + A_{gcA}}$$

$$\text{(d)} \quad A_{gaC} = \frac{A_{gCa}}{1 + A_{gCa}}$$

The loop forms given in Fig. 2 show more clearly than the direct forms the fact that the basic circuit form is the same for all three methods of operation. It will be noted that the feedback configuration in all four cases is of the series-shunt type with unit-β feedback, as indicated in the crossed-chain-line divided circuit (a).

Practical Sources

Practical sources are always non-ideal in that they do not have zero or infinite internal impedances. It is convenient, therefore, to think of voltage sources only, with internal impedances which are non-zero but finite, and to regard them as good approximations to the idealized sources when their impedances are very low or very high in relation to their load impedances. Moreover, in practical valve and transistor circuits the

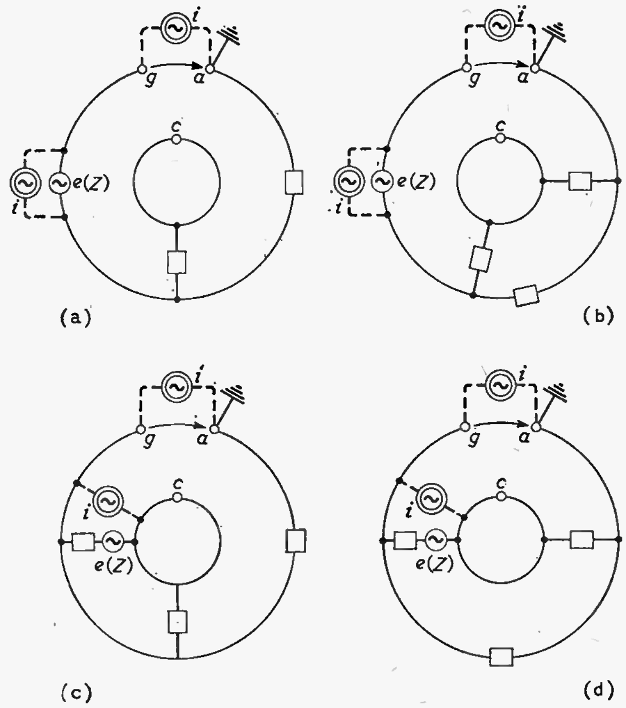


Fig. 3. *The four basic methods of applying feedback to the bootstrap amplifier [part (b) of Figs. 1 and 2]: (a) series-series, (b) series-shunt, (c) shunt-series and (d) shunt-shunt*

form is rarely so simple as shown in Fig. 1; there will always be a path of non-zero admittance across the valve, between the upper input and output terminals. This may be due to the presence of an actual resistance, such as a grid leak resistor in a cathode-follower circuit, or to leakage, or to stray and/or interelectrode capacitance.

The effect of this path will be to provide shunt-shunt feedback between output and input. In terms of the direct transmission form shown in Fig. 1 this feedback will, for a resistive circuit, be negative for *gca* transmission [(a), (b)], but positive for *gac* transmission [(c), (d)]; in terms of the feedback forms this additional feedback path will be opposed to the existing inherent feedback in both cases, so that, if it exceeds 50%, it will have the effect of causing a reversion from *gca* to *gac* transmission, or vice versa. When, for example, the impedance between grid and cathode of a cathode-follower is smaller than the source impedance the circuit properties are more nearly those of the ideal bootstrap amplifier than of the ideal cathode-follower, since the 100% negative feedback by which the cathode-follower is obtained from the bootstrap amplifier is largely cancelled by the positive feedback through the grid–cathode admittance. It follows that, for the classification made in Figs. 1, 2 to hold, the source impedance with voltage operation must be small compared to any impedance bridging input and output, while for current operation the input impedance element used to develop an input voltage from the practical current source must have a small impedance compared with that of the source. There is no difficulty in obtaining these conditions in practice; the present point is that these are the conditions which determine the circuit action and hence the naming and designation of the circuit.

Application of Feedback

In many applications of the bootstrap amplifier feedback is applied, and it will be clear that, in so far as the bootstrapped form differs only by the earth shift from the normal amplifier, there will be a bootstrapped version of each method of applying feedback to the normal amplifier. There are four distinct feedback amplifier configurations[5]; viz., series-series, series-shunt, shunt-series, and shunt-shunt; these are shown in basic form at (a)–(d), respectively, in Fig. 3. Practical examples of these various forms will be given in the later parts of this paper.

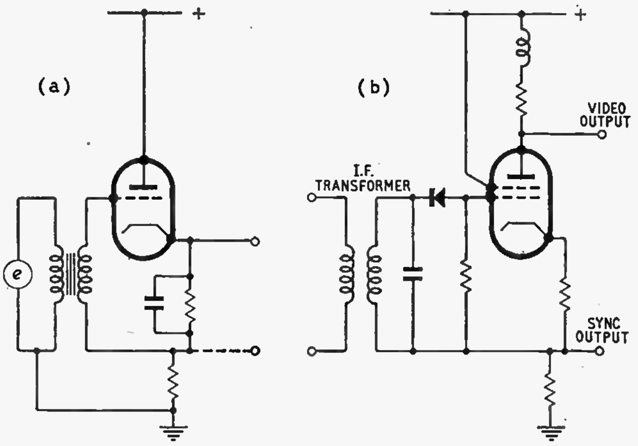
When feedback is applied there is an advantage to be gained in replacing the single triode by a multi-stage amplifier having a higher gain. Two two-valve examples are of particular interest: in each the gain of the single valve is augmented by associating with it an additional amplifier valve; in one a cathode-input amplifier is inserted between the existing amplifier and load, thereby forming the so-called cascode arrangement, in the other the additional valve is itself a bootstrap amplifier. Since both additional amplifiers are non-inverting the overall gain is increased without incurring reversal of output signal polarity; moreover, both arrangements admit of series h.t. feed. In the case of the added bootstrap amplifier there are two variants, according as the load is fed from the grid or cathode of the bootstrap: in the former connection the bootstrapped valve is used in a two-terminal impedance network, with the advantage that it effectively increases the effective magnitude of the existing amplifier load, as already explained under 'Impedance Multiplication'; with the latter connection the bootstrapped valve acts in a transfer network, and the output current has to traverse its grid–cathode impedance element before passing through the load. Examples of these two-valve arrangements will be given later.

2. VOLTAGE-OPERATED BOOTSTRAP AMPLIFIERS

Transformer Input

In its practical form the basic voltage-operated bootstrap amplifier has its load impedance and h.t.

Fig. 4. *Transformer-coupled voltage-operated bootstrap amplifiers: (a) single output, (b) double output*



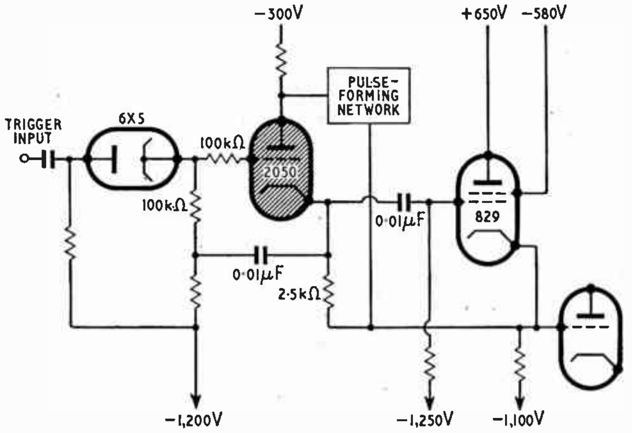


Fig. 5. *Bootstrapped radar pulse modulator. (Slightly modified version of circuit given in "Principles of Radar", 2nd edn.). The two halves of the* 6*X*5 *and of the* 829 *are paralleled*

supply transposed by comparison with the basic normal amplifier from which it is derived; otherwise the shift of earth connection from cathode to anode would leave the h.t. supply floating. Moreover, it is usual to earth the anode indirectly, through the supply, by connecting the earth lead to the negative end of the supply. Even so, the source of signal input voltage remains entirely floating, and this feature constitutes the one drawback of the bootstrap amplifier. This difficulty may, of course, be overcome by interposing a transformer, as shown in Fig. 4 (a), but at the cost of incurring additional shunt capacitance across the signal path due to the p.d. between the secondary as a whole and earth, particularly if a step-up transformer is used for matching purposes, or to obtain increased gain; also, the leads between the transformer and valve must be kept as short as possible. When only a non-inverted output of greater-than-unity gain is required, and the use of a transformer is permissible, the voltage-operated bootstrap has no advantage over the normal amplifier, since the latter could be preceded by an inverting transformer, and would not have the disadvantage of a floating secondary winding. There are occasions, however, when a pair of amplified outputs are desired, and a useful feature of bootstrap voltage drive is that with a load in both anode and cathode leads such outputs may be obtained. Also, there are situations in which a transformer already exists in the circuit preceding the amplifier, as in the case of the detector circuit of a radio or television receiver in which the diode detector is almost invariably fed from the final i.f. transformer. In a sound receiver the detector may be bootstrapped to the a.f. amplifier, and the latter loaded equally at anode and cathode to give an amplified pair of anti-phase outputs; in a vision receiver one output may be used to feed the picture tube, the other to supply the sync. separator[8]; such an arrangement was, in fact, employed in a pre-war Cossor television receiver [see Fig. 4 (b)].

Pulse Amplifiers and Modulators

In radar practice the bootstrap amplifier has been used as a driver for pulse modulators[6, 7]. When a valve is required to handle large current pulses it is desirable that its quiescent current (i.e., the current which flows between pulses) should be small, and that the valve should be driven into heavy current only during the applied pulse, rather than stand at a heavy current and be cut off by the pulse; the latter must therefore be positive-going. If, however, output is taken from the anode of the pulsed valve for application to the following stage it will be negative-going; with bootstrap connection of the input to the pulsed driver valve the output to the next stage will, however, be positive-going, as required. In radar modulator circuits one or more of the stages immediately preceding the modulator may be bootstrapped in this way; a typical skeleton circuit is shown in Fig. 5, in which the gas-filled valve is bootstrapped to a positive trigger-pulse source at its input, and also to a following driver stage at its output.

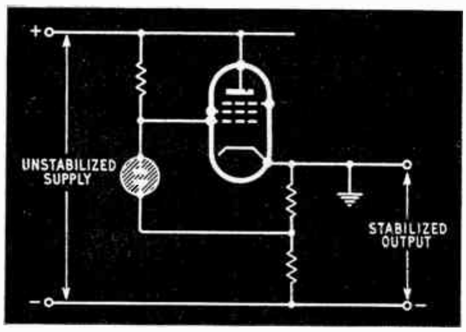
Application of Feedback: Bootstrap Voltage Regulator

Feedback may, of course, be applied to the voltage-operated bootstrap amplifier in any of the four basic ways shown in Fig. 3. Thus, an unbypassed resistor (i.e., series-series feedback) is often used in the bootstrap amplifier for bias purposes, as in Fig. 4 (b), but the shunt-series feedback configuration 3 (b) is rarely encountered. An example of series-shunt feedback occurs in the voltage-regulator circuit given in Fig. 6. This circuit is useful for cases where the output voltage is considerably greater than that of the gas-filled voltage reference tube and it is desired to earth the positive side of the output, as in the case of high-voltage supplies for cathode-ray tubes and television cameras[9].

The Bootstrap Integrator

The most common application of the shunt-shunt feedback arrangement occurs in the well-known bootstrap integrator[10–14], whose basic form is shown at (a) in Fig. 7. As noted by F. C. Williams[12], this circuit may be obtained from the basic single-valve Blumlein-Miller integrator by the earth-shift load-supply transposal transformation. The usual practical form of this circuit is shown at (b) in which the input voltage source of (a) has been replaced by the capacitor C, which is charged intermittently through the diode V_2; the integrating capacitor is discharged periodically through V_3. When this circuit is used to generate a sawtooth

Fig. 6. *Voltage-operated bootstrapped regulator*



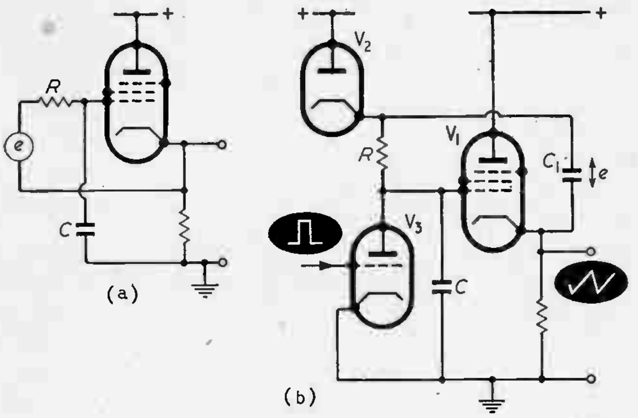


Fig. 7. *The bootstrap integrator: (a) basic circuit, (b) usual practical form*

wave V_3 is operated by a positive pulse wave at its grid. V_3 conducts during pulses, and the negative change at its anode produces a similar change at V_1 cathode, which change switches on diode V_2, thereby recharging the input capacitor while the integrating capacitor is being discharged through V_3. When V_3 turns off, on cessation of the input pulse, the voltage rise at V_1 cathode turns off V_2 also, leaving capacitor C_1 to supply charging current through R to the integrating capacitor C. The integration error is small but may be reduced further by including a second integrator the resistor of which is usually shunted by an additional discharge diode. A further development is the incorporation of the discharge valve V_3 in a monostable multivibrator ('flip-flop') circuit.

Increase of Bootstrap Amplifier Gain

It is desirable in these feedback circuits that the bootstrap amplifier over which the feedback is applied should have a high voltage gain. There are two ways in which a second valve may be associated with the bootstrap in such a way as to increase its gain without incurring phase reversal. In one the additional valve is connected in cascode with the bootstrap; the complete circuit is, indeed, a bootstrapped cascode amplifier, obtained from the normal form of the cascode by the usual bootstrap transformation; it is shown at (a) in Fig. 8. (It may be of interest to note that if the lower side of the input-voltage source is connected to the other end of the load, this circuit becomes the cascode cathode-follower.) In the other method (b) the load itself is also bootstrapped (in the single-ended current-operated sense) in order to multiply its effective impedance and thereby make the stage gain approach more closely the limit set by the amplification factor of the valve. With these modifications the gain of the basic bootstrap amplifier is increased (provided $R \gg r_2/\mu_2$) from $+\mu R/(r_1 + R)$ to:

$$\text{(a)}: \quad + \frac{\mu_1 R}{r_1 + R} \cdot \frac{(\mu_2 + 1)(r_1 + R)}{(1 + \mu_2) r_1 + r_2 + R}$$

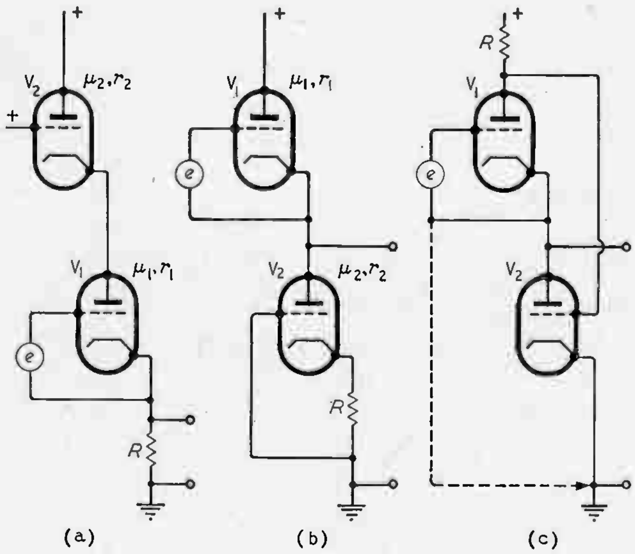
and

$$\text{(b)}: \quad \frac{\mu_1 R'}{r_1 + R'}, \text{ where } R' = r_2 + (\mu_2 + 1) R$$

respectively. Method (b) will be mentioned again later as an application of the current-operated bootstrap.

In an important variant of the bootstrap-loaded bootstrapped amplifier shown at (b) the load is connected across the additional valve V_2, by shifting the lower output connection and the earth lead to the other end of the coupling resistor R; the latter may then be moved to the other side of the h.t. supply; i.e., into the anode lead of V_1, giving the arrangement shown at (c). The 100% feedback form of this circuit, obtained by connecting the lower input terminal to the bottom of the load, is the E. L. C. White stacked cathode-follower, which will be mentioned again in a later section. If the earth connection in (b) and (c) is shifted to the other side of the load, and V_2 moved to the other side of the h.t. supply, the two variants of the normal-amplifier driven bootstrapped amplifier given in Fig. 10 (a) will be obtained.

Fig. 8. *Use of additional valve to increase bootstrap amplifier gain: (a) the bootstrapped cascode amplifier, (b) the bootstrap amplifier with bootstrapped (current sense) load, (c) another form of the bootstrap-loaded bootstrap amplifier. (The dashed line indicates modification required to obtain* 100% *feedback, as in White's cathode-follower)*



Transistor Forms

One method of stabilizing the operating conditions of a junction transistor amplifier operating in the common-emitter connection is to apply a potential to the base in the manner shown at (a) in Fig. 9, and insert a resistor in the emitter lead. This resistor may have a relatively large value; accordingly, feedback of the signal potential developed across it is sometimes eliminated by returning the lower side of the input source to the emitter rather than to the other side of the emitter current-defining resistor. In the examples shown at (a) and (b) output is taken from the collector alone, or from both collector and emitter, but in the third example at (c) only emitter output is developed[15, 16].

3. CURRENT-OPERATED BOOTSTRAP VALVE CIRCUITS

Methods of Coupling

The current-operated bootstrap valve is generally employed as the anode load impedance of a normal

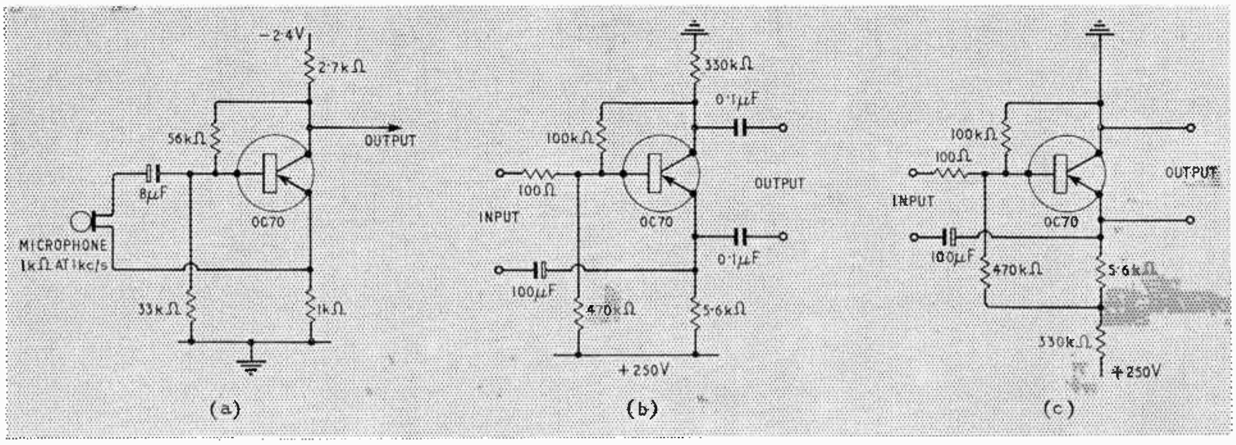


Fig. 9. *Bootstrapped transistor circuits*

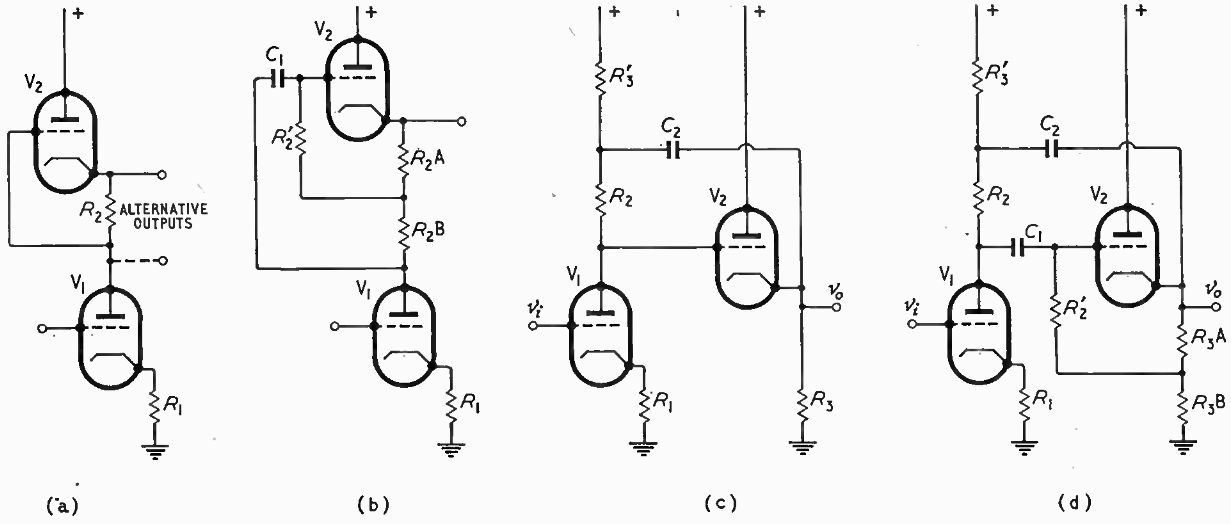
amplifier stage. There are four distinct arrangements according as the grid and cathode of the bootstrapped load valve are a.c. or d.c.-coupled to the anode circuit of the driver, as follows (see Fig. 10):

(a) d.c. couplings to both cathode and grid, (d.c.–d.c.)
(b) a.c. coupling to grid, d.c. coupling to cathode, (a.c.–d.c.)
(c) d.c. coupling to anode, a.c. coupling to cathode, (d.c.–a.c.)
(d) a.c. coupling to both cathode and grid. (a.c.–a.c.)

In (a) and (b) the two valves may have series d.c. feed, as shown in Fig. 10, while (c) and (d) have parallel feeds; (a) and (b) may also be arranged for parallel valve feeds by replacing C_1 and/or C_2 by (a) constant-voltage two-electrode tubes, or by using a resistive-divider coupling. In all cases the bootstrapped valve may be used purely as a two-terminal load impedance, with output from the anode of the driver valve, or as a transfer stage, with output from the cathode of the bootstrapped valve itself. The difference in gain is small when $R_2 \ll (\mu + 1)\, r_a$, but the latter arrangement is generally preferred, despite any gain loss, when it is desired to obtain fast positive-going transitions in circuits handling pulse- or step-type signals, since the bootstrapped valve may then charge the load capacitance directly, rather than through the coupling resistor R_2.

The simple form (a) is restricted by the fact that R_2 not only acts as the coupling impedance but also provides bias for V$_2$, and a suitable single resistance value for both purposes may not be feasible unless R_2 is required to be small, as in a wide-band circuit. Apart from this point it is the most convenient circuit when d.c. coupling cannot be avoided. It has been used by Philips as an audio output stage[17] driving a high-impedance loudspeaker. With a.c. signals, and for cases where the load is required to be much larger than the bias value, the form (b) is more suitable [see Fig. 17 (b)]. When the parallel-feed form must be used, because of lack of adequate h.t. voltage, for example, circuit (c) has the advantage over (d) of needing only one *CR* coupling and of allowing V$_2$ to operate into a high cathode resistance without the need for a negative rail. Direct

Fig. 10. *An amplifier valve* (V_1) *having a single-ended type current-operated bootstrapped impedance load* (V_2): (*a*) *d.c.–d.c. coupling*, (*b*) *a.c.–d.c. coupling*, (*c*) *d.c.–a.c. coupling*, (*d*) *a.c.–a.c. coupling*



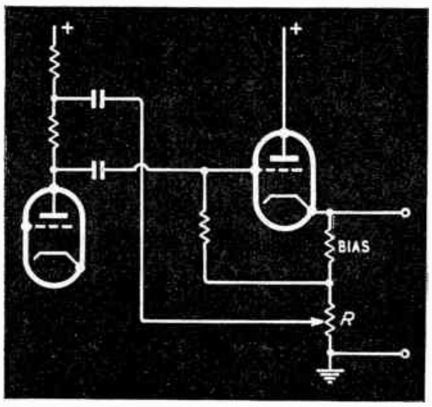


Fig. 11. *A feedback method of controlling the gain and output impedance of the bootstrap-loaded amplifier*

coupling of V_2 may, however, make a *CR* coupling in the output lead more necessary than for (d), in which such a coupling could in some cases be avoided by arranging the quiescent cathode voltage to be zero. With a negative rail available $R_{3A,B}$ of (d) may be combined into a single resistor and the lower side of R_2' returned to earth, or to a divider across the negative supply. An example of a.c.–a.c. coupling (d) is given in Fig. 14 (b).

Gain Control

A convenient method of gain or output impedance control for the double a.c.-coupled parallel-feed circuit is shown in Fig. 11; with the slider at its lower extreme position both gain and output impedance are minimum, and increase to a maximum as the slider is raised, the range of control obtained depending on the relative values of R and r_a of V_1.

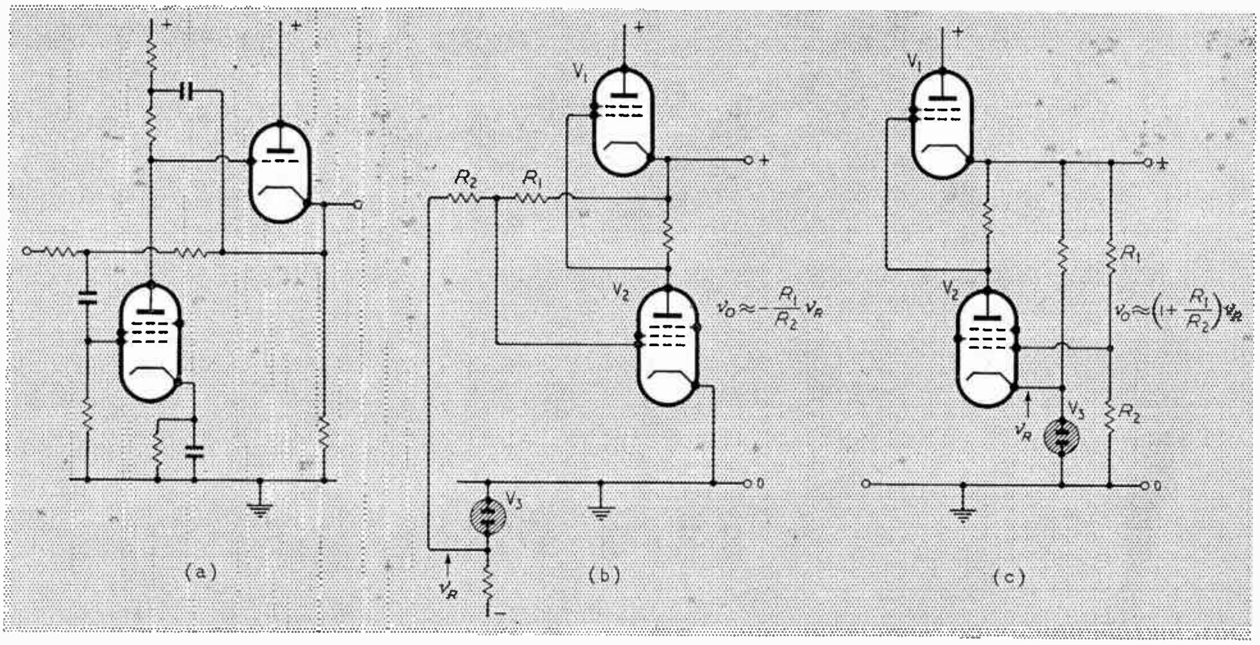
Application of Feedback

In all of these circuits feedback may, of course, be applied in any of the ways employed over a normal amplifier stage since the bootstrapped valve leaves the output signal phase substantially unchanged. From the feedback point of view the use of the bootstrap connection is of importance since it allows the forward gain to approach more closely the limit μ set by the amplification factor of the driver valve, particularly when the latter has a very high r_a, as with a pentode. The most widely used feedback arrangement is the shunt-shunt form shown at (a) in Fig. 12. The most common application of this type of feedback occurs in h.t. voltage-stabilizers of the kind shown in Fig. 12 (b), in which the series regulator valve V_1 is bootstrapped to the load of the amplifier V_2 and feedback applied over the potential divider R_1, R_2 to the voltage reference tube V_3. The latter in effect provides a d.c. signal (v_R) into the amplifier, the high gain of which maintains the output potential closely equal to R_1/R_2 times the reference voltage, even when a relatively low-impedance load is imposed on the output terminals. In the more familiar form at (c) the same components are rearranged into the series-shunt form of feedback configuration, for which the output voltage is closely equal to $(1 + R_1/R_2)\, v_R$.

Bootstrap Driver for the Cathode-Follower

The current-operated bootstrap transfer stage has been used as a driver for a cathode-follower. When a cathode-follower works into a relatively low load its input impedance may not be much greater than that of the same valve operated as a normal (i.e. common-cathode) amplifier stage. For example, when a cathode-follower works into a matched load its gain is one-half and its input impedance is then only double that obtained with common cathode operation, so that input capacitance may be excessive, particularly when the cathode-follower is fed from a signal source of high impedance. Also, in such a situation the cathode-follower would act as a current-operated bootstrap amplifier, and would therefore not have the desired low

Fig. 12. *Application of negative feedback to the bootstrap-loaded amplifier:* (*a*) *shunt-shunt feedback applied to a voltage amplifier,* (*b*) *shunt-shunt feedback regulator,* (*c*) *series-shunt regulator*



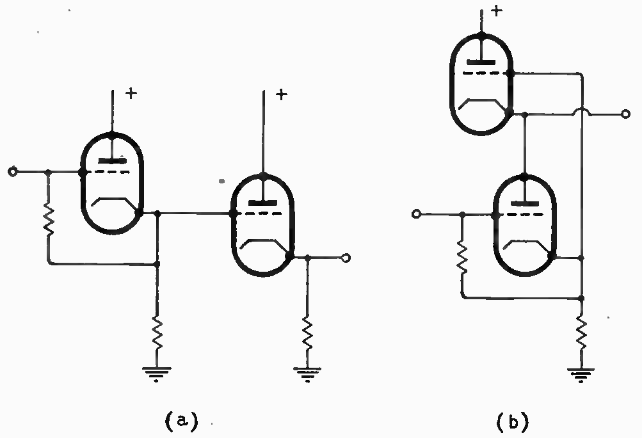


Fig. 13. *The bootstrap driven cathode-follower: (a) parallel d.c. feed form, (b) series d.c. feed form*

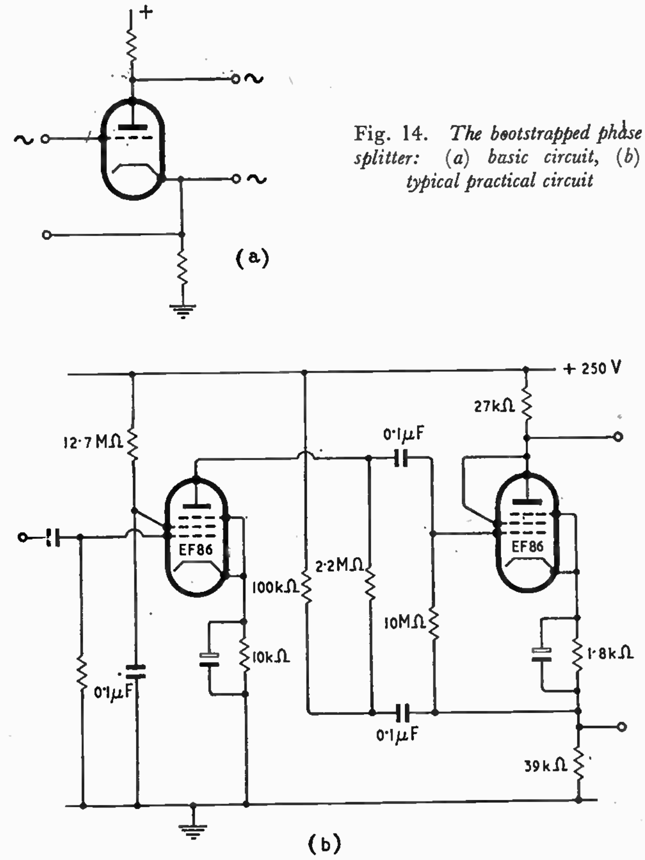


Fig. 14. *The bootstrapped phase splitter: (a) basic circuit, (b) typical practical circuit*

output impedance. The remedy would appear to be to interpose a second cathode-follower having a cathode load considerably higher than that of the first as shown in Fig. 13 (a). The latter would set an upper limit on the effective source impedance presented to the first-mentioned stage and could be chosen to be sufficiently less than the input impedance of this stage to ensure that it always operates substantially as a cathode-follower even when the actual source impedance is infinite. The additional stage would, however, work as a bootstrap with a high source impedance, and would function as a cathode-follower only when the source impedance is low. In a further development of this arrangement the cathode of the output valve is connected to the anode of the input valve, so that the effective anode impedance of the latter provides the cathode load of the former, and the two valves have series feed. This is a double application of the bootstrap principle, since the input valve multiplies the load value of its cathode impedance from the point of view of the output valve; i.e., it constitutes a current-operated bootstrap load, as in Fig. 8 (b). This arrangement has been called[18] the cascade cathode-follower.

4. BOOTSTRAPPING IN PUSH-PULL AMPLIFIERS

Bootstrapped Phase-Splitter

When the load of the common-cathode amplifier is divided into two parts and the earth connection shifted to their junction a reduced output is obtained at the anode; the remainder of the output appears at the cathode and is in phase with the input, as indicated in Fig. 14 (a). This case is intermediate between the two extremes of normal (zero cathode output) and bootstrap (zero anode output) operation. With equal subdivision of the load the two antiphase outputs are of equal amplitude; this special case is useful as a phase-splitter for push-pull stages, and has the advantage over the more familiar form, which may be derived from it by returning the lower side of the input to the other end of the cathode load, of giving a gain up to the limit of $\mu/2$, instead of unity, over each channel. A typical practical circuit having single-ended current input[19, 20] is given in Fig. 14 (b).

Combined Normal Amplifier—Bootstrap Amplifier Output Stages

The bootstrap phase-splitter may be regarded as a combination of a normal amplifier with a voltage-operated bootstrap amplifier in which the valve is shared. Another composite arrangement occurs in the series-connected type of push-pull amplifier or output

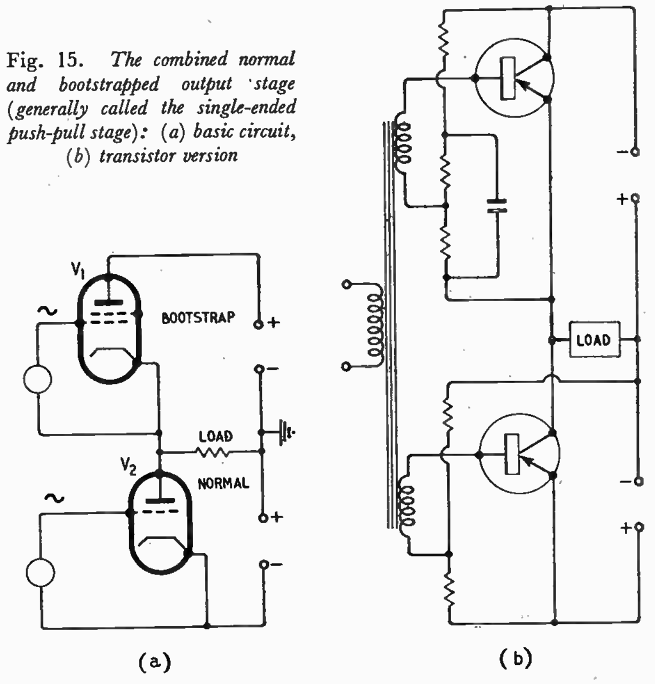


Fig. 15. *The combined normal and bootstrapped output stage (generally called the single-ended push-pull stage): (a) basic circuit, (b) transistor version*

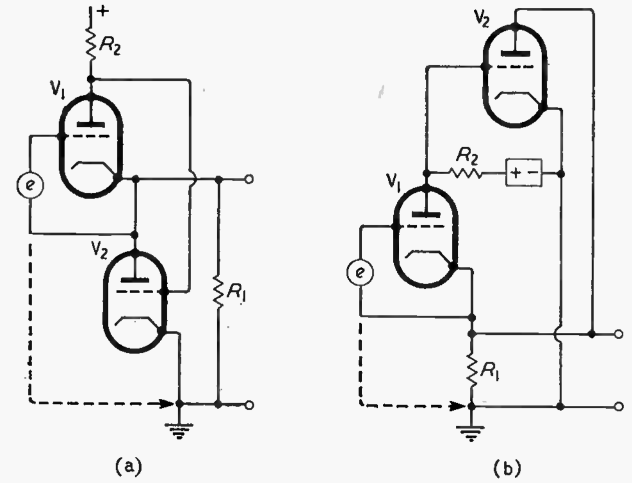


Fig. 16. *The series push-pull stage with internal phase-splitting connection: (a) as normally drawn, (b) redrawn to show current-bootstrapping of* V_2

stage, the basic circuit of which is shown in Fig. 15 (a)[21–26]. In this two valves, one (V_1) bootstrapped, the other (V_2) normally connected, operate in push-pull into a common load. It will be noted that the two inputs have no terminal in common, so that, with transformer input, two separate secondary windings are necessary; despite this inconvenience this circuit has been used with transistors[26], operating in the connection (viz., common emitter) equivalent to common cathode [Fig. 15 (b)]. In this, as in the preceding circuit, V_1 may, of course, be current driven from a preceding stage, as explained in an earlier section.

The need for dual input to the common load circuit of Fig. 15 may be avoided by inserting a load in V_1 anode lead of magnitude equal to that of the common load (including the shunting effect of V_2, where this is appreciable), and feeding the inverted signal so obtained to the grid of the lower valve V_2 as shown in Fig. 16 (a). The upper valve then acts as a phase-splitter of the type given in Fig. 14 (a), and the entire circuit is a combination of this circuit with the common-load one. When the circuit is redrawn as shown in Fig. 16 (b) it becomes apparent that if the earth connection is shifted to the other end of R_1, V_1 will revert to a normal amplifier, while V_2 becomes a current-operated bootstrap, at least for the usual case of $R_1 \ll r_{a1}$. As previously noted, when the lower side of the input to V_1 is connected to the bottom of the load, thereby incurring 100% feedback, the circuit becomes the series-connected (or 'stacked') cathode-follower pair due to E. L. C. White[21].

The method of driving V_2 shown in Fig. 16 is less satisfactory in power-output stages than the arrangements shown in Fig. 17, where the dual drive is obtained from a phase-splitter, the anode load of which is current-bootstrapped on the upper output valve. Either the degenerative or the bootstrapped forms of phase-splitter may be employed, the latter being preferred in Fig. 17 because of its increased gain. The figure also shows two methods of applying feedback, for gain stabilization, distortion reduction, or other purposes; series-shunt feedback is shown at (a), shunt-shunt at (b).

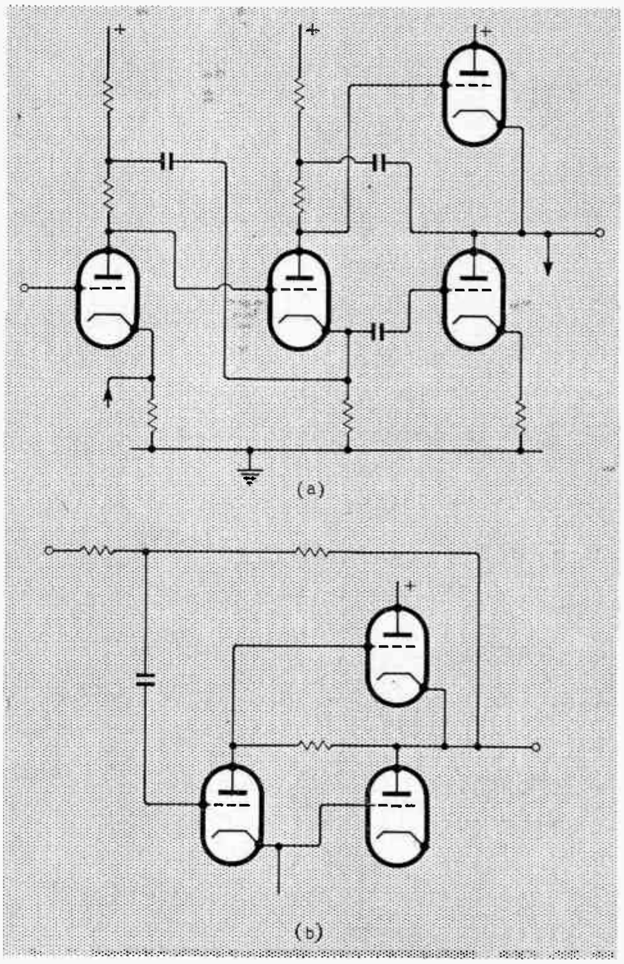


Fig. 17. *Single-ended push-pull output stage with drive circuit and feedback: (a) series-shunt feedback, (b) shunt-shunt feedback*

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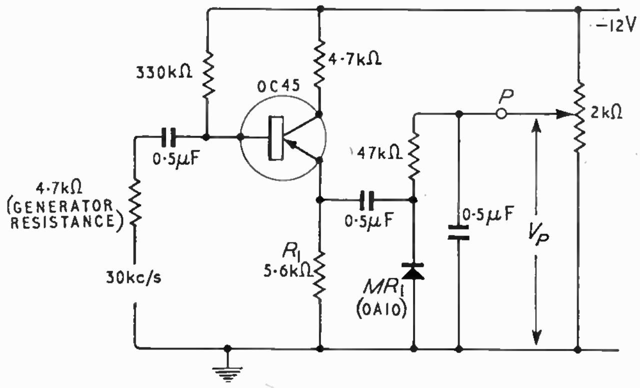
Correspondence

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

Transistor Amplifier Gain Control

Sir,—The gain of a common-emitter amplifier is frequently controlled by varying the d.c. emitter current I_e. This variation of I_e may be achieved in a number of ways.[1, 2] The gain reduction is mainly due to the increase in the emitter resistance r_e, and hence the increase in negative feedback, with decreasing I_e ($r_e \propto 1/I_e$). One serious disadvantage of the above method is that the change in d.c. operating conditions may lead to excessive distortion or even overloading in a.g.c. applications of the form described in references 1 and 2.

A method of achieving gain control by altering the emitter-to-earth resistance without affecting I_e is shown in the accompanying diagram, which is the subject of British Patent Application 12577/57. Here the effective a.c. emitter resistance consists of r_e in series with the parallel combination formed by R_1 and the resistance of a forward-biased junction diode MR_1. The a.c. resistance of the latter varies inversely as the direct current through it (this behaviour is analogous to the $r_e \propto 1/I_e$ relation mentioned previously). The setting of the slider P controls the d.c. current through MR_1 thereby controlling the gain. In the circuit shown the transducer gain at 30 kc/s referred to the 4·7-kΩ collector load decreased by 26 dB when V_p was reduced from -12 V to zero. In a practical application a.g.c. action could be provided by returning P to the collector of a common-emitter detector[3].



British Telecommunications Research Ltd.,
Taplow, Bucks.
16th July 1958.

L. J. Herbst.

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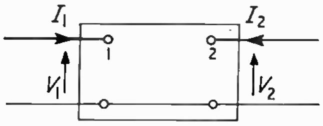
Parallel Four-Terminal Networks

Sir,—It would appear that Mr. Rogers has gone to unnecessary length in his article to derive a relationship which follows immediately from the admittance equations of a four-terminal network. With the usual notation these equations are:



$$\begin{aligned} I_1 &= Y_{11} V_1 + Y_{12} V_2 \\ I_2 &= Y_{21} V_1 + Y_{22} V_2 \end{aligned} \quad \ldots \quad (1)$$

If n such networks are in parallel the total currents I_{1n} and I_{2n} are:

$$\begin{aligned} I_{1n} &= V_1 \Sigma Y_{11} + V_2 \Sigma Y_{12} \\ I_{2n} &= V_1 \Sigma Y_{21} + V_2 \Sigma Y_{22} \end{aligned} \quad \ldots \quad (2)$$

Now if the load admittance is included in ΣY_{22} then I_2 is zero and

$$\frac{V_2}{V_1} = -\frac{\Sigma Y_{21}}{\Sigma Y_{22}} \quad \ldots \quad (3)$$

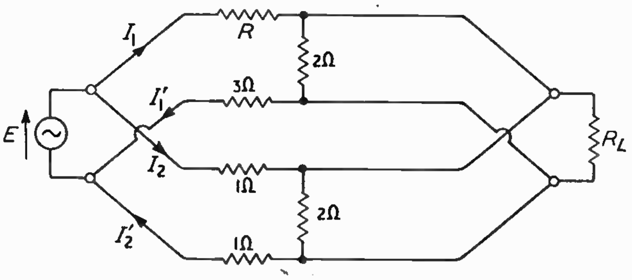
which is the relationship quoted in the article. As equality of Y_{12} and Y_{21} is not required this relationship can be applied to active networks.

It should be noted that this result is derived for three-terminal networks, as shown in the diagram, and in fact all Mr. Rogers' illustrations are three-terminal networks. Although the transfer characteristics of four-terminal networks may be represented by equations (1), the act of paralleling such networks constrains each of three independent nodal voltages and this information is not contained in the equations.

Norwood Technical College
West Norwood, London, S.E.27.
10th July 1958.

W. J. J. Raven.

Sir,—In his letter in your August issue, Mr. Rogers states that the theorem he discusses is valid only for "networks free from structural interaction". I am not sure what Mr. Rogers means by this phrase, but it can hardly apply to the circuit shown below, for which the theorem is invalid if $R \neq 3\ \Omega$ and is valid if $R = 3\ \Omega$.



It may be noted that, as pointed out by Guillemin, the theorem is valid if $I_1 = I'_1$ and $I_2 = I'_2$, with obvious extensions to cases where more than two networks are in parallel. It will thus be true for three-terminal networks and for balanced four-terminal networks, but it may not be true for unbalanced four-terminal systems.

I certainly agree with Mr. Rogers that there is a need for articles which bring to the notice of electrical engineers published works which they might find useful, and if Mr. Rogers had given more detailed references it would have been easier to deduce that such was his purpose. However, I would suggest that in such articles it is of prime importance to make clear the limitations of the work discussed, in order to reduce the risk of the results or theorems being used in an erroneous manner.

Electrical Engineering Dept.,
The University of Birmingham.
6th August 1958.

J. T. Allanson.

The Roentgen, the Rep and the Rem

Sir,—Congratulations on the article "The Roentgen, the Rep and the Rem" by Quantum in your July issue, explaining to the wider realms of the electronic industry some of the terminology of the nucleonics fringe.

Unfortunately, having defined the roentgen and explaining that one roentgen of electromagnetic radiation liberated slightly different amounts of energy per unit mass in different absorbing media, the article did not continue to state that in recent years it has become customary to express radiation dosage in another unit—the rad—corresponding to an energy absorption of 100 ergs per gm of absorbing material.

Tolerance levels for particular types of radiation are quoted, both in industry and in the Ministry of Labour's proposed Factories Regulations, in rads per hour (or week). While a knowledge of the physical conception of the various units is an obvious asset, the rad is the one unit with which the non-specialist should be familiar.

Lest this subject be considered beyond the fringe of electronics, let me cite the sad consequence of ignorance in the case of the radio engineer who, when presented with a faulty ion chamber dose-rate meter—which, being a box containing valves, was assumed to lie within his orbit—expressed judgment that the value of the grid leak in the input stage was much too high!

Plessey Nucleonics Ltd., G. D. Smith
Northampton.
13th August 1958.

New Books

Network Synthesis, Vol. 1

By David F. Tuttle, Jr. Pp. 1175. Chapman & Hall, 37 Essex Street, London, W.C.2 for John Wiley and Sons. Price £9 8s.

This book is the first part of a treatise on network synthesis that promises to be very comprehensive indeed, for its 1,175 pages are devoted to the steady-state synthesis of two-terminal networks alone. The author's intention to devote his second volume to four-terminal networks is probably wise, for such networks are compounded of two-terminal ones, and the mathematical foundations laid in this first volume should serve most of the requirements of the second.

The outstanding features of this book are its thoroughness and lucidity. The author has recognized that mathematical relationships do not in themselves afford adequate explanation, and appears to share this reviewer's opinion: "Mathematics tells one what it's equal to, but not what it is". Mathematical details are accordingly intermingled with a considerable amount of prose, which greatly facilitates understanding of the subject.

The book comprises fifteen chapters. The first two are of an introductory kind, and recapitulate most of the well-known principles of network analysis. The term "immittance" is introduced as a useful generality embracing the alternatives of impedance and admittance, and an adequate explanation is given of the procedure for normalizing values as an aid to calculation.

The complex-frequency variable is introduced in Chapter 3 in a simple way, and the necessary theory is very well developed over 83 pages. The inclusion of this relevant mathematical background is a commendable feature.

The fourth chapter is devoted to realizability requirements, from energy considerations in a passive network. The author refers, in his chapter-heading, to "—— some remarkable conclusions"; but one cannot help feeling that the treatment is unnecessarily long and complex in relation to the conclusions reached: it does not seem very remarkable for example, that the power supplied instantaneously to a network is the sum of the powers instantaneously stored and dissipated; nor even that the driving-point immittance of a realizable passive network must have a positive real-part.

The properties of driving-point immittance functions, commenced in the fifth chapter, are followed by the synthesis of non-dissipative Foster and Cauer type networks and then *RL* and *LC* networks. The general properties are taken up again in Chapter 8, with particular reference to the minimum reactance/susceptance and resistance/conductance conditions, the important reactance and resistance integral theorems, and computational methods based on the relationships between an immittance and its components, such as those due to Brune, Gewertz and Bode. This chapter is particularly clear and has a practical appeal.

RLC networks are covered in great detail in Chapters 9 and 10, with particular emphasis on Brune's approach and its extensions, and methods of realization not involving mutual inductance.

The theme is then interrupted by a short chapter illustrating some practical applications, followed by one concerned with the impurities in practical network elements. These are important for the realism that they impart, though the author is restricted in his illustrations by the exclusion of four-terminal networks from the preceding sections of the book.

The topic of approximation is explained clearly and in detail, in relation to the practical problem: "The approximation problem we have to solve is to determine a function $Z(p)$ or $Y(p)$, that is rational and positive real, and whose behaviour as a function of frequency is acceptably close to the requirements laid down by the 'customer'" (p. 741).

The Potential Analogy in Chapter 14 embraces 219 pages and, in conjunction with the pole-zero theory developed in the earlier chapters, it possibly represents the most integrated treatment in existence. An extensive knowledge of electrostatics is not assumed and, after the basis of the analogy is demonstrated by reference to the Cauchy-Riemann and Laplace equations, the analogy is developed smoothly to an advanced level by expansion of two-dimensional electrostatic theory from the simple case of a line of charge.

The final chapter, devoted to some thorough numerical illustrations, is followed by two appendices on computation. These include extensive tables for evaluation of resistance and reactance integral functions.

The many numerical illustrations in the text give expression to the author's wise contention on p. 342: "Numerical examples are essential in explaining any synthesis process". But the problems terminating each chapter, though well chosen and sometimes accompanied by useful suggestions, are unaccompanied by answers.

Whereas the adequacy of the explanatory prose has been commended, the author's enthusiasm has often led to the use of superfluous words and phrases. For example, on p. 30, "$j\omega$ being replaced by p, *if you will*"; on p. 70, referring to some equations, "—— *for the time being we merely place them in our tool kit with our other shiny new tools*": and on p. 87, "*We have now finished our excursion into the wonderland of function theory*". A more concise style might have saved a substantial number of pages.

The attractiveness of the publishers' production facilitates and encourages reading. The author is to be congratulated on his compilation of a book, suitable both for systematic study and for reference, which represents a marked advance in exposition of this difficult subject.

F.E.R.

The Ionosphere: Its Significance for Geophysics and Radio Communications

By Karl Rawer. Translated from the German by Ludwig Katz. Pp. 202. 72 Figs. Crosby Lockwood & Son Ltd., 26 Old Brompton Road, London, S.W.7. Price 42s.

Some research physicists and engineers interested in the propagation of radio waves by way of the ionosphere may already be acquainted with the German edition of "*Die Ionosphäre*" by K. Rawer, published in 1952. The publication of an English translation of the book will be welcomed by these and the large number of others who are now concerned with the ionosphere and the part it plays in long-distance radio communications. Furthermore, many individuals who are engaged in making ionospheric observations of all kinds as part of the programme of the International Geophysical Year will wish to continue their study and analysis of the scientific results obtained during this period, which ends at 31st December next, and they will find the volume under review of great assistance in this connection.

In the first half of the book the author, who was formerly with the French naval ionospheric prediction service, deals with the physics of the ionosphere and its characteristics as determined by radio-wave soundings from a number of observatories in different parts of the world. The information obtained in this way is supplemented by a knowledge of aurorae, geomagnetism and the general influence of the sun's radiation on the upper atmosphere.

A chapter in the second half describes in some detail the regular features of the ionosphere and the manner in which these depend on frequency, time, season and location. The less regular changes associated with a solar eclipse, or with the co-related phenomena of aurorae and disturbances of the earth's magnetic field, are outlined and illustrated. The final chapter deals with the propagation of radio waves around the earth by successive reflections between the earth's surface and the ionosphere, and the associated problem of forecasting the most suitable frequency to use for long-distance radio communication.

The book is well illustrated with some photographs of ionospheric sounding records and with many diagrams, a large proportion of which are based on experimental observations. The text is of a suitable length for those—whether engineers or scientists—who require a straightforward introduction to the ionosphere; and a bibliography of references to original papers is provided for those who wish to study the subject in more detail. R.L.S.R.

An Introduction to the Theory of Random Signals and Noise

By W. B. Davenport, Jr. and W. L. Root. Pp. 393 & ix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 77s. 6d.

On first receiving this book and glancing haphazardly through the pages I was both pleased and impressed by the inclusion of so many topics in which I was interested, and on which I knew I required further enlightenment. However, on settling down to read through the book more thoroughly and as I progressed slowly from chapter to chapter I was disappointed—not that I had been mistaken in the subjects nor were they scantily treated but I felt an opportunity had been missed.

During the past ten to fifteen years a number of excellent original papers have been published on the problem of random noise and the detection of signals masked by such noise. Unfortunately—perhaps a reflection on an engineer's training—most of these works have caused engineers to burn a lot of midnight oil in order to understand the mathematics involved. In fact this situation has resulted in many cases of quotation of results from these papers without a proper understanding of all the assumptions. This perhaps is particularly true of quotations from Rice's epic work on noise and Shannon's "Mathematical Theory of Communication". What I feel has been missing (except for one or two notable exceptions) is an approach which will encourage the engineer to explore this wide field and make use of the powerful tools it provides.

Most of the material covered in the early chapters of this book (i.e., the fundamentals of probability theory) is already widely published and there are many standard textbooks available. In rewriting such material an opportunity arose to present the information in engineers' language rather than a mathematician's. Davenport and Root have travelled some way along this road, but in my opinion not far enough. Woodward's book "Probability and Information Theory with Applications to Radar" is a good example of the approach I have in mind.

Davenport and Root have based their book on a series of lectures given to a 1st Year Graduate Course at M.I.T., and judging by the content the intake to this Course must have been of a fairly high standard. The authors have obviously worked very hard to collect and collate the scattered material into a logical sequence suitable for presentation in such a course, and there are numerous examples of some very careful thought on the part of the authors—in particular the chapter on 'averages'.

As an example of the use of the methods described in earlier chapters, and also to introduce the reader to further ideas, the problem of shot noise in thermionic valves is given a fairly thorough treatment in chapter 7.

Perhaps the three most useful chapters are 11–14 inclusive, in which the authors deal with optimum linear systems utilizing Weiner's theory, non-linear devices, and the statistical detection of signals. The earlier chapters may be said to provide the tools with which to tackle these problems; e.g., the determination of the probability distributions when the variable is transformed and the relations between correlation functions and frequency spectra. Some knowledge of matrices and vector notation is required, and fairly extensive use is made of transform theory of both Fourier and Laplace. Two appendices deal with impulse functions and integral equations of which considerable use is made.

The final chapter on statistical detection of signals is perhaps one of the most important in the book since it suggests a standard by which we may measure the performance of a system. In many papers this subject is rather glossed over and yet the particular decision criterion used can have a substantial effect on the estimated performance.

A short bibliography lists most of the important contributions to the subject and is, unlike so many bibliographies, international in character. It also refers to Chessin's well-known bibliography on noise published by the I.R.E. which in itself is very comprehensive.

This book which is the first of a group of books to be published by the Lincoln Laboratory, M.I.T., should serve as a very useful reference for engineers in this field, but will present a lot of hard work to anyone trying to start from scratch. There are a number of problems at the end of each chapter which provide useful exercises for the industrious.

J.W.R.G.

Chambers's Technical Dictionary (3rd Edition)

Edited by C. F. Tweney and L. E. C. Hughes, A.C.G.I., D.I.C., B.Sc.(Eng.), Ph.D., M.I.E.E., F.R.S.A. Pp. 1028. W. & R. Chambers Ltd., 11 Thistle Street, Edinburgh 2. Price 35s.

The first edition appeared in 1940. This is the third revised edition and includes a supplement. It aims at giving "definitions of terms that are of importance in pure and applied science, in all branches of engineering and construction, and in the larger manufacturing industries and skilled trades".

The field covered is enormous and the book is the collated work of numerous specialist contributors. The definitions are short and clear. No single reviewer can hope to assess their accuracy as a whole; he can merely regard those in his own specialized field with a critical eye.

The definitions in the electronics and radio field are, in the main, adequate, but there are a few exceptions. For example, negative feedback is stated to be "Interconnexion of the input and output terminals of an amplifier in such a manner that the output opposes the input, resulting in a reduction in amplification, but also in a *corresponding increase in output power obtainable with a given degree of harmonic distortion*". This should read "corresponding decrease of harmonic distortion for a given output power". Feedback does many useful things, but it does not work miracles.

The multivibrator is said to be "characterised by an irregular wave-form of oscillation". One knows that what is meant is a waveform in which the current or voltage does not vary smoothly with time, but those who do not know may well think that it varies erratically and unpredictably, so that successive cycles are unlike.

The definition of a rejector circuit is marred by the concluding words "when placed in series with the antenna circuit of a receiver", for the usage of rejectors is by no means confined to the aerial circuit. In fact, that is where they are now least used. The use of "antenna" instead of "aerial" is irritating and the two words themselves are not properly defined. "Antenna" is stated to be "An elevated and/or extended system of conductors used for the transmission and/or reception of electromagnetic waves."

When we turn to "aerial", however, we find: "Any exposed wire capable of radiating or receiving the energy to or from an electromagnetic wave. The term is preferably restricted to such, and should not be applied to aerial systems which are designed to have special characteristics, to which the term antenna is applicable".

The author may think this is a desirable usage of words but it is not in accord with present or past practice. The two words are synonymous, antenna merely being the American for aerial.

One cannot expect that a work of this magnitude can be free from all error and the pedantic can always find fault with definitions. The foregoing criticism of a few matters must not be taken to imply too much, therefore. There is no doubt at all that the book is an extremely useful one and the reviewer, for one, will certainly keep it handy. W.T.C.

Conductance Curve Manual

By Keats A. Pullen, Jr., Eng.D. Pp. 114. John F. Rider, Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $4·25.

The major part of this book is taken up by characteristic curves for 71 American receiving-type valves. In the case of triodes, the anode-voltage–anode-current curves are given; and for pentodes and tetrodes, screen-voltage–anode-current curves. Anode dissipation contours are drawn on these, also two other sets of contours, one showing anode conductance and the other mutual conductance. These latter are claimed by the author to "help in the design of circuits which, when actually built, conform closely to the predictions of the calculated design". In fact, the additional information is implicit in the current–voltage curves, and the author states that in the case of triodes his additional curves have been obtained from standard anode characteristic curves. It is certainly convenient to be able to select an operating point directly in terms of g_m and $1/r_a$, but this is not a prerequisite of good design. The screen-voltage–anode-current curves are in themselves useful, because valve manufacturers seldom publish these. The curves are plotted for screen voltage equal to half the anode voltage, but they are accompanied by diagrams which enable the user to estimate electrode current and mutual conductances for other ratios.

G.W.S.

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Pp. 254. Published by the Polish Academy of Sciences, Panstwowe Wydawnictwo Naukowe, Warsaw, Poland. Price Zl. 41.

Contains summaries of forty-six papers and communications on the theory and applications of ultrasonics. Most of these are printed in English, but three are in French.

Nachrichtentechnische Fachberichte, Band 10: Fernwirktechnik II

Edited by Dipl.-Ing. JOHANNES WOSNIK. Pp. 85. Friedr. Vieweg & Sohn, (20b) Braunschweig, Burgplatz 1, Germany. Price DM 14.

Collection of 17 articles, mostly on heavy-current subjects.

Guide to Broadcasting Stations 1958-59

Pp. 80. Includes long and medium-wave European stations, v.h.f. sound transmitters in the U.K., short-wave stations of the world, and U.K. television stations. Price 2s. 6d.

Radio Valve Data. 6th Edition

Pp. 136. Tabulated data on 3,000 valves, transistors, rectifiers, cathode-ray tubes, etc. Price 5s.

Both the above are published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1.

COLOUR TELEVISION COURSE

The Ministry of Education has given permission to the Northern Polytechnic, Holloway, London, N.7, to hold a full evening course on the principles and practice of colour television.

The course is intended to cover the fundamentals of television transmission and reception in colour. The entry qualification required is at least City and Guilds Radio III or City and Guilds Final Certificate, and some knowledge of monochrome principles will be necessary.

The syllabus covers the eye and perception of colour, colour synthesis and photometry. A résumé of monochrome technique is made, and the type of camera for colour is considered. Possible transmission systems are compared, and the N.T.S.C. system considered in detail. Receiver design forms a large part of the work, and display tubes for projection or direct viewing are considered. Test equipment and measurements are a vital section of the syllabus, and closed-circuit systems are surveyed.

Polytechnic staff will be dealing with the bulk of the course, but specialist lecturers will be engaged for items of particular interest.

Provision has been made for practical work on colour receivers and, in the event of experimental transmissions taking place during the run of the course, full facilities exist for their display.

MEETINGS

The Television Society

19th September. "Experimental Colour Receiver: Setting up and Adjustment", by E. Ribchester, B.Sc.

2nd October. "Printed Circuit Techniques applied to a Television Tuner", by P. C. Ganderton.

These meetings will be held at 7 o'clock at the Cinematograph Exhibitors' Association, 164 *Shaftesbury Avenue, London, W.C.2.*

The Society of Instrument Technology

30th September. "Frequency Response Analysis using Describing Function Method", by Professor J. C. West, D.Sc., Ph.D., to be held at Manson House, Portland Place, London, W.1, at 6 o'clock.

Radar and Electronics Association

23rd September. "The Three Esses—Stereophonic Sound Systems", by J. Moir, to be held at the Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Resumption of Day-Time Transmission by MSF 60 kc/s

The extensive work of aerial maintenance at Rugby is now almost complete and from the 15th November it is planned that MSF 60 kc/s will revert to its customary transmitting period at 14.29–15.30 G.M.T. each day. By the same date the high-power telegraphy transmitter, GBR 16 kc/s, will also be in operation after an absence of nearly two years. It will be recalled that the carrier frequency of this station is derived from the crystal oscillator controlling the MSF signals and the published values of frequency deviation therefore relate to both MSF and GBR.

Deviations from nominal frequency for July* 1958

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

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

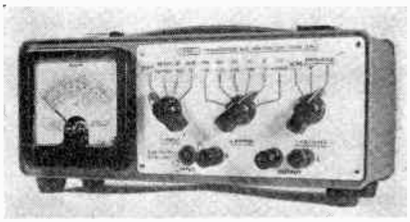
New Products

Transistor A.C. Voltmeter Type TM2

This instrument may be used as an a.c. voltmeter or a general-purpose a.c. amplifier. Voltages from $75\,\mu$V to 500 V may be measured in 12 ranges. The frequency response is within ±3 dB from 6 c/s to 250 kc/s and is the same in either function.

The instrument contains a high-gain transistor amplifier with negative feedback. When it is used as a voltmeter the output of the amplifier drives a full-wave rectifier and a $3\frac{1}{2}$-in. moving-coil meter. Negative feedback is then derived from the current in the rectifier circuit so that linear scales are ensured. An output of 0·5 V at f.s.d. is available to permit oscilloscope observations to be made while signals are measured.

Alternative output facilities are available when the instrument is used as an amplifier. Voltage feedback can be applied to give a low output impedance with a maximum gain of 60 dB; the instrument is then suitable for use as a pre-amplifier for an oscilloscope. Alternatively, current feedback is applied to give a high output impedance, which is suitable for pen recorders, the output being 350 mA per volt input with a maximum output of 1 mA peak.



The instrument operates from self-contained batteries with a service life longer than 1,000 hours. The absence of a mains connection permits the use of the instrument in floating circuits.

Levell Electronics,
High Street, Edgware, Middx.

Add and Subtract Counter

This counter is a self-contained unit with a five-figure presentation. The number wheels are actuated by two escapement mechanisms operated from separate d.c. or a.c. supplies.

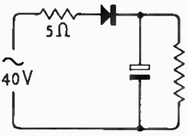
The first escapement operates to add, at the same time releasing the second and, when subtracting, this operation is reversed.

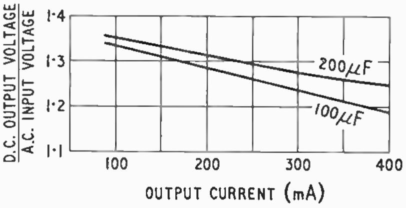
Counting is possible up to speeds of 17 per second and the design is such that damage cannot occur if both adding and subtracting pulses are supplied at the same time.

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

Silicon Rectifier

Silicon junction rectifier type Si 3 is described as a hermetically-sealed wire-ended component of very small dimensions and low weight, designed for use in all types of electronic equipment operated at mains voltages up to 480 V r.m.s. for a resistive load up to 0·5 A and at mains voltages up to 240 V r.m.s. for a capacitive load up to 0·4 A. Characteristics: Maximum peak inverse voltage = 750 V, maximum reservoir capacitor = $200\,\mu$F, minimum source impedance with capacitive load = $5\,\Omega$, maximum ambient temperature = 50° C, average reverse current at 750 V peak inverse voltage and nominal operating temperature = 1 mA, maximum forward voltage drop = 1·1 V, maximum case temperature = 110° C.





The diagram shows the rectifier's operating characteristics with a capacitive load.

The rectifier is available in the U.K. from *R. H. Cole (Overseas) Ltd.,*
2 Caxton Street, Westminster, London, S.W.1.

Low-Voltage Pencil Soldering Iron

The salient features of this soldering iron are as follows: Weight $1\frac{1}{2}$ oz.; overall length 8 in.; $\frac{5}{16}$-in. diameter barrel with no screw heads or other projections, so that the most inaccessible joints can be made without damaging neighbouring components; standard voltage is 6 or 12, but elements can be wound specially up to 50 V; heating-up time, approximately 2 minutes; power consumption 18 W.

Browning's Electric Co. Ltd.,
Boleyn Castle, Green Street, Upton Park,
London, E.13.

Miniature Ferrite-Cored Coils

A range of coils for transistor receivers is now available. It consists of a 4-in. ferrite-rod aerial with a $400\text{-}\mu$H winding ($Q = 125$ at 1 Mc/s), a pot-cored oscillator coil ($230\,\mu$H, $Q = 90$) and 470-kc/s i.f. coils also on pot cores ($450\,\mu$H, $Q = 110$ or 130). The coils are said to be suitable for use with Mullard transistors OC44 and OC45, and the coil manufacturers can supply a circuit diagram for a 6-transitor medium-wave receiver. The potted coils are in aluminium screening cans $\frac{1}{2}$ in. square and $\frac{11}{16}$ in. high.

Denco (Clacton) Ltd.,
357-9 Old Road, Clacton-on-Sea, Essex.

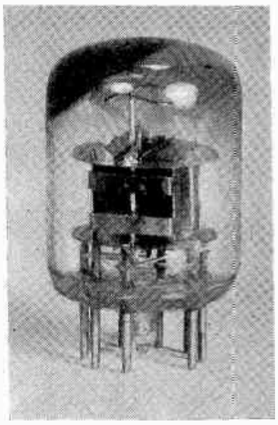
General-Purpose Audio Transistor

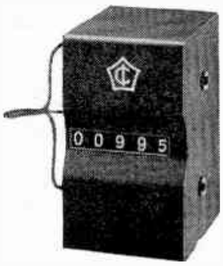
A new transistor, type XB.104, is intended for general-purpose applications in the audio-frequency range. Some of its operating characteristics are as follows: current amplification, common emitter $\beta = 30$; noise factor, common emitter = 6 dB; maximum collector-to-emitter leakage current (base open-circuit, $V_{ce} = -10$ V) = $250\,\mu$A.

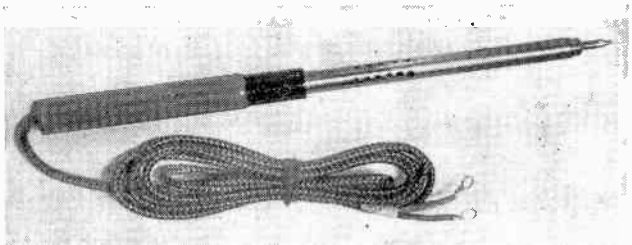
Siemens Edison Swan Ltd.,
155 Charing Cross Road, London, W.C.2.

Valve for Wide-Band Amplifiers

The A2674 is a new high figure-of-merit tetrode for use in wide-band amplifiers. It is described as a direct equivalent of the Western Electric 436A. The slope-to-capacitance ratio is particularly high, the slope being 32 mA/V at 23 mA anode







current, and the input capacitance 15 pF.

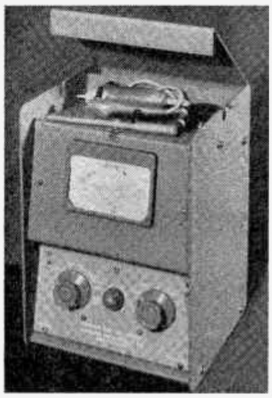
The valve and its special socket employ gold-plated pins and connections, ensuring freedom from contact noise. The socket has wide pin spacings and has been designed to enable tag-to-earth connections to follow the shortest possible path.

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

300-Mc/s Voltmeter

This valve voltmeter type TF1300 is described as a moderately-priced instrument for the laboratory or production engineer who requires a sturdy, reliable voltmeter of medium range, good accuracy and good stability.

The a.c. measurement range is 0·1 to 100 V, 20 c/s to 300 Mc/s. D.C. is measured up to 300 V and resistance up to 5 MΩ. The indicating meter is direct-reading on a.c., d.c., and ohms ranges, and no correction factors are necessary. Zero stability is claimed to be of a high order with respect to both time and mains variation, and only one zero setting is required for all a.c. or d.c. ranges. Both a.c. and d.c. inputs are isolated from chassis.

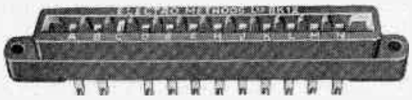


A.C. measurements are made with a lightweight cylindrical probe allowing direct contact with the circuit under test. The rectified output from the probe is applied to one arm of a valve bridge comprising two double-triodes; a second arm of the bridge is coupled to a diode current-balancing network including a potential divider across the h.t. supply and a 3-volt stabilizer element. The indicating meter is automatically protected from overload.

Marconi Instruments Ltd.,
St. Albans, Herts.

Printed Circuit Connectors

Series BK and BKM connectors have been produced to meet the demand for a connector which will accommodate a printed-circuit card based on a 0·1-in. grid. The BK series, with contact centres 0·2 in., are available with 8 and 12 contacts and the BKM series, with contact centres 0·1 in., are available with 17 and 25 contacts. These connectors, which are interchangeable, have end-fixing and polarizing pins which can be fitted in any position.

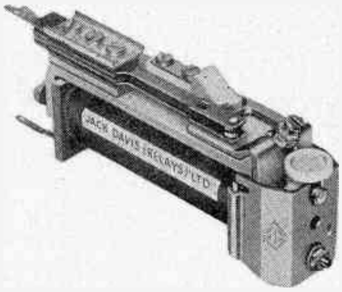


A transistor holder and connector, based on Series M plug and socket, which has been especially adapted to carry several transistors, in the series OC, OCP and 2N, has also been introduced.

Electro Methods Ltd.,
Caxton Way, Stevenage, Herts.

3000-Type Magnetic Relay Latch

This is a simple device which can be fitted to a 3000-type relay for latching the relay after energizing. Release is effected by push-button above the latch or, for remote releasing, by an extended cable.



Jack Davis (Relays) Ltd.,
Tudor Place, London, W.1.

Subminiature Coaxial Plugs and Sockets

The plugs of this range of subminiature plugs and sockets are securely retained by a knurled nut which screws on to a thread on the socket. A rubber washer is fitted to the junction of this thread with the fixing flange, to prevent loosening of the knurled nut by vibration. P.T.F.E. insulation is used and all contact surfaces are gold-plated.

The plugs are suitable for use with coaxial cable having outside diameter of $\frac{1}{16}$ in. approximately.

The range comprises items with the following specifications:

L.1403/FP Free Plug

The cable to this plug enters along the axis of the plug. The rubber cover protects the cable clamp from damage and gives the plug a streamlined appearance.

L.1403/RFP Right-Angle Entry Free Plug

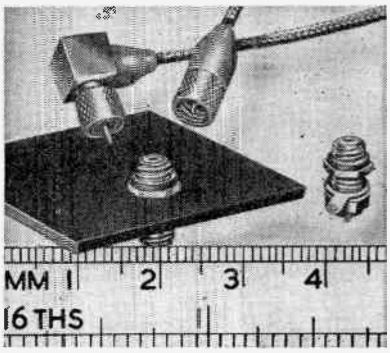
This plug has the cable entry at right angles to the axis of the plug enabling the cables to be 'dressed-down' against a panel without the danger of sharp bends causing cable failure.

L.1403/CS Chassis-Mounting Socket

This is designed for use where the feeder enters equipment and is not required to continue as a coaxial cable, as the components associated with it are in close proximity to the socket. It is fitted with gold-plated soldering tags.

L.1403/BS Bulkhead Adaptor

This socket-to-socket panel-mounting adaptor provides for the passage of a coaxial cable through a panel in such a way that the cable can be disconnected on either side. It is ideal for use where a coaxial lead needs to be continued behind the main panel of equipment.



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

Valve Voltmeter

This instrument is essentially a centre-zero reading device, but facility for a left-hand dial zero is provided for a.c. and ±d.c. measurements. The offsetting control is so arranged that no re-zeroing of the meter is necessary when going from balanced to plus or minus d.c. or a.c., and vice versa.

A silicon junction diode is incorporated for a.c. measurements, and the d.c. part of the circuit consists of a double-triode with the meter connected between the cathodes.

The zero stability is given as about 1% of full-scale deflection for any 12-hour period (after an initial warming-up period of 10 minutes).

The mains input voltage is 110 or 220 V a.c. 50–60 c/s and the instrument is stabilized against ±10% variations in mains voltage. The maximum meter deflection for this variation is 1% full scale.

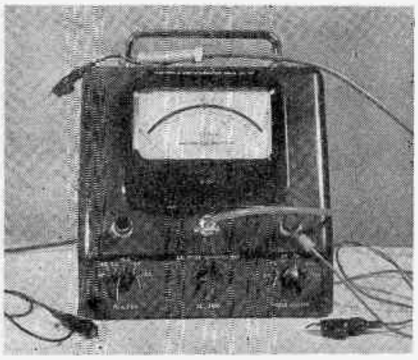
The frequency response of the probe is claimed to be within 1 dB between 50 c/s and 10 Mc/s.

The d.c. input resistance on all ranges is 30 MΩ (shunted by approximately 30 pF), and the a.c. input impedance is 5 MΩ shunted by 5 pF (max.). The capacitance of the negative terminal to earth is approximately 200 pF.

Ranges: In balanced d.c. position (accuracy 2%), ±2·5, ±7·5, ±25, ±75, ±250, ±750 V. In a.c. position (accuracy 3%), 5, 15, 50 and 150 V.

All voltages measured may be superimposed on ±450 V d.c.

Winston Electronics Ltd.,
Govett Avenue, Shepperton, Middx.



Abstracts and References

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

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

ACOUSTICS AND AUDIO FREQUENCIES

534.2-8-14 **2610**

Investigation of an Ultrasonic Field in a Liquid.—B. Labory & G. Laville. (*C. R. Acad. Sci., Paris*, 21st Oct. 1957, Vol. 245, No. 17, pp. 1401–1403.) Measurements have been made to determine the force exerted on an Al cylinder in the sound field of a quartz transducer operating at frequencies up to 7 Mc/s in water. Results are compared with theoretical calculations of the radiation field.

534.2-8-14 : 534.6 **2611**

Apparatus for Measuring the Attenuation of Ultrasonic Propagation in Liquids.—G. Laville. (*C. R. Acad. Sci., Paris*, 28th Oct. 1957, Vol. 245, No. 18, pp. 1523–1526.) Description of apparatus designed in accordance with the results of earlier measurements (2610 above), and a report on its application in determining the ultrasonic absorption of sugar solutions.

534.2-8-14 : 534.6 **2612**

Two-Crystal Interferometric Method for Measuring Ultrasonic Absorption Coefficients in Liquids.—R. S. Musa. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 215–219.) Two matched quartz crystals are used. A standard signal generator drives one of the crystals continuously. The amplitude of the resulting pressure waves is measured at the other crystal by means of a communications receiver and a voltmeter.

534.22-14 : 546.212 **2613**

On the Temperature Variation of the Velocity of Sound in Water.—C. Salceanu. (*C. R. Acad. Sci., Paris*, 21st Oct. 1957, Vol. 245, No. 17, pp. 1371–1374.) Results of measurements at a frequency of 1315 c/s at 27°–81°C show that the velocity has a maximum value at about 62°C. See also 1296 of May (Greenspan & Tschiegg).

534.78 **2614**

Dynamic Analogue Speech Synthesizer.—G. Rosen. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 201–209.) "A dynamically controllable electrical analogue of the vocal tract capable of synthesizing sequences of speech sounds is described. The acoustic transmission line between the glottis and lips in the human vocal tract is realized electrically by eleven electronically controlled variable *LC* sections plus three fixed sections."

534.78 : 621.391 **2615**

Message Procedures for Unfavourable Communications Conditions.—I. Pollack. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 196–201.) "Several message procedures, designed to improve speech communications under extremely unfavourable speech-to-noise ratios, were examined. A message procedure based upon the informational principle of successive selections among a reduced number of alternatives was strikingly superior to a message procedure based upon the repetition of a single selection among a larger number of alternatives."

534.79 **2616**

Loudness of Periodically Interrupted White Noise.—I. Pollack. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 181–185.)

534.846 **2617**

Studio for Listening Tests.—L. O. Dolansky. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 175–181.) The design, construction, acoustical testing and articulation testing of a studio giving reasonably constant listening conditions are described. The reverberation time was found to be about 0·5 s.

621.395.61 **2618**

Noise Shield for Microphones used in Noisy Locations.—M. E. Hawley. (*J. acoust. Soc. Amer.*, March 1958, Vol. 30, No. 3, pp. 188–190.) The rubber noise shield developed significantly improves the speech/noise ratio of microphones for military use.

AERIALS AND TRANSMISSION LINES

621.372.2 **2619**

Electromagnetic Wave Propagation in an Almost Periodic Waveguide.—A. A. Sharshanov & K. N. Stepanov. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1474–1481.) A mathematical analysis is presented of (*a*) wave propagation in cavity resonators, (*b*) propagation at a frequency near to the transmission threshold, and (*c*) the propagation of e.m. waves in a cylindrical waveguide loaded with dielectric disks. The solution of differential equations with slowly varying coefficients is considered.

621.372.2 **2620**

Transmission-Line Discontinuities.—K. W. H. Foulds. (*Electronic Radio Engr*, July

1958, Vol. 35, No. 7, pp. 263–267.) The conditions necessary for a transmission line with two identical discontinuities to be matched in the steady state to the generator are examined. The way in which the steady-state conditions are derived from the transient behaviour is explained.

621.372.2 **2621**

The Effects of Reflections from Randomly Spaced Discontinuities in Transmission Lines.—R. K. Moore. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 121–126. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

621.372.2 **2622**

The Probability of Specified Losses at Mismatched Junctions.—J. H. Craven. (*J. Brit. Instn Radio Engrs*, May 1958, Vol. 18, No. 5, pp. 293–296.) "Probability contours are presented for specified losses (<1 dB to 6 dB) at junctions between networks or lines for a range of voltage s.w.r. from 1 to 10."

621.372.2 **2623**
: [621.372.826 + 621.396.677.7

Single-Slab Arbitrary-Polarization Surface-Wave Structure.—R. C. Hansen. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 115–120. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.) See also 3733 of 1957 (Plummer & Hansen).

621.372.2 : 621.385.029.6 **2624**

Fast Waves in a Coaxial Helical Line.—L. N. Loshakov & E. B. Ol'derogge. (*Radiotekhnika, Mosk.*, June 1957, Vol. 12, No. 6, pp. 25–30.) Waves with a phase velocity exceeding that of waves in an infinite dielectric are investigated on the assumption that the inner conductor (helix) of the line can be replaced by an anisotropically conducting surface. Conditions are established under which the propagation of such waves is possible.

621.372.22 **2625**

The Transmission Characteristics of a Line consisting of Axially Stacked Insulated Metal Rings.—G. Piefke. (*Arch. elekt. Übertragung*, Oct. & Nov. 1957, Vol. 11, Nos. 10 & 11, pp. 423–428 & 449–454.) The mathematical treatment developed earlier (3035 of 1957) is applied to an investigation of the 'ring-element waveguide'. The axial thickness of the insulated rings is much less than the wavelength. The propagation constants for various modes are calculated.

621.372.821 **2626**

Coupled Strip Transmission Lines with Rectangular Inner Conductors.—J. D. Horgan. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 92–99. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

621.372.823 : 621.372.852.1 **2627**

The Effects of Mode Filters on the Transmission Characteristics of Circular Electric Waves in a Circular Waveguide.—W. D. Warters. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 657–677.) Measurements have been made showing the mode conversion effects for the TE_{01} mode in a circular waveguide at a frequency of 9 kMc/s using a pulse technique. The conversion to other modes has been measured and mode filters have been developed which are effective against all the spurious modes generated. The effect of filter spacing on pulse distortion and attenuation is shown and the spacing of the filters is discussed.

621.372.825 **2628**

Semicircular Ridges in Rectangular Waveguides.—J. Van Bladel & O. Van Rohr, Jr. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 103–106. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

621.372.829 : 621.372.852.1 **2629**

Research Models of Helix Waveguides.—C. F. P. Rose. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 679–688.) The construction of a waveguide for the TE_{01} circular mode is discussed. The waveguide has a helix on the inner wall which acts as a mode filter. The design and construction of such a waveguide for use at 55 kMc/s is described.

621.372.829 : 621.385.029.6 **2630**

Helical Waveguides—Closed, Open and Coaxial.—G. M. Clarke. (*J. Brit. Instn Radio Engrs*, June 1958, Vol. 18, No. 6, pp. 359–361.) "The application of helical waveguides to electron-beam amplifiers employing slow-wave structures is described. The performances of the various possible configurations are discussed." See 353 of February (Waldron).

621.372.852.1 **2631**

The Design of Inductive Post-Type Microwave Filters.—M. H. N. Potok. (*J. Brit. Instn Radio Engrs*, May 1958, Vol. 18, No. 5, pp. 263–272.) Using experimental and theoretical results it is shown that filters can be designed to have a desired voltage s.w.r. within the pass band and a given insertion loss outside it. Optimum design of three- and four-cavity filters is discussed.

621.372.855 **2632**

Theoretical Analysis of the Design of a High-Power Waveguide Load.—V. Hlubucek. (*Slab. Obz., Praha*, July 1957, Vol. 18, No. 7, pp. 420–425.) Conditions for the uniform absorption of power along the load with minimum voltage s.w.r. are derived. Practical design details are given.

621.396.67 : 538.221 **2633**

Magnetic-Field Antenna.—W. J. Polydoroff. (*Electronic Ind.*, March 1958, Vol. 17, No. 3, pp. 66–68.) The use of magnetic materials, such as ferrites, in the design of high-gain aerials is briefly discussed. The pick-up of such aerials is best considered in terms of the magnetic rather than the electric field.

621.396.676 **2634**

Flush-Mounted Aircraft Aerials.—M. Lorant. (*Wireless World*, July 1958, Vol. 64, No. 7, pp. 337–338.) The use of scale models in the study of the radiation patterns of the aerials is briefly outlined.

621.396.677.062.8 **2635**

Aerial Selectors.—L. Leng. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 446–450.) Indoor-type aerial switching installations for s.w. transmitters are described.

621.396.677.3 **2636**

The Directional Aerial Installation of the 'Deutsche Welle'.—K. Alt & H. König. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 427–433.) Description of the s.w. directional aerial system erected at Jülich, Germany.

621.396.677.3 : 621.396.712 **2637**

Long-Distance H.F. Broadcasting.—T. W. Bennington. (*Wireless World*, July 1958, Vol. 64, No. 7, pp. 331–336.) The characteristics of the transmitting aerial array for good broadcast coverage of a distant area are discussed.

621.396.677.71 **2638**

A Low-Frequency Annular-Slot Antenna.—J. R. Wait. (*J. Res. nat. Bur. Stand.*, Jan. 1958, Vol. 60, No. 1, RP2822, pp. 59–64.) "The radiation characteristics of an annular slot cut in an ideally conducting ground plane are discussed. The voltage impressed between the concentric edges is assumed to be constant around the slot. The annular slot is backed by a hemispherical cavity which has imperfectly conducting walls. For a specified voltage, the power radiated in the upper half-space and the power absorbed by the hemispherical cavity are calculated. It is indicated that the power absorbed can be reduced greatly by lining the walls of the cavity with a wire mesh. A flush-mounted antenna of this type at low frequencies may have certain practical advantages over the more conventional monopole." See also 1326 of May.

621.396.677.81 : 538.566 **2639**

The Impedance of a Wire Grid Parallel to a Dielectric Interface.—J. R. Wait. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 99–102. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.) See also 3273 of 1957.

621.396.677.833.1 **2640**

Far-Field Radiation of a Cheese Aerial.—R. F. Kyle. (*Electronic Radio Engr*, July 1958, Vol. 35, No. 7, pp. 260–262.) The phase and amplitude distributions across the aperture of an X-band cheese aerial were measured and the radiation pattern was calculated. Good agreement was obtained with the measured patterns.

621.396.679 **2641**

On the Calculations of Transverse Current Loss in Buried-Wire Ground Systems.—J. R. Wait. (*Appl. sci. Res.*, 1958, Vol. B7, No. 1, pp. 81–86.) An expression is derived for the impedance between a wire grid buried in a homogeneous ground and an overhead conducting plane. The power loss due to transverse currents in buried-wire systems for aerials is calculated.

AUTOMATIC COMPUTERS

681.142 **2642**

A Survey of Delay Lines for Digital Pattern Storage.—S. Morleigh. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 380–387.) A review of e.m. and ultrasonic delay lines for use over a wide range of delay times. Wire-type acoustic delay lines using magnetostrictive transducers, acoustic lines using liquids or solids as delay media, and both the continuous and lumped-parameter types of e.m. delay line are examined.

681.142 **2643**

Modern Trends in Analogue Computation.—B. B. Murphy. (*IVA, Stockholm*, 1957, Vol. 28, No. 8, pp. 336–350. In English.)

681.142 **2644**

A Special Analogue Computer for Calculating the Oscillations of Diaphragms.—H. Dehnert. (*Elektronik*, Oct. 1957, Vol. 6, No. 10, pp. 301–303.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.001.4 : 621.395.64 **2645**

Passive Components for Submarine Telephone Cable Repeaters.—M. C. Wooley. (*Trans. Inst. Radio Engrs*, Nov. 1957, No. PGRQC-12, pp. 14–24.) The design and inspection of inductors, resistors and transformers are briefly described. Details of the inspection of silvered mica capacitors are given as an example of the procedure adopted.

621.3.042.4 (083.57) **2646**

Nomogram for Air-Gap Design.—A. C. Sim. (*Electronic Radio Engr*, July 1958, Vol. 35, No. 7, pp. 250–251.) A nomogram is given for determining the relation between the actual air gap in a choke or transformer and its effective value for normal types of laminations.

621.316.825 **2647**

Thermistors. — K. R. Patrick. (*Electronic Radio Engr*, July 1958, Vol. 35, No. 7, pp. 242–249.) A review is given of the various types of thermistor, their method of manufacture and the physical theory of their operation. Some typical applications are outlined.

621.318.435 : 621.373.443 **2648**

Saturable Reactors fire Radar Magnetrons.—H. E. Thomas. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 72–75.) 'Magnetic modulator uses saturable reactors to convert input sine wave into narrow, high-peak-power output pulses. Basic action of current-pulse compression with magnetic modulators is explained. Polarizing and differentiating circuits, delay-line wave shaping, pulse permeability measurements, cancellation effects and related features leading to improved design are discussed."

621.318.57 : 621.387 **2649**

A Multichannel Dekatron Scaling Unit.—F. W. Lovick. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 394–395.) Equipment is described for counting simultaneously in up to six channels. A variety of counting rates can be obtained.

621.319.4 **2650**

The Duroplast Capacitor. — R. Bretschneider. (*Nachr Tech.*, Oct. 1957, Vol. 7, No. 10, pp. 460–465.) A tubular paper capacitor impregnated with epoxy resin which eliminates the need of a sealed metal can. Constructional and electrical data are given.

621.372 : 621.396.822 : 530.162 **2651**

Thermal Fluctuations in a Nonlinear System.—N. G. van Kampen. (*Phys. Rev.*, 15th April 1958, Vol. 110, No. 2, pp. 319–323.) An electric circuit containing a voltage-dependent resistance $R(V)$ and a capacitance is studied. A general method is given for finding the spectral density of current fluctuations, and explicit results are obtained for two special forms of $R(V)$.

621.372.4 **2652**

A Numerical Method for the Determination of a Two-Pole Function which Approximates a Given Complex Function in a Band of Real Frequencies.—R. Unbehauen. (*Arch. elekt. Übertragung*, Nov. 1957, Vol. 11, No. 11, pp. 440–448.)

621.372.413 : 621.318.134 **2653**

On the Theory of Anisotropic Obstacles in Cavities.—W. Hauser. (*Quart. J. Mech. appl. Math.*, Feb. 1958, Vol. 11, Part 1, pp. 112–118.) Variational expressions are derived for the resonance frequencies of a cavity containing a material with tensor electromagnetic properties.

621.372.413 : 621.372.54.029.6 **2654**

Circularly Polarized Microwave Cavity Filters.—C. E. Nelson. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 136–147. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.) See also 41 of 1957.

621.372.5 **2655**

Synthesis of Driving-Point and Transfer Functions by Continued-Fraction Expansion.—R. E. Vowels. (*Aust. J. appl. Sci.*, Sept. 1957, Vol. 8, No. 3, pp. 151–168.) "A method of synthesis of driving-point functions has been developed in which more than one network element may be included in each branch of the ladder network. The method is first applied to *RL* and *LC* two-element-type networks and then extended to the general *RLC* network. Four-terminal networks having prescribed driving-point and transfer impedances are synthesized in a similar manner."

621.372.54 **2656**

The Unbalanced Symmetrical Parallel-T Network taking account of Condenser Loss.—K. Posel. (*Trans. S. Afr. Inst. elect. Engrs*, July 1957, Vol. 48, Part 7, pp. 243–257. Discussion, pp. 257–260.) The network is analysed without adopting the usual approximations, and its application to wave-analyser and oscillator circuits is discussed. See 3425 of 1957.

621.372.54.029.64 **2657**

Synthesis of a Class of Microwave Filters.—H. Seidel. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 107–114. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

621.372.543.3 : 621.375.126.029.3 **2658**

Bifilar-T Trap.—A. Hendry & A. G. McIntosh. (*Electronic Radio Engr*, July 1958, Vol. 35, No. 7, pp. 254–259.) Details are given of a circuit with a bifilar-T trap designed for rejecting the second harmonic of a 1-kc/s signal in an infrared detector amplifier. Other applications of the bifilar-T trap are considered, e.g. a tuned a.f. amplifier having a bandwidth of a few hundred cycles with a narrow band of high attenuation at the centre, and a narrow-band feedback amplifier.

621.373.2.029.65 **2659**

Production of Millimetre Waves by a Spark Generator.—J. Hart. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, p. 743.) A brief account of experiments which suggest that the spark generator does not offer a practical means of extending the radio spectrum to wavelengths less than 5 mm.

621.373.43 **2660**

Circuits for Forming Pulses from a Sinusoidal Voltage using a Relatively Low Supply Voltage.—V. I. Zabavin. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 73–77.) Circuits are considered which have a number of advantages over multivibrator and trigger circuits. Valve, transistor or hybrid circuits can be used.

621.373.431.1 **2661**

The Schmitt Multivibrator.—G. L. Swaffield. (*Wireless World*, July 1958, Vol. 64, No. 7, pp. 344–348.) A detailed design procedure is given and applications of the circuit are outlined.

621.373.431.1 : 621.314.7 **2662**

A Method for Sharpening the Output Waveform of Junction-Transistor Multivibrator Circuits.—A. E. Jackets. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 371–374.) Extension of previous work (2340 of 1956). A method of designing the circuit to reduce the recharging time of the coupling capacitor is given. An improvement in waveform at the collector of one transistor is achieved with an accompanying deterioration of the other output.

621.373.44 : 621.317.34 **2663**

Design of a Dual Pulse Code Train Generator.—A. M. Leuck & T. I. Humphreys. (*Electronic Equipm.*, Nov. 1957, Vol. 5, No. 11, pp. 22–25.) Description of a pulse-train generator, developed for delay-line testing, which provides a double train of pulses of variable separation.

621.373.52 **2664**

Transistor Oscillators and their Load-Independent Properties.—W. Herzog. (*Nachrichtentech. Z.*, Nov. 1957, Vol. 10, No. 11, pp. 564–569.) The equivalence of the 'internal' and the external feedback in a transistor oscillator circuit is established. Two types of load-independent oscillator are considered, one with a load impedance in parallel with the equivalent passive quadripole, and the other with the impedance in series. The conditions for obtaining an oscillator with two load-independent impedances, and a load-independent oscillator in the form of a six-terminal network are examined. See also 2699 of 1957 (Frisch & Herzog).

621.373.52 **2665**

Experimental and Theoretical Investigation of a Frequency-Stabilized Transistor Oscillator for 8 Mc/s.—H. Schaffhauser & M. J. O. Strutt. (*Arch. elekt. Übertragung*, Nov. 1957, Vol. 11, No. 11, pp. 455–460.) The condition for oscillation of a feedback oscillator is derived. The amplifier, its load and the feedback network are treated as a single cascaded quadripole. Output voltage and frequency variations as a function of temperature, load and supply voltage were measured on a practical circuit containing a quartz crystal resonator. A general rule for adjusting the emitter series resistance so that the input parameters remain constant during variations of temperature could not be established.

621.373.52 **2666**

Designing Transistor Circuits—Sinusoidal Transistor Oscillators: Part 1.—R. B. Hurley. (*Electronic Equipm.*, Sept. 1957, Vol. 5, No. 9, pp. 22–25.) A criterion for the generation of a single sinusoidal waveform is developed from linear network theory, and applied to basic transistor oscillator circuits. Practical problems of distortion are discussed.

621.373.52.012.8 **2667**

Designing Transistor Circuits—Small-Signal Parameters and Equivalent Circuits.—R. B. Hurley. (*Electronic Equipm.*, Nov. 1957, Vol. 5, No. 11, pp. 28–33.) The merits of various equivalent circuits for transistors are discussed, with emphasis on the common-base-derived *y*, *h* and T arrangements.

621.374.4+621.376.223] : 621.314.632 **2668**

Performance of Three-Millimetre Harmonic Generators and Crystal Detectors.—J. M. Richardson & R. B. Riley. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 131–135. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

621.375 **2669**

A Method for Investigating Amplifying Circuits with Characteristic Equations of the Third Degree for Short Build-Up Times.—E. N. Mokhov. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 54–61.) A method is proposed in which a number of difficulties in design are eliminated. As an example, results are given of an investigation of a correction circuit using an inductance in the grid circuit.

621.375.1.029.3 : 621.395.625.3 : 621.396.665 **2670**

Squelch Circuit mutes Magnetic-Tape Echoes.—D. Cronin. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 66–67.) A biased-diode type of a.v.c. is described which eliminates 'echoes' due to print-through effect by rejecting outputs lower than 40 dB below peak signal level.

621.375.2.018.75 **2671**

The 'Cathoguard'.—L. G. White. (*Wireless World*, July 1958, Vol. 64, No. 7, pp. 312–313.) A method of reducing the input and output capacitances of a wide-band pulse amplifier is described.

621.375.226 : 621.372.54 **2672**

The Asymmetry of Two-Circuit Tuned and Coupled Filters due to Feedback via the Grid-Anode Capacitance C_{ga}.—E. G. Woschni. (*Hochfrequenztech. u. Elektroakust.*, July 1957, Vol. 66, No. 1, pp. 15–19.) The asymmetry due to feedback in the characteristics of an i.f. amplifier is calculated (*a*) for a circuit with tuned filters, and (*b*) for coupled filters. The limiting conditions for instability are also determined. See also 2689 below.

621.375.4.126.029.3 **2673**

Transistor *Q*-Multiplier for Audio Frequencies.—G. B. Miller. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 79–81.) The use of positive feedback to increase the selectivity of a tuned circuit is discussed. The performance of a practical circuit giving, for example, a *Q* of 1 000 at 200 c/s is described.

621.375.4.13 **2674**

Transistor A.C. Amplifier uses Multiple Feedback.—H. Lefkowitz. (*Electronics*, 23rd May 1958, Vol. 31, No. 21, pp. 84–85.) The use of shunt and series negative-feedback loops simultaneously in each stage is considered.

621.375.9 : 537.312.62 : 621.317.32.024 **2675**

A Sensitive Superconducting 'Chopper' Amplifier.—A. R. De Vroomen & C. Van Baarle. (*Physica*, Aug. 1957, Vol. 23, No. 8, pp. 785–794.) An amplifier based on a magnetically modulated Tl wire and similar in principle to that due to Templeton (518 of 1956) is described. It is designed for the study of thermoelectricity at liquid-He temperatures.

621.375.9 : 538.569.4.029.64 **2676**

21-Centimetre Solid-State Maser.—S. H. Autler & N. McAvoy. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 280–281.) A three-level solid-state maser using $K_3Co(CN)_6$ doped with $K_3Cr(CN)_6$ has been operated as an amplifier at 1 382 Mc/s. The design is somewhat different from that previously reported [2037 of July (Artman et al.)] which operates in the same frequency range.

621.375.9 : 538.569.4.029.64 **2677**

Two-Level Solid-State Maser.—P. F. Chester, P. E. Wagner & J. G. Castle, Jr. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 281–282.) Microwave amplification and oscillation have been observed at 4.2°K using two-level electron-spin systems. The materials used were single crystals of quartz and of magnesium oxide, each containing paramagnetic defects.

621.375.9 : 538.569.4.029.6 **2678**

Maser Oscillator with One Beam through Two Cavities.—W. H. Wells. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 714–717.) The behaviour is investigated using a geometrical representation of the Schrödinger equation, and a qualitative explanation is given of the ability of the maser to oscillate simultaneously at two frequencies.

621.375.9 : 621.3.011.23 : 621.314.63 **2679**

A Low-Noise Wide-Band Reactance Amplifier.—B. Salzberg & E. W. Sard. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, p. 1303.)

621.375.9 : 621.3.011.3 **2680**

Circuit Analogues of Suhl-Type Masers.—K. W. H. Stevens. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 275–279.) A tuned circuit in which inductance varies with time, and two coupled tuned circuits in which mutual inductance varies with time, are analysed and shown to act as amplifiers under certain conditions. These considerations are preliminary steps to the final problem leading to the Suhl-type maser (3076 of 1957). See also 2681 below.

621.375.9 : 621.3.011.3 **2681**

Amplification using a Precessing Magnetic Moment.—K. W. H. Stevens. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 280–284.) "An account is given of the analysis of an arrangement in which two tuned circuits are coupled to a precessing magnetic moment. It is shown that with $\omega_1 + \omega_2 = \omega$, where ω_1 and ω_2 are the resonance frequencies of the circuits and ω the precession frequency, useful amplification can be obtained at ω_1. In the course of the analysis the Bloch equations are solved to a higher order in approximation than is usual."

621.375.9 : 621.385.029.63 : 537.533 **2682**

Parametric Amplification of the Fast Electron Wave.—Adler. (See 2934.)

621.375.9.029.63 : 621.3.011.23 : 621.314.63 **2683**

Experimental Characteristics of a Microwave Parametric Amplifier using a Semiconductor Diode.—H. Heffner & K. Kotzebue. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, p. 1301.) A rectangular cavity simultaneously resonant at 3·5, 2·3, and 1·2 kMc/s is used with a Ge junction diode having a zero-bias capacitance of 1 pF and a spreading resistance of 5 Ω. With a pumping signal applied at 3·5 kMc/s, amplification can be obtained at either of the lower frequencies. At 16 dB gain preliminary measurements show a noise figure less than 4·8 dB.

621.375.9.029.64 : 621.3.011.23 : 621.314.63 **2684**

Noise-Figure Measurements on Two Types of Variable Reactance Amplifiers using Semiconductor Diodes.—G. F. Herrman, M. Uenohara & A. Uhlir, Jr. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1301–1303.) The use of semiconductor diodes in an 'up-converter' (460 Mc/s to 9·4 kMc/s) and in a negative-resistance amplifier at 6 kMc/s is described. A bandwidth of 8 Mc/s was obtained with the latter at a maximum gain of 18 dB, and a noise figure as low as 3 dB has been achieved.

621.376 **2685**

An Experimental Investigation of the Method for Obtaining an Optimum Amplitude Phase Modulation.—S. I. Tetel'baum & Yu. G. Grinevich. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 42–47.) A report on experiments confirming the main premises of the theory of the method proposed by Tetel'baum (*ibid.*, Feb. 1950, Vol. 5, No. 2). See 257 of 1956 (Vereschagin).

621.376 : 538.632 : 537.311.33 **2686**

Balanced Modulators based on the Hall Effect in Semiconductors.—L. S. Berman, S. S. Raĭkhman & Z. A. Khalfin. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1597–1598.) Balanced modulators based on the Hall effect in HgSe and *n*-type Ge are investigated. Sensitivities are too low for practical use.

621.376.232 : 621.317.61 : 621.372.8 **2687**

A Phase-Sensitive Detector.—K. G. Beauchamp. (*Electronic Ind.*, March 1958, Vol. 17, No. 3, pp. 74–77.) A detailed circuit is given of a system for measuring the voltage s.w.r. in a waveguide by detecting a modulated carrier. By using a very narrow effective bandwidth detection is possible at signal levels as low as 0·01 μV.

621.376.232.2 **2688**

Analysis of Transient Processes in Diode Detection by a Method based on the Use of Low-Frequency Equivalent Circuits.—L. S. Gutkin & O. S. Chentsova. (*Radiotekhnika, Mosk.*, June 1957, Vol. 12, No. 6, pp. 31–44.) A method is proposed for analysing the transient processes in a system consisting of a h.f. amplifier and a diode detector. The method is based on the linear approximation of the processes occurring during detection and on the replacement of the detector and of the amplifier circuits by equivalent l.f. circuits. Various types of circuit and various shapes of the envelopes of the input signal are considered.

621.376.3 : 621.372.54 **2689**

The Distortion of Frequency-Modulated Oscillations due to Asymmetry in Two-Circuit Filter Networks.—E. G. Woschni. (*Hochfrequenztech. u. Elektroakust.*, July 1957, Vol. 66, No. 1, pp. 11–15.) Second- and third-order distortion factors for quasi-static and dynamic distortion conditions are calculated using formulae derived from a series expansion of the filter phase and attenuation characteristics.

GENERAL PHYSICS

535.325 : 535.14 **2690**

Quantum Theory of the Refractive Index.—C. A. Mead. (*Phys. Rev.*, 15th April 1958, Vol. 110, No. 2, pp. 359–369.) A perturbation theory is used to obtain a unitary transformation which approximately diagonalizes the total Hamiltonian for a system made up of a radiation field in interaction with gas molecules.

535.37 **2691**

A Model of Phosphors on the Basis of Quantum Mechanics: Part 1—Kinetics of Electron and Phonon Reactions.—H. Stumpf. (*Z. Naturf.*, Feb. 1957, Vol. 12a, No. 2, pp. 153–167.) The optical behaviour of a phosphor is discussed assuming that two types of defect centre exist in the crystal.

537.226.1 **2692**

A Property of Dielectric Constants of Dielectrics in Thermal Equilibrium.—M. Peter. (*J. Math. Phys.*, Jan. 1958, Vol. 36, No. 4, pp. 347–350.) Theoretical discussion of dielectrics with or without internal sources of energy. A lumped-circuit model of a molecular amplifier with Lorentz-shaped bandpass characteristic is derived.

537.311.33 : 539.2 **2693**

Irreversible Thermodynamics and Carrier Density Fluctuations in Semiconductors.—K. M. van Vliet. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 50–61.) The formalism of irreversible thermodynamics is applied to the kinetics of carrier transitions in semiconductors. The thermodynamic forces, the generalized resistances and admittance functions are introduced. It is shown that the thermodynamic forces which establish the regression of a perturbed state to equilibrium are the differences of the quasi-Fermi levels that have to be assigned to each group of carriers; the generalized resistances are simply related to the transition rates. The kinetic equations for the rate of change of the various carrier concentrations are then written in a unified form, the spectral density matrix of the spontaneous carrier fluctuations being found from the admittance matrix. The results are then expressed in a closed form which is valid for nondegenerate as well as for degenerate semiconductors. The theory is applied explicitly to electronic noise in extrinsic and near-intrinsic crystals with and without recombination centres.

537.311.62 **2694**

The Theory of the Anomalous Skin Effect in Metals for Obliquely Incident Radiation.—J. G. Collins. (*Appl. sci. Res.*, 1958, Vol. B7, No. 1, pp. 1–40.) The effective surface impedance of a metal varies with the direction of polarization of incident radiation in the upper visible and ultraviolet ranges. Formulae for refractive indices, reflectivities and principal angles at oblique incidence are derived in terms of two surface impedances. Experimental results are in good agreement with the theory.

537.5 **2695**

The Microfield in Assemblies with Coulomb Interaction.—G. Ecker. (*Z. Phys.*, 22nd July 1957, Vol. 148, No. 5, pp. 593–606.) Critical assessment of Holtsmark's theory in the light of the results of recent investigations. See also 384 of 1956.

537.5 : 538.63 **2696**

Investigation of the Transverse Motion of Ions in a Discharge in a Strong Longitudinal Magnetic Field.—A. Zharinov. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1803–1810.) The influence of the gas pressure on the average transverse velocity of ions was investigated using a rotating probe. In an argon discharge, it was found that for pressures $<3\times10^{-3}$ mm Hg and in spite of a strong magnetic field of about 2 300 oersteds, a considerable quantity of ions reached the side walls of the discharge chamber, so that an estimate could be made of the energy acquired by the ions in the radial electric field.

537.52 **2697**

The Demonstration of Single Electron Avalanches and their Secondary Processes in Gases.—J. K. Vogel. (*Z. Phys.*, 8th May 1957, Vol. 148, No. 3, pp. 355–373.) Continuation of earlier experimental investigations [2413 of 1957 (Vogel & Raether)] on electron avalanches in oxygen, nitrogen and air. An interpretation of the observed delay in the formation of avalanches in nitrogen and air is given.

537.523.3 **2698**

Electrical Discharges between Coaxial Electrodes.—J. S. T. Looms. (*Nature, Lond.*, 8th March 1958, Vol. 181, No. 4610, pp. 696–697.) Luminous corona discharges occur when transient voltages are applied between a wire and a small coaxially mounted ring in air. Photographs show that the shapes of the discharges differ according to the polarity of the electrodes.

537.525 **2699**

The Initiation of a V.H.F. Pulse Discharge in Neon.—V. E. Golant. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1482–1494.) An investigation carried out at 2 800 Mc/s with pulses of duration 10^{-6} sec is described. The distribution function of the electron velocities, the kinetic coefficient of the electron interaction with the gas molecules, the h.f. conductivity and the onset of discharge induced by a rectangular h.f. pulse are examined; the results are presented graphically.

537.53 **2700**

Pulse Ionization Chamber as a Device for Simultaneous Investigation of the

Energy and Angular Distributions of Charged Particles.—B. A. Bochagov, A. A. Vorob'ev & A. P. Komar. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1575–1577.)

537.533 **2701**

Small Oscillations in an Electron Beam.—R. V. Polovin & N. L. Tsintsadze. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1466–1473.) The stability conditions for the propagation of small-amplitude e.m. oscillations in noncompensated electron streams are examined and solutions for some particular cases are obtained.

537.533 **2702**

On the Space-Charge Divergence of an Axially Symmetric Beam.—E. R. Harrison. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 193–200.) The usual assumptions made in estimating the space-charge effect are consistent with a coaxial, cylindrical conducting boundary. For small amounts of divergence a hyperbola solution is suggested; this may in some cases be as good as or better than the usual solution even when divergence is not small.

537.533 **2703**

Small-Signal Power Conservation Theorem for Irrotational Electron Beams.—J. W. Klüver. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 618–622.) The theorem describes the balance between the r.f. kinetic and r.f. electromagnetic powers in a laminar beam; the thermal velocity distribution at the cathode is neglected. The kinetic power flow is expressed in terms of the Chu potential, the longitudinal component of the r.f. beam current and the equivalent longitudinal surface current on the beam boundary. See also 3701 of 1957 (Hans & Bobroff).

537.533.8 **2704**

Variation of Secondary Emission with Primary Electron Energy.—B. K. Agarwal. (*Proc. phys. Soc.*, 1st May 1958, Vol. 71, No. 461, pp. 851–852.) A simple formula is given, relating secondary emission to primary electron energy, which is more in agreement with experimental results for MgO and Ge than previous theoretical formulae.

537.56 **2705**

Effective Cross-Sections of Electrostatic Interaction in Plasmas. Definitions.—M. Bayet. (*C. R. Acad. Sci., Paris*, 13th Nov. 1957, Vol. 245, No. 20, pp. 1708–1710.) For calculations see *ibid.*, 23rd Dec. 1957, Vol. 245, No. 26, pp. 2493–2496. See also 1111 of April.

537.56 **2706**

Effects of Electron-Electron Interactions on Cyclotron Resonances in Gaseous Plasmas.—R. C. Hwa. (*Phys. Rev.*, 15th April 1958, Vol. 110, No. 2, pp. 307–313.) From the Boltzmann-Fokker-Planck equation, a set of equations is derived and solved numerically. The results indicate that the real part of the electrical conductivity of the plasma (power absorption) is reduced by electron-electron interaction at the peak of resonance, and the width of resonance is increased.

537.56 **2707**

Nonlinear Phenomenological Theory of Plasma Oscillations.—L. Gold. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 219–226.) Damped oscillations are shown to arise under certain conditions; the intensity of the standing waves being augmented by collisions with a loss of efficiency. In general, increasing current and plasma temperature tend to promote oscillations, while the role of plasma density is more complex. The influence of a magnetic field in the limiting case of no damping is also investigated.

538.3 : 530.12 **2708**

Completely Integrable Singular Electromagnetic Field.—L. Mariot. (*C. R. Acad. Sci., Paris*, 21st Oct. 1957, Vol. 245, No. 17, pp. 1386–1388.) Derivation of expressions for the components of the completely integrable singular electromagnetic tensor with reference to Maxwell's equations.

538.524 **2709**

Induction by an Oscillating Magnetic Dipole over a Two-Layer Ground.—J. R. Wait. (*Appl. sci. Res.*, 1958, Vol. B7, No. 1, pp. 73–80.) Expressions for the mutual electromagnetic coupling between two small loops and for the self-impedance of a loop over a two-layer earth are derived.

538.566 : 531.51 : 530.12 **2710**

Effect of a Gravitational Field, due to a Rotating Body, on the Plane of Polarization of an Electromagnetic Wave.—N. L. Balazs. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 236–239.) "It is shown that for electromagnetic waves the gravitational field of a rotating body acts as an optically active medium. Thus, the plane of polarization of the wave rotates while it passes through this field. The effect is small. The angle of rotation due to the gravitational field of the sun is about 10^{-12} radian."

538.566 : 535.42 **2711**

Approximate Calculations of the Diffraction of Plane Electromagnetic Waves by some Metallic Bodies: Part 1.—P. Ya. Ufimtsev. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1840–1849.) A new approximate method of calculation is described which is applicable to the diffraction of e.m. waves by wedges or strips. The method applies to wavelengths much smaller than the linear dimensions of the diffracting body.

538.566 : 537.56 **2712**

Basic Microwave Properties of Hot Magnetoplasmas.—J. E. Drummond. (*Phys. Rev.*, 15th April 1958, Vol. 110, No. 2, pp. 293–306.) The usual conductivity tensor of a uniform plasma in a uniform static magnetic field is generalized to include four effects of random motion of plasma electrons. The results are applied to the evaluation of the index of refraction for microwave signals.

538.569.4 : 538.22 : 537.228 **2713**

Investigations of the Paramagnetic Nuclear Resonance Absorption in a Capacitor Field.—U. Gersch & A. Lösche. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 167–172.) The advantages and disadvantages of using a capacitor field for high-resolution nuclear induction spectroscopy are briefly outlined.

538.569.4 : 539.14.098 **2714**

Theory of Saturation in Nuclear Magnetic Resonance.—G. Vojta. (*Z. Naturf.*, April 1957, Vol. 12a, No. 4, pp. 282–294.)

539.2 **2715**

Energy-Band Splitting and Inter-band Terms in the Dislocation of Lattice Atoms in a Solid.—A. Haug & A. Schönhofer. (*Z. Phys.*, 22nd June 1957, Vol. 148, No. 4, pp. 513–526.)

539.2 : 537.226 **2716**

On the Theory of a Fast Polaron.—Yu. I. Gorkun. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1764–1769.) Expressions for the dependence of the polaron energy on its initial velocity are obtained. In connection with polarons, the breakdown voltages for ionic dielectrics are also examined. Results of investigations on alkali halide crystals are tabulated.

539.2 : 538.222 **2717**

Spin-Lattice Relaxation.—M. W. P. Strandberg. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 65–69.) It is pointed out that phonon relaxation times can be the dominant quantity measured in the usual saturation spin-lattice relaxation measurements. The analysis indicates how pulse measurements may be used to evaluate the actual spin-lattice relaxation time independently of the phonon relaxation time.

539.2 : 548.0 **2718**

One-Dimensional Impurity Bands.—M. Lax & J. C. Phillips. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 41–49.) "The density of states of one-dimensional crystals consisting of δ functions randomly distributed has been calculated on the IBM 650 computer. The chains contained 500–1 000 impurity atoms, and the most probable error in the integrated density of states at various energies was estimated to be at most ½%. Calculations were performed for various values of the parameter $\epsilon = n/\kappa_0$, where n is the density of atoms and κ_0 the attenuation constant appropriate to the isolated bound state. The results at different densities are compared with those obtained from various physical models."

539.2 : 548.0 **2719**

The 'Spur' [trace] Method in the Problem of the Exciton.—I. M. Dykman. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1731–1743.) The excitons in ionic crystals are examined. The interaction of electrons and holes with lattice vibrations is considered. It is shown that the change of energy at low temperatures is primarily determined by the weakening of the interaction of charged particles with the field of polarized vibrations when the temperature rises without increase of kinetic energy. The effective mass of the exciton is calculated. See also 3124 of 1957 (Dykman et al.) and 2796 of 1957.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 **2720**

The Initial Radius of Meteoric Ionization Trails.—L. A. Manning. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 181–196.) Kinetic theory is applied to the formation of regions of ionized and neutral atoms of meteoric material about the path of a meteor. It is found that the high initial velocity of the diffusing particles causes the trail to reach an 'initial radius' quickly, and normal diffusion then ensues. A reflected signal may be computed as if the initial radius (14 mean free paths) were reached instantaneously.

523.5 : 621.396.11.029.62 **2721**

The Random Occurrence of Meteors in the Upper Atmosphere.—T. J. Keary & H. J. Wirth. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 67–75.) The distribution of time intervals between bursts of 43·5-Mc/s signals over a 690-km path was studied statistically, using data obtained in two periods of two hours (778 and 446 bursts respectively). Results were consistent with forward-scattering from meteor trails produced by random entry of meteors into the atmosphere.

523.74 **2722**

Identification of *M*-Regions on the Sun.—K. Toman. (*Nature, Lond.*, 1st March 1958, Vol. 181, No. 4609, pp. 641–642.)

523.752 **2723**

The Duration of Emission and the Slope of the Pulse Spectrum of Solar Ultraradiation during the Chromospheric Eruption of 23/2/1956.—G. Pfotzer. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 26–41.) Recordings made at various stations are analysed and discussed in relation to various theoretical models. 52 references.

523.755 : 523.164.32 **2724**

The Calculation of the Centre–Limb Variations and the One-Dimensional Intensity Profiles at $\lambda = 20$ cm and $\lambda = 60$ cm for a Rotationally Symmetric Corona Model at Constant Temperature.—G. Wallis. (*Z. Naturf.*, April 1957, Vol. 12a, No. 4, pp. 337–345.) The calculated results are compared with those obtained from measurements by Christiansen & Warburton (1706 of 1956) and Swarup & Parthasarathy (1707 of 1956).

550.385 : 523.75 **2725**

Geomagnetic Disturbances associated with Solar Flares with Major Premaximum Bursts at Radio Frequencies $\leqslant$ 200 Mc/s.—H. W. Dodson & E. R. Hedeman. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 77–96.) A close association is found between solar flares in the early phase of which radio noise at frequencies $\leqslant$ 200 Mc/s is received, and sudden-commencement ionospheric storms on the earth. It is suggested that this r.f. emission may be from the sun's outer atmosphere and corona and thus may indicate which solar flares are accompanied by the ejection of storm-producing particles. Although flare data are incomplete, the proportion of flares having 'major early bursts' is estimated at $< 10\%$, and the number of flares is comparable to the number of ionospheric storms not identifiable with well-established 27-day recurrent series. An average time delay of 2½ days occurs between flare and geomagnetic disturbance, and the intensity of the disturbance is related to the position of the flare on the sun's disk.

550.389.2 : 629.19 **2726**

Progress of Sputnik 2 (1957β).—D. G. King-Hele. (*Nature, Lond.*, 15th March 1958, Vol. 181, No. 4611, pp. 738–739.) Estimates of the orbital period, semi-major axis and eccentricity on 9th Nov. 1957, 2nd March 1958 and two intermediate dates are given. These are based on observational data from different sources.

550.389.2 : 629.19 **2727**

Determination of Orbital Characteristics of an Earth Satellite from Single-Station Radio-Transit Observations.—G. Grant, A. L. Jones, R. W. Burhans, P. S. Fay & D. Frazier. (*Nature, Lond.*, 29th March 1958, Vol. 181, No. 4613, pp. 900–901.) Meridian observations of the satellite 1958α have been made by means of an interferometer located at Cleveland, Ohio. Because of the generally low altitude of the satellite at this station horizontal dipoles were replaced by commercial Yagis.

551.510.535 **2728**

The Mesopause Region of the Ionosphere.—J. B. Gregory. (*Nature, Lond.*, 15th March 1958, Vol. 181, No. 4611, pp. 753–754.) Observations have been made at 1·75 Mc/s with pulse sounding apparatus at Christchurch, New Zealand, since early in 1955. Initial results (see 1433 of 1957) are confirmed. Extended observations indicate the continuous existence of ionization throughout a range of heights from about 80 to 100 km.

551.510.535 **2729**

New Rocket Measurement of Ionospheric Currents near the Geomagnetic Equator.—L. J. Cahill, Jr, & J. A. Van Allen. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 270–273.) Rocket-borne magnetometer measurements indicating current sheets at heights of 97–110 km and 118–121 km are described. Comparison with other data indicate that at 121 km only half the total current system had been penetrated.

551.510.535 **2730**

Ionosphere Electron-Density Measurements with the Navy Aerobee-Hi Rocket.—J. E. Jackson & J. C. Seddon. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 197–208.) Electron densities were measured at heights between 80 km and 260 km above New Mexico by the Doppler-shift method in which the beat note between a 7·75-Mc/s signal and a reference signal at 46·5 Mc/s was recorded. The results, corrected for obliquity effects, confirm the general structure of the ionosphere previously measured, and are consistent with simultaneous $P'f$ records; the ionosphere remains dense between E and F_2 regions. Sporadic E was associated with a sharp spike in the distribution at 101 km, approximately 1 km thick.

551.510.535 **2731**

Differential Absorption in the D and Lower E Regions.—J. C. Seddon. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 209–216.) A method of measuring electron densities in the D region is described, in which use is made of the Faraday effect in rocket flights at White Sands, New Mexico. At noon on a day of high absorption, a value of 2 000 electrons/cm^3 was obtained for an altitude of 76 km.

551.510.535 : 523.5 **2732**

Concerning Ionospheric Turbulence at the Meteoric Level.—H. G. Booker. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 97–107.) A discussion on the application of fluid mechanics to meteoric phenomena is continued, with particular reference to a 'rough trail paradox', which concerns the relative turbulent velocities of large and small eddies. The paper is mainly concerned with the relative merits of work by Booker & Cohen (1417 and 1441 of 1957) and by Manning & Eshleman (111 of January) regarding the explanation of radar echo fading.

551.510.535 : 523.75 **2733**

Solar-Flare Effect in the F_2 Layer of the Ionosphere.—C. M. Minnis & G. H. Bazzard. (*Nature, Lond.*, 8th March 1958, Vol. 181, No. 4610, pp. 690–691.) Δf is plotted as a function of local mean time, where Δf is the difference between the measured critical frequency and the 'normal' value based on that measured one hour previously. The Δf plot for the period of the intense solar flare of 23rd February 1956 shows a positive 'pulse' of amplitude 1·1 Mc/s followed by a second smaller pulse 5 h later. An analysis of the statistics of the fluctuations in Δf has been made.

551.510.535 : 523.75 **2734**

Medium-Frequency Observations of the Lower Ionosphere during Sudden Disturbances.—J. B. Gregory. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 273–275.) Records obtained during S.I.D.'s by pulse sounding at 1·75 Mc/s are analysed in detail. The observed behaviour gives support to the theory of Friedman and Chubb (*The Physics of the Ionosphere*, 1955, pp. 58–62) that soft X-rays and not Lyman-α radiation are responsible for S.I.D.'s.

551.510.535 : 550.385 **2735**

Variations in the Height of Ionospheric Layers during Magnetic Storms.—E. Tandberg-Hanssen. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 157–160.) Statistical work carried out on Washington ionospheric data for 1952–1956 and magnetic planetary K_p figures from C.R.P.L. publications, shows that the E layer remains unchanged in height, while the F_1 layer rises during storms. Application to Parker's theory (1422 of 1957) gives 100–200 km as the height where storm-associated solar energy is absorbed.

551.510.535 : 551.515.2 2736

An Apparent Ionospheric Response to the Passage of Hurricanes.—S. J. Bauer. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 265–269.) Evidence is given for the association of the passage of a hurricane with a departure of the critical frequency of the F_2 layer from its monthly mean value. The four hurricanes discussed resulted in rises in critical frequency of up to 2 Mc/s. A possible mechanism is suggested.

551.594.6 2737

Radio Noise from Lightning Discharges.—F. Horner & C. Clarke. (*Nature, Lond.*, 8th March 1958, Vol. 181, No. 4610, pp. 688–690.) Records obtained at Slough during two local thunderstorms in July and August 1957 are reproduced including simultaneous recordings on 11 Mc/s and 6 kc/s with bandwidths 300 and 200 c/s respectively.

551.594.6 2738

Between the Atmospherics.—G. Reber. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 109–123.) An examination is made of the background cosmic radiation between atmospherics at a frequency of 520 kc/s. Records are explained in terms of an 'ionospheric shutter' operated by two mechanisms: D-region absorption by day, and shift of coupling level between ordinary and extraordinary modes relative to the gyro-level at night. Observations of precipitation static and local atmospherics are also described.

550.389.2 2739

Annals of the International Geophysical Year: Vol. 3—Parts 1, 2, 3 & 4. [Book Review]—Publishers: Pergamon Press, London and New York, 1957, 381 pp., £6 or $17. (*Nature, Lond.*, 15th March 1958, Vol. 181, No. 4611, pp. 726–727.) The first volume to be published of a series of instruction manuals. Titles of the parts of Vol. 3 are: Ionospheric Vertical Soundings, Measurement of Ionospheric Absorption, the Measurement of Ionospheric Drifts, and Miscellaneous Radio Measurements.

LOCATION AND AIDS TO NAVIGATION

621.396.9 2740

Doppler Effect in Radio and Radar.—N. M. Rust. (*Wireless World*, July & Aug. 1958, Vol. 64, Nos. 7 & 8, pp. 304–307 & 373–377.) The general principles of the Doppler effect are explained; it is shown how directional discrimination can be obtained using phase-sensitive detectors. Various practical applications are described.

621.396.93 2741

Dectra: a Long-Range Radio-Navigation Aid.—C. Powell. (*J. Brit. Instn Radio Engrs*, May 1958, Vol. 18, No. 5, pp. 277–290. Discussion, pp. 291–292.) The tracking and ranging functions of the system are discussed with reference to the time-sharing technique on which the tracking pattern is based. The receiving and display equipment is described and details are given of the transmitting stations which are common to the Dectra and Decca services. Accuracy and performance are considered.

621.396.932.1 : 621.396.677.53 2742

The Polarization Direction-Finder.—J. Grosskopf & K. Vogt. (*Nachrichtentech. Z.*, Nov. 1957, Vol. 10, No. 11, pp. 572–579.) Report on experimental investigations made in 1944 on behalf of the German navy on a s.w. d.f. system with a combined dipole and loop aerial. Three different arrangements are considered: (*a*) vertical loop with vertical dipole, (*b*) horizontal loop with horizontal dipole, and (*c*) vertical loop with horizontal dipole. The theory of operation of the system is detailed and diagrams of phase and amplitude difference as a function of azimuth angle are given for various types of polarization. A combination of arrangements (*a*) and (*b*) is the most generally useful. Interference due to the incidence of two waves is also considered. The polarization type of direction finder can provide results as accurate as those obtained by the Adcock system with the additional advantage of a much more compact aerial arrangement.

621.396.933.2 2743

Long-Range Radio Beacons.—K. Lutz. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 451–452.) A 10-kW radio beacon for aircraft and marine navigation is described. It operates in the 200–415-kc/s band and the transmitter can be used for telephony.

621.396.933.2 2744

On the Accuracy of Determining Position by Radio Navigation Methods.—A. G. Saibel'. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 62–66.) Various methods for assessing accuracy are reviewed and compared on the basis of distribution functions of error probabilities. It is suggested that this accuracy should be characterized by the mean-square error in position.

621.396.96 : 551.515.3 2745

R.H.I. Radar Observation of a Tornado.—J. Schuetz & G. E. Stout. (*Bull. Amer. met. Soc.*, Dec. 1957, Vol. 38, No. 10, pp. 591–595.)

621.396.967 : 621.396.932 2746

The Shore-Based Radar System for the New Rotterdam Waterway.—N. Schimmel. (*Tijdschr. ned. Radiogenoot.*, March 1957, Vol. 22, No. 2, pp. 59–86. In English.)

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 2747

Photoconductivity.—D. A. Wright. (*Brit. J. appl. Phys.*, June 1958, Vol. 9, No. 6, pp. 205–214.) The general process of photoconductivity is reviewed, and the preparation and properties of photoconducting materials are described, in particular CdS. The effects on sensitivity of defects in the crystal lattice and of the nature of impurities are discussed. Applications are outlined. 34 references.

535.215 : 537.311.33 2748

Diffusion of Charge Carriers in Photoconductors taking account of Local Levels.—J. Auth. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 210–214.) A model is derived for an n-type photoconductor and approximate solutions of the diffusion equations are obtained.

535.215 : 546.482.21 2749

The Influence of Dielectric After-Effects on Conduction in CdS Single Crystals in the Region of Breakdown Field-Strength.—K. W. Böer & U. Kümmel. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 303–314.) See also 135 of 1956 and 3792 of 1956 (Böer et al.).

535.215 : 546.817.231 2750

The Effect of Various Gases and Vapours on the Semiconductor Properties of Vapour-Deposited Lead Selenide Films.—H. Gobrecht, F. Niemeck & K. E. Boeters. (*Z. Phys.*, 8th May 1957, Vol. 148, No. 3, pp. 281–297.) Tests were made on specimens prepared by the method outlined in 2634 of 1955 (Gobrecht et al.) to determine the effect of Hg and I vapours on photoconductivity and dark resistance. Measurements were also carried out with air, oxygen, nitrogen, hydrogen and argon to find the relation between photoconductivity and gas pressure.

535.37 : 534-8 2751

The Effect of Ultrasonic Radiation on the Luminescence of Zinc Sulphide Fluorescent Screens under Continuous Excitation by Light.—M. Leistner. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 129–141.) Full report and analysis of test results previously noted in 3158 of 1957 (Leistner & Herforth).

535.376 2752

Enhancement by Electric Fields of the Sensitivity of Certain Luminescent Products to X Rays, and 'Photoelectroluminescence'.—G. Destriau. (*C. R. Acad. Sci., Paris*, 18th Nov. 1957, Vol. 245, No. 21, pp. 1797–1800.) Research on (CdS, ZnS)-Mn does not support the theory of Cusano (2980 of 1955) and Williams (2981 of 1955).

535.376 2753

Sensitization, by Traces of Gold, of Products with an Electroenhancement Effect.—G. Destriau. (*C. R. Acad. Sci., Paris*, 25th Nov. 1957, Vol. 245, No. 22, pp. 1913–1916.) The addition of traces of Au to the Mn activator in CdS-ZnS mixtures increases both the enhancement ratio, defined as the ratio of brightness under combined X-ray and electric-field excitation to that under X-ray excitation alone, and the field sensitivity.

537.226/.228.2 : 546.431.824-31 2754

Two Experiments on the Electromechanical Characteristics of Barium

Titanate Ceramics in the Curie Region.—G. Schmidt. (*Z. Phys.*, 8th May 1957, Vol. 148, No. 3, pp. 314–320.) The results are discussed of further tests on thickness variation as a function of polarization and field strength, (*a*) with a superimposed 400-c/s field, and (*b*) at field strengths up to 3 kV/cm. See also 465 of 1957.

537.226/.228.1 : 546.431.824-31 **2755**

Piezoelectric and Dielectric Characteristics of Single-Crystal Barium Titanate Plates.—A. H. Meitzler & H. L. Stadler. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 719–738.) The dielectric constant of *c*-domain $BaTiO_3$ single-crystal plates has been measured as a function of their average polarization, using alternating voltages much smaller than the coercive voltage. At frequencies well above thickness resonance the dielectric constant has a maximum at zero polarization. The piezoelectric output voltage of a single crystal subjected to a recurrent strain pulse of constant amplitude is not exactly proportional to its average polarization. Plate-shaped single crystals have a significantly lower dielectric constant and smaller capacitance ratio when used as resonators in the thickness-extensional mode than similar resonators of ceramic $BaTiO_3$.

537.226/.227 : 546.431.824-31 **2756**

High Permittivity of some Solid Solutions with Ferroelectric Properties.—V. A. Bokov. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1784–1793.) Investigations of the temperature dependence of the dielectric constant and the loss angle of solid solutions of the types $(Ba,Sr)TiO_3$, $Ba(Ti,Sn)O_3$ and $Ba(Ti,Zr)O_3$ in weak fields. Results are presented graphically.

537.226/.227 : 546.431.824-31 **2757**

Some Anomalies of the Hysteresis Loop in Ceramic Barium Titanate and Piezolan.—E. Hegenbarth. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 20–25.) Analysis and discussion of hysteresis loop and *V/I* oscillograms. See e.g. 3758 of 1956 (Heywang & Schöfer).

537.227 : 539.2 **2758**

Appearance of Spontaneous Polarization in Crystals.—G. A. Smolenskiĭ. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1778–1783.) Ferroelectric active ions and chemical bonds in ferroelectrics and antiferroelectrics are considered. Some experimental results concerning the structure of electron shells occurring in the oxygen octahedrons of ferroelectrics containing no hydrogen are discussed.

537.227 : 546.431.824-31 : 621.318.57 **2759**

Electrical Stability of $BaTiO_3$ Single Crystals at –195°C.—H. L. Stadler. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 743–744.) Crystals at room temperature, used as matrix-type digital storage systems exhibit a gradual loss of response with continuous switching sequences. This decay does not occur at –195°C and is not primarily dependent on crystal phase.

537.227 : 547.476.3 **2760**

Ferroelectric Polarization Reversal in Rochelle Salt.—H. H. Wieder. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 29–36.) Besides the slow, lateral motion of *b*- and *c*-domain walls, there is observed a fast reversal process characteristic of the nucleation and wall propagation in the orthorhombic *a*-direction.

537.227 : 621.375.5 : 621.396.822 **2761**

Noise of Cyclic Repolarization of Ferroelectrics.—I. A. Andronova. (*Dokl. Ak. Nauk S.S.S.R.*, 1st March 1958, Vol. 119, No. 1, pp. 68–70.) Noise occurring in dielectric amplifiers due to cyclic repolarization is attributed to a ferroelectric domain structure. The temperature dependence of this noise is examined.

537.311.33 **2762**

Semiconductors Again.—(*Wireless World*, July 1958, Vol. 64, No. 7, pp. 339–343.) A nonmathematical explanation, from the energy-level point of view, of the properties of semiconductors.

537.311.33 **2763**

Electrons, Holes, and Traps.—W. Shockley. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 973–990.) "The statistics of recombination and of trapping of electrons and holes through traps of a single species are presented. The results of the Shockley-Read recombination theory are derived and more fully interpreted. A level of energy known as the equality level is introduced. When the Fermi level lies at this level, the four basic processes of electron capture, electron emission, hole capture, and hole emission all proceed at equal rates. Transient cases for large trap density are presented."

537.311.33 **2764**

Recombination in Semiconductors.—G. Bemski. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 990–1004.) The kinetics of hole-electron recombination, either direct or by means of recombination centres, are reviewed. The latter process accounts for recombination in Ge and Si, but the former becomes important in semiconductors with narrower energy gaps. Trapping and surface recombination are also considered in relation to volume recombination. Experimental methods of measuring lifetime are reviewed, and results are discussed in terms of the various recombination processes. 131 references.

537.311.33 **2765**

The Preparation of Semiconductor Devices by Lapping and Diffusion Techniques.—H. Nelson. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1062–1067.) Large semiconductor wafers can be processed to a point where they can be diced into numerous and identical devices. Lapping instead of etching is used for all shaping and the high degree of precision built into the lapping apparatus is passed on to all the devices. Unipolar, photo-unipolar and bipolar transistors and negative-resistance devices have been made.

537.311.33 **2766**

On Junctions between Semiconductors having Different Energy Gaps.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1307–1308.)

537.311.33 **2767**

Thermodynamic Treatment of Electron Processes in Semiconductor Boundary Layers.—H. A. Müser. (*Z. Phys.*, 8th May 1957, Vol. 148, No. 3, pp. 380–390.) An interpretation of the processes occurring in *p-n* junctions is derived on the basis of thermodynamics. The principle is applied to a consideration of photoelectric efficiency and the process of rectification at a *p-n* junction.

537.311.33 **2768**

Semiconductors considered as Ionized Media.—M. Demontvignier. (*Rev. gén. Élect.*, Oct. 1957, Vol. 66, No. 10, pp. 495–512.) An analogy is drawn between the theory of tetravalent semiconductors and the theory of electrolytes.

537.311.33 **2769**

The Temperature Dependence of Mobility in Nonpolar Semiconductors.—D. Dorn. (*Z. Naturf.*, Jan. 1957, Vol. 12a, No. 1, pp. 18–22.) Bloch's theory is extended to allow for electron energy changes due to collision with thermal lattice vibrations. The resulting correction leads to an increase in mobility compared to the $T^{-3/2}$ law.

537.311.33 **2770**

Measurements of Lifetime of Charge Carriers in Semiconductors.—M. I. Iglitsyn, Yu. A. Kontsevoĭ, V. D. Kudin & A. A. Meĭer. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1414–1424.) A theoretical and experimental investigation based on the modulation of conductivity at the point-contact is presented. Formulae taking account of the diffusion and recombination at this point are derived.

537.311.33 **2771**

Simple Methods for Investigating the Zone Structure of some Semiconductor Compounds.—V. E. Khartsiev. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1713–1722.) Some general expressions determining the band structure in semiconductors are derived. After a general survey of the one-dimensional problem, and with the help of matrix relations, three-dimensional lattice models, with Zn and NaCl types of structure, are examined.

537.311.33 **2772**

Low-Temperature Anomalies in Impurity Semiconductors.—M. I. Klinger. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1915–1922.) Theoretical investigation of *p*-type and *n*-type semiconductors. See also 3184 of 1957.

537.311.33 : 061.3 **2773**

All-Union Conference on the Theory of Semiconductors.—M. F. Deĭgen, I. M. Dykman & K. B. Tolpygo. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1628–1642.) Summaries are given of some 40 papers read during the Conference at Kiev, 9th–13th October 1956. The papers are grouped under the following headings: (*a*) multi-electron theory of solids, (*b*) exciton processes in semiconductors, (*c*) interaction of current carriers with the crystal lattice; theory of polaron, (*d*) theory of the local state of an electron in semiconductors, (*e*) the band structure in semiconductors, (*f*) magnetic properties of semiconductors, and (*g*) phenomenological theory of semiconductors.

537.311.33 : 539.16 **2774**

Irradiation of *P-N* Junctions with Gamma Rays: a Method for Measuring Diffusion Lengths.—R. Gremmelmaier. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1045–1049.) The short-circuit current in an irradiated p-n junction is given by the product of the electron charge, the generation rate (g) of minority carriers, and a quantity equal to the diffusion length if the junction position is suitably chosen; g can be calculated for irradiation by γ rays from a Co^{60} source, and the short-circuit current measured. Diffusion lengths are observed up to 8 μ in GaAs and 130 μ in InP.

537.311.33 : [546.28+546.289 **2775**

Properties of Silicon and Germanium —II.—E. M. Conwell. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1281–1300.) Revision of an earlier paper (738 of 1953) giving the physical background of more recent advances and best current values of important physical quantities. 117 references.

537.311.33 : [546.28+546.289 **2776**

Scattering of Holes in Germanium and Silicon.—G. E. Pikus. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1606–1609.) The lattice scattering of heavy and light holes within and between the two valence bands of Ge and Si is considered. The matrix elements and angular distribution for this scattering are examined and the relations of concentrations of light and heavy holes, and also of their respective mobilities are tabulated.

537.311.33 : [546.28+546.289 **2777**

Formation of Junction Structures by Solid-State Diffusion.—F. M. Smits. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1049–1061.) "The diffusion of group III and group V impurities into germanium and silicon is reviewed. Observed and possible variations of the diffusion coefficient with concentration are discussed, followed by a summary of the diffusion coefficients and of solutions to the diffusion equation. Finally, methods for the evaluation of diffused layers and diffusion techniques are described."

537.311.33 : [546.28+546.289 **2778**

Galvanomagnetic Effects in *n*-Type Ge or *n*-Type Si Single Crystals in Strong Magnetic Fields.—M. I. Klinger & P. I. Voronyuk. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1609–1613.) A mathematical analysis is presented in which the electrical resistance and Hall constant of n-type semiconductors are examined.

537.311.33 : [546.28+546.289 **2779**

The Effects of Neutron Irradiation on Germanium and Silicon.—G. C. Messenger & J. P. Spratt. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1038–1044.) A theoretical expression is derived for the dependence of grounded-emitter current gain of a transistor on accumulated neutron dose. Observed changes in transistor parameters after irradiation are explained in terms of this theory, and in the case of Ge enabled the position of the recombination site in the forbidden band, and its capture cross-sections for holes and electrons, to be determined.

537.311.33 : 546.28 **2780**

Preparation of Pure Silicon by the Thermal Decomposition of Silane.—S. I. Kleshchevnikova, Ya. E. Pokrovskiĭ & E. I. Rumyantseva. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1645–1648.) The method and apparatus are described and results presented graphically.

537.311.33 : 546.28 **2781**

Evaluation of the Surface Concentration of Diffused Layers in Silicon.—G. Backenstoss. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 699–710.) A diffused layer in a semiconductor is characterized by the impurity distribution. A method for determining the surface concentration of diffused impurity layers is described. It is shown that the surface concentration can be found for the properties of the junction and calculations have been made for typical impurity distributions in silicon.

537.311.33 : 546.28 **2782**

Inversion Boundary Layers on *p*-Type Silicon.—S. Müller. (*Z. Naturf.*, Feb. 1957, Vol. 12a, No. 2, pp. 112–122.) The formation in water vapour of an inversion layer in p-type Si is investigated and the conductivity of the layer is determined, which is of the order of $10^{-5} - 10^{-6}\,\Omega^{-1}$. Two types of surface state are observed, one in which equilibrium with the conduction band is rapidly reached, and the other with relaxation times of 10–50 minutes. The density of the former states is much higher in Si than in Ge. See also 2434 of 1956 (Statz et al.).

537.311.33 : 546.28 **2783**

Measurement of Lifetime of Charge Carriers in Si Single Crystals.—M. I. Iglitsyn, Yu. A. Kontsevoĭ & V. D. Kudin. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1425–1430.) The lifetime is evaluated by measuring the modulated conductivity in the region of the point contact by passing two consecutive current pulses. The effect of temperature on the charge-carrier lifetime is investigated. The results seem to be in good agreement with the Shockley-Read theory (420 of 1953).

537.311.33 : 546.28 **2784**

Evidence of Dislocation Jogs in Deformed Silicon.—W. C. Dash. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 705–709.)

537.311.33 : 546.28 : 548.5 **2785**

Silicon Crystals Free of Dislocations.—W. C. Dash. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 736–737.) A brief description of a method for growing crystals free from detectable oxygen and without observable dislocations.

537.311.33 : 546.281.46 **2786**

Properties of Silicon Carbide under Pulsed Conditions.—R. Goffaux. (*Rev. gén. Élect.*, Sept. 1957, Vol. 66, No. 9, pp. 463–472.) A model is proposed to explain the observed I/V characteristic; explanations based on the thermal ionization of donor centres and on the tunnel effect are disputed. See 434 of 1954 (Teszner et al.).

537.311.33 : 546.289 **2787**

Energy Spectrum of Current Carriers in Semiconductors of the Germanium Type.—A. G. Samoilovich & K. D. Tovstyuk. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1753–1763.) A mathematical analysis is presented and the energy spectrum of current carriers of Ge and Si is examined. The energy spectrum of holes seems to depend not on the orbital spin interaction but on the distribution of clouds of valence electrons, that is, on the nature of chemical bonds. By a proper choice of wave functions, an electron energy spectrum can be obtained which is in agreement with experimental data.

537.311.33 : 546.289 **2788**

Determination of Relaxation Time of Surface States in Germanium.—A. E. Yunovich. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1707–1712.) Results of measurements of the field effect in the frequency range 400 c/s–80 kc/s on n-type Ge samples indicate the strong dependence of the relaxation time of surface states on the ambient atmosphere.

537.311.33 : 546.289 **2789**

Direct Optical Transitions and Further Exciton Effects in Germanium.—G. G. Macfarlane, T. P. McLean, J. E. Quarrington & V. Roberts. (*Proc. phys. Soc.*, 1st May 1958, Vol. 71, No. 461, pp. 863–866.) Measurements of absorption coefficient in Ge at temperatures in the range 20°K–291°K show a theoretically predicted absorption line, due to exciton production, at a photon energy less than E_0, the energy gap, by an amount E_{ex}, the binding energy of an exciton of zero wave vector. Values of E_{ex} agree well with estimates based on the effective mass, but there is a marked discrepancy between the low-temperature values of E_0 and those obtained by Zwerdling et al. (1449 of May) in work on the oscillatory magneto-absorption effect. See also 1463 of May.

537.311.33 : 546.289 **2790**

Impact Ionization in Germanium Single Crystals in the Temperature Range 4·2°K–10°K.—G. Kinke & G. Lautz. (*Z. Naturf.*, March 1957, Vol. 12a, No. 3, pp. 223–225.) The effect interpreted as impact ionization [see e.g. 754 of 1955 (Ryder et al.)] is investigated as a function of temperature, purity of Ge, and transverse magnetic field strength.

537.311.33 : 546.289 **2791**

Field Modulation of Liquid-Induced Excess Surface Currents on Germanium *p-n* Junctions.—W. T. Eriksen. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 730–733.) Experiments show that modulation is only possible for polar liquids and increases with decreasing temperature down to the melting point of the liquid. A qualitative explanation is discussed in terms of a model in which the charge carriers move in the liquid outside the semiconductor.

537.311.33 : 546.289 : 535.215 **2792**

Change of Lifetime in Thin Germanium Plates under the Action of an External Electric Field and Adsorption.—V. I. Lyashenko. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1613–1615.) Experiments show that the photo-e.m.f. and carrier lifetime in *n*-type and *p*-type Ge increase or decrease according to the direction of the applied field. The effects of the adsorption of liquid films such as acetone, alcohol, benzene, water and vaseline oil are also investigated.

537.311.33 : 546.289 : 535.32 **2793**

Optical Constants of Germanium: 3 600 Å to 7 000 Å.—R. J. Archer. (*Phys. Rev.*, 15th April 1958, Vol. 110, No. 2, pp. 354–358.) The optical constants, obtained from the ellipticity of reflected polarized light, show two maxima, and are consistent with the Kramers-Kronig relations.

537.311.33 : 546.289 : 537.32 : 538.63 **2794**

The Magnetic Variation of Thermoelectric Power of Ge Single Crystals at Low Temperatures.—J. Erdmann, H. Schultz & J. Appel. (*Z. Naturf.*, Feb. 1957, Vol. 12a, No. 2, pp. 171–174.) Results of measurements on pure *n*-type specimens are reported and briefly discussed. The changes of thermoelectric power as a function of magnetic field strength in the range 0–17 kG are plotted for 22·9°, 59·0° and 82·4°K. See also 2796 of 1956 (Lautz & Ruppel).

537.311.33 : 546.47-31 : 546.824-31 **2795**

Nonlinear $ZnO - TiO_2$ Semiconductors. — Kh. S. Valeev & M. D. Mashkovich. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1649–1651.) The electrical properties of $2ZnO–TiO_2$ are examined for concentrations of up to 40 % of TiO_2. Graphs are included which show, as a function of concentration, (*a*) the specific resistance, measured with constant potential of 16 V, and (*b*) the temperature coefficient of resistance in the interval 20–120°C. Up to 30 % of TiO_2, *n*-type conductivity was observed and above this, *p*-type conductivity.

537.311.33 : 546.49.241 : 538.63 **2796**

Nernst - Ettingshausen Effect in Mercuric Telluride.—I. M. Tsidil'kovskiĭ. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1744–1752.) Experiments carried out in the range 125–600°K showed that the temperature dependence of the Nernst-Ettingshausen effect could be explained by thermomagnetic theory. The temperature dependence of the electron and hole mobility was estimated. A quantitative confirmation of the theory was obtained in strong magnetic fields with equal current-carrier concentrations. Depending on the purity and polycrystalline content, HgTe at room temperature had a maximum electron mobility of $1 \cdot 15 \times 10^4$ cm^2/V. sec.

537.311.33 : 546.681.19 **2797**

Electron Bombardment of *p-n* Barrier Layers in GaAs.—H. Pfister. (*Z. Naturf.*, March 1957, Vol. 12a, No. 3, pp. 217–222.) The changes of e.m.f. and short-circuit current due to electron bombardment at different intensities are investigated. These effects can be used for the measurement of electron beam intensity up to 315 keV before radiation damage sets in.

537.311.33 : 546.682.19 : 538.63 **2798**

The Magnetic Resistance Change in InAs.—H. Weiss. (*Z. Naturf.*, Jan. 1957, Vol. 12a, No. 1, p. 80.) The resistance of a Corbino disk of InAs increased 5·5-fold in a field of 10 000 G. For a long single-crystal rod the resistance change amounted to 2·4 % at 10 000 G. For tests on InSb see 143 of 1955 (Weiss & Welker).

537.311.33 : 546.682.86 : 537.312.8 **2799**

The Transverse Magnetoresistance Effect in Indium Antimonide.—C. H. Champness. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 201–218.) "Measurements have been made on a wide range of nearly single crystal specimens at different temperatures. While most of the results can be explained from the measured carrier concentrations and Hall mobilities assuming lattice and impurity scattering, it is found that the magnetoresistance at liquid air temperature is larger than expected."

537.311.33 : 546.682.86 : 539.23 **2800**

Preparation and some Electrical Properties of Thin Films of InSb.—C. Paparoditis. (*C. R. Acad. Sci., Paris*, 28th Oct. 1957, Vol. 245, No. 18, pp. 1526–1528.) Films of *n*-type material 200 – 1 700 Å thick were deposited on mica by evaporation in vacuo followed by reheating in argon. Hall effect and resistivity were measured and the structure of some samples was studied by X rays and by electron microscope. The mean free path in samples reheated to 280°C was found to be much less than that in the bulk material and also less than the thickness of the samples.

537.311.33 : 546.682.86 : 539.23 **2801**

Theoretical Results relating to Thin Films of Indium Antimonide. — A. Colombani & J. Launay. (*C. R. Acad. Sci., Paris*, 4th Nov. 1957, Vol. 245, No. 19, pp. 1607–1608.) For previous work see 2470 and 2471 of August.

537.311.33 : 546.873.241 : 537.323 **2802**

Thermoelectric Properties of Bismuth Telluride and its Alloys.—D. A. Wright. (*Nature, Lond.*, 22nd March 1958, Vol. 181, No. 4612, p. 834.) Progress since 1954 [see 1465 of 1955 (Goldsmid & Douglas)] in the production of pure *p*- and *n*-type Bi_2Te_3 is reported. A maximum temperature difference of 65°C, at a mean temperature of 17°C, and Hall electron mobilities as high as 300 cm^2/V.sec have been obtained with these materials, and a maximum temperature difference of 80°C at the same mean temperature has been obtained by combinations of alloys of analogous materials.

537.311.33 : 548.5 **2803**

Oriented Growth and Definition of Medium-Angle Semiconductor Bicrystals.—H. F. Mataré & H. A. R. Wegener. (*Z. Phys.*, 22nd July 1957, Vol. 148, No. 5, pp. 631–645. In English.) The physical and metallurgical procedure in the preparation of semiconductor bicrystals is described, methods of seed orientation are given and precision limits are discussed. High-precision boundaries with tilt angle between 1° and 25° can be grown; a number of microphotographs of grown grain boundary structures are shown.

537.311.33 : 621.314.7 **2804**

The Status of Transistor Research in Compound Semiconductors. — D. A. Jenny. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 959–968.) The most promising semiconducting materials likely to compete with Ge and Si in transistor fabrication are among the group III–V and group IV–IV compounds. On the basis of performance at high temperatures and high frequencies, it is shown that, of compounds evaluated to date, GaAs and InP are the most promising, while AlSb, GaP and SiC will be useful at higher temperatures, at the cost of h.f. performance. Two new methods of junction preparation are described, namely surface diffusion and the formation of a wide-gap junction, which can be used to increase emitter efficiency. InP and GaAs transistors have been demonstrated, and electron lifetimes in the material estimated from measurements on them.

537.311.33+535.215] : 621.396.822 **2805**

Noise in Semiconductors and Photoconductors.—K. M. van Vliet. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1004–1018.) A survey of theory and experimental results relating to generation-recombination noise in semiconductors and photoconductors, $1/f$ noise in single crystals, and modulation noise in granular materials. 107 references.

538.221 **2806**

The Dependence of the Coercive Force on Dislocation Density.—Z. Málek. (*Z. angew. Phys.*, June 1957, Vol. 9, No. 6, pp. 279–280.) Critical assessment of Kersten's theory (1825 of 1957) in the light of recent measurements, followed by Kersten's comment (*ibid.*, pp. 280–281).

538.221 **2807**

The Effect of Dislocations on the Initial Permeability of Nickel in the Recrystallized and in the Plastically Deformed State.—M. Kersten. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 337–344.) Further discussion of the theory outlined in 1825 of 1957 with reference to experimental findings.

538.221 **2808**

Magnetization Curves of Tubular Specimens of Nickel under Torsion and Pressure.—K. Strnat. (*Z. Naturf.*, Jan. 1957, Vol. 12a, No. 1, pp. 76–79.)

538.211 : 538.24 **2809**

Radiation Magnetization. — C. H. Becker. (*Z. Phys.*, 8th May 1957, Vol. 148, No. 3, pp. 391–401. In English.) "Radiation magnetization represents a new electron induction effect within ferrites resulting in real permanent magnetization, dynamically induced and detected by means of microwaves."

538.221 : 621.318.134 **2810**

Domain Patterns on a Single Crystal of Manganese Ferrite.—L. F. Bates, D. J. Craik & P. M. Griffiths. (*Proc. phys. Soc.*, 1st May 1958, Vol. 71, No. 461, pp. 789–796.) Domain patterns on a cleaved surface were studied by optical and electron microscopy, using a colloid film technique. Separate fine and coarse patterns were observed, which are believed to represent independent domain structures, due respectively to internal domain structure and to a surface effect produced by cleavage.

538.221 : 621.318.134 **2811**

Magnetic and Crystallographic Properties of Substituted Yttrium-Iron Garnet, $3Y_2O_3 . xM_2O_3 . (5-x)Fe_2O_3$.—M. A. Gilleo & S. Geller. (*Phys. Rev.*, 1st April 1958, Vol. 110, No. 1, pp. 73–78.) A solid solution of Y-Fe garnet with Y-Ga or Y-Al garnet was formed over the entire range of composition, the lattice-constant variation being nearly linear. The trivalent ions, Sc^{3+}, In^{3+} and Cr^{3+} were substituted for Fe^{3+} to a limited extent. The effect of the substitutions on the magnetic moment and Curie temperature is described.

538.221 : 621.318.134 **2812**

Interpretation of Variations with Temperature of the Ferrimagnetic Resonance Fields in Gadolinium Garnet.—J. Paulevé. (*C. R. Acad. Sci., Paris*, 4th Nov. 1957, Vol. 245, No. 19, pp. 1604–1607.) Results of recent experimental work on pure material [see 538 of February (Calhoun et al.)] supplement previous work (3214 of 1957) and show reasonable agreement with theoretical hyperbolic curves.

538.221 : 621.318.134 : 621.372.413 : 621.317.411 **2813**

Frequency Shifts in Cavities with Longitudinally Magnetized Small Ferrite Disks.—H. Seidel & H. Boyet. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 637–655.) The tensor permeability components of a magnetized ferrite can be found from the frequency shift produced by the ferrite in a resonant cavity. Mathematical expressions are found for the permeability tensor for any natural mode in a cavity. Examples are given for various cavities.

538.614 : 539.23 : 621.317.44 **2814**

Investigation of the Magnetic Properties of Vapour-Deposited Iron Films by means of the Faraday Effect.—L. Reimer. (*Z. Phys.*, 22nd June 1957, Vol. 148, No. 4, pp. 527–532.) Note on the method outlined in 3592 of 1957 and further experimental results.

538.632 **2815**

Hall Coefficient of Technically Pure Metals from 80°K to 800°K.—V. Frank. (*Appl. sci. Res.*, 1957, Vol. B6, No. 5, pp. 379–387 & 1958, Vol. B7, No. 1, pp. 41–51.)

Part 1—Results for Cu, Ag, Au, Pd and Pt.

Part 2—Results for Zr, W, Mo, Ta, Nb and Al. Survey of Results for the 4*d* and 5*d* Transition Group of Metals.

621.315.61 **2816**

The Electrical Conductivity of Solid Organic Insulators.—N. Riehl. (*Ann. Phys., Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 93–128.)

621.315.61 : 537.221 **2817**

Triboelectricity and Electron Traps in Insulating Materials: some Correlations.—E. Fukada & J. F. Fowler. (*Nature, Lond.*, 8th March 1958, Vol. 181, No. 4610, pp. 693–694.) Some correlation is found between triboelectric properties and the distribution of electron traps as determined by a study of X-ray-induced conductivity [see 848 of 1957 (Fowler)].

621.315.61 : 538.566.029.6 **2818**

Low-Reflection Dielectric Walls for Microwaves.—H. Meinke. (*Nachrichtentech. Z.*, Nov. 1957, Vol. 10, No. 11, pp. 551–558.) Methods of obtaining a small reflection coefficient over a narrow as well as a wide band of frequencies are reviewed. The properties of dielectric media, particularly those containing graphite and iron powder, are discussed. 28 references.

621.315.612.6 : 537.226 **2819**

Preparation and Investigation of Dielectric Properties of a Group of Glasses with Increased Permittivity.—G. I. Skanavi & A. M. Kashtanova. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1770–1777.) Three groups of glasses were investigated: boron, tellurite and boron lead glasses. Best results were obtained with 24% TiO_2, 54·4% PbO glasses giving a dielectric constant of 35 and breakdown voltage of 2×10^6 V/cm. Results are shown in graphical and tabular form.

621.315.612.6 : 537.226.31 **2820**

Nature of Dielectric Losses in Sodium Aluminosilicate Glasses.—V. I. Ioffe. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1454–1461.) An investigation of silicate and borate glasses in the temperature range 20–290°K at 230, 1 000 and 1 700 kc/s. Dielectric losses were observed to decrease with increasing oxygen content. The results are presented graphically.

MATHEMATICS

512.831 : 518.2 **2821**

Tables for Diagonalizing Second-Order Matrices.—R. E. Trees & C. D. Coleman. (*J. Res. nat. Bur. Stand.*, March 1958, Vol. 60, No. 3, RP2838, pp. 201–214.) "Sets of tables are given to facilitate the evaluation of the eigenvalues and eigenvectors of second-order matrices."

512.831 : 621.372.6 **2822**

Matrices all of whose Elements and Subdeterminants are 1, –1 or 0.—I. Cederbaum. (*J. Math. Phys.*, Jan. 1958, Vol. 36, No. 4, pp. 351–361.) See also 2009 of July.

517.512.2 **2823**

The Relationship of Physical Applications of Fourier Transforms in Various Fields of Wave Theory and Circuitry.—E. F. Bolinder. (*Trans. Inst. Radio Engrs*, April 1957, Vol. MTT-5, No. 2, pp. 153–158. Abstract, *Proc. Inst. Radio Engrs*, July 1957, Vol. 45, No. 7, p. 1036.)

517.512.4 : 518.2 **2824**

Table of First 700 Zeros of Bessel Functions—$J_\iota(x)$ and $J'_\iota(x)$.—C. L. Beattie. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 689–697.) The zeros of the Bessel functions and Bessel function derivatives are identified by standard waveguide notation which also serves as a code for more general mathematical applications.

518.6 : 621.372.2 **2825**

Application to Electrodynamic Problems of an Approximation of a Logarithm by a Power Law.—W. Rehwald & O. Zinke. (*Arch. elekt. Übertragung*, Oct. 1957, Vol. 11, No. 10, pp. 397–402.) The function $\log_e x$ is approximated by a power law of the form Ax^c. The method is applied to the problem of e.m. wave propagation along a bare wire of finite conductivity (Sommerfeld line) and along a helical conductor.

MEASUREMENTS AND TEST GEAR

529.78 : 621.374 : 621.314.7 **2826**

Transistor Chopper drives Accurate Clock.—R. H. Williams. (*Electronics*, 23rd May 1958, Vol. 31, No. 21, pp. 64–65.) A 400-c/s synchronous clock motor is driven from an 800-c/s transistorized crystal oscillator using a voltage-doubling chopper circuit and a 28-V d.c. power supply.

531.76 : 621.373.43 **2827**

The Measurement of Periodically Recurring Short Time Intervals.—R. Gerharz. (*Z. angew. Phys.*, June 1957, Vol. 9, No. 6, pp. 282–286.) The use of pulse generators with high repetition frequency (see 2065 of 1957) for calibration purposes is discussed. By using two identical pulse generators to produce a beat frequency curve on a c.r.o. screen, time interval measurements with a resolution of at least 10^{-11} s have been made.

621.317.31.024.42 **2828**

Method of Measuring and Recording 'Ultra Small' Currents.—L. L. Dekabrun. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1578–1583.) Electrometer valve circuits are described by means of which currents up to 10^{-14} A are measured. This is accomplished by passing the current through resistances of 10^{10}–10^{12} Ω and by measuring the voltage drop. The instrument has a good stability and an accuracy within ±10%.

621.317.32 : 621.396.822 **2829**

Noise Voltage Measurements on Low-Impedance Circuit Elements by means of a Valve Amplifier with Input Transformer.—W. Nonnenmacher. (*Nachrichtentech. Z.*, Nov. 1957, Vol. 10, No. 11, pp. 559–563.) Equivalent noise impedances of less than 1 Ω can be determined at low and medium frequencies by the method described. Additional noise due to the input transformer can be eliminated.

621.317.32.024 : 621.375.9 : 537.312.62 **2830**

A Sensitive Superconducting 'Chopper' Amplifier.—De Vroomen & Van Baarle. (See 2675.)

621.317.331 **2831**

Measurement of Sheet Resistivities with the Four-Point Probe.—F. M. Smits. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 711–718.) See also 1502 of 1954 (Valdes).

621.317.331.028.3 **2832**

The Measurement of High-Value Resistances.—J. K. Wood. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 374–377.) Several methods using standard laboratory equipment are described together with sources of error and ways of reducing them.

621.317.34 : 621.397.5 **2833**

Measurement of the Reflection Coefficient of Television Lines and Equipment.—E. Thinius. (*Nachrichtentech. Z.*, Nov. 1957, Vol. 10, No. 11, pp. 548–550.) Using a frequency-sweep oscillator a response curve is obtained on a c.r.o. screen, from which the reflection coefficient and phase angle can be determined.

621.317.361 **2834**

Frequency Measurement by Time Delay of a Signal.—A. I. Danilenko. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 67–72.) Two voltages are applied to a summation device, one from the oscillator under investigation and the other from the same oscillator but delayed in time. The amplitude of the resulting voltage will depend on frequency. The method might be used to determine the characteristics of delay lines.

621.317.411 : 621.372.413 : 538.221 : 621.318.134 **2835**

Frequency Shifts in Cavities with Longitudinally Magnetized Small Ferrite Disks.—Seidel & Boyet. (See 2813.)

621.317.61 : 621.372.8 : 621.376.232 **2836**

A Phase-Sensitive Detector.—Beauchamp. (See 2687.)

621.317.71 : 537.228.1 **2837**

Simple Apparatus for the Direct Determination of the Charge Output of Piezoelectric Materials at High Forces.—D. S. Schwartz. (*Rev. sci. Instrum.*, April 1958, Vol. 29, No. 4, pp. 321–323.) Compressions varying from 50 to 15 000 lb are obtained with hydraulic vices. The charge is measured using a Miller integrator circuit which presents a very low impedance to the sample.

621.317.733 : 621.372.412.029.62 **2838**

Plug-in Bridge Checks V.H.F. Quartz Crystals.—D. W. Robertson. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 82–85.)

621.317.755 : 537.52 **2839**

A Multiple-Beam Oscilloscope for the Study of High-Voltage Transient Discharges.—K. G. Beauchamp. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 358–365.) High-voltage transient discharge phenomena are studied by means of trace expansion and selective brightening. The effects of corona and main discharge can be separated in the instrument. The complete design is given.

621.317.755 : 537.52 **2840**

The Synchronized Plasmograph.—R. Ledrus, M. Hoyaux, A. Vanavermaete & P. Gans. (*Rev. gén. Élect.*, Oct. 1957, Vol. 66, No. 10, pp. 513–521.) Refinement of an instrument described earlier [532 of 1956 (Ledrus et al.)].

621.317.763.089.6 : 621.372.414.029.63 **2841**

Measuring Decimetric Wavelengths.—H. B. Dent. (*Wireless World*, July 1958, Vol. 64, No. 7, pp. 319–323.) Lecher wires are used to calibrate an absorption wavemeter covering the range 450–750 Mc/s.

621.317.799 : [621.314.63 + 621.385.029.6 **2842**

Diode and Klystron Test Set for the 3.2-cm Region.—W. Otto. (*NachrTech.*, Oct. & Nov. 1957, Vol. 7, Nos. 10 & 11, pp. 454–460 & 502–506.) The equipment described is designed for production tests on Si diodes and reflex klystrons used as mixers and local oscillators respectively.

621.317.799 : 621.385.1 **2843**

High-Speed Tester Checks Tubes in Groups.—E. S. Gordon. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 76–78.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.7 : 681.142 **2844**

The Synchro Resolver as a Shaft Position Transducer.—M. B. Wood. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 366–370.) The resolver is part of a phase-shifting unit with 400-c/s sinusoidal input. The difference in time-phase of the output zero crossover in the reference and measuring channels is measured by a period counter. The relation between shaft and crossover pulse position is linear to within ±0·1 %.

551.508.7 : 621.316.825 **2845**

Captive Balloon Refractovariometer.—A. L. Crozier. (*Rev. sci. Instrum.*, April 1958, Vol. 29, No. 4, pp. 276–279.) A system is described for measuring and recording rapid (3-c/s) fluctuations of refractive index, humidity and temperature using fast-response thermistor beads. Measurements can be obtained up to 5 000 ft above the measuring site.

621.383.4 : 612.84 **2846**

Retinal Type of Photovoltaic Cell.—I. Levin. (*Nature, Lond.*, 22nd March 1958, Vol. 181, No. 4612, p. 832.) A multi-electrode photocell analogous to the retina and an amplifier unit analogous to the nervous system of the eye are described.

621.383.8 : 537.311.33 : 535.34.096 **2847**

A New Thermal Image Converter.—W. R. Harding, C. Hilsum & D. C. Northrop. (*Nature, Lond.*, 8th March 1958, Vol. 181, No. 4610, pp. 691–692.) A simple image converter can be made based on the temperature dependence of the absorption threshold in a semiconductor. In the example described a self-supporting film of Se is metallized on one side and mounted in a vacuum at the focus of a parabolic mirror. The film is viewed by transmitted light from a sodium lamp. Variations in temperature or emissivity in the scene focused on the semiconductor appear as differences in transmitted intensity.

621.384.6 **2848**

Focusing of High-Velocity Electrons in Linear Electron Accelerators.—K. N. Stepanov & A. A. Sharshanov. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1863–1869.) The radial movement of high-velocity relativistic electrons is examined and equations are derived.

621.384.6 **2849**

The Focusing of Beams of Charged Particles by High-Frequency Fields.—M. A. Miller. (*Dokl. Ak. Nauk S.S.S.R.*, 21st March 1958, Vol. 119, No. 3, pp. 478–480.) Mathematical analysis applied to an example of the focusing of a rectilinear beam in the field of symmetry of a TE_{10}-mode wave propagated in a circular waveguide with ideally conducting walls.

621.384.6 + 621.385.029.6] : 621.372.2 **2850**

The Propagation of Slow Waves.—Dain. (See 2933.)

621.384.612 **2851**

Resonance Perturbation of Synchrotron Oscillations in Particle Accelerators.—I. S. Danilkin & M. S. Rabinovich. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1558–1570.) The influence of the resonance harmonic disturbances on the phase shift and the intensity of the beam of accelerated particles is investigated. A synchrophasotron of 10 kMeV is used.

621.384.613 **2852**

Investigation of Electron Capture Processes in a Betatron.—Yu. S. Korobochko. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1603–1605.) The average number of revolutions of an electron in a 15-MeV betatron was estimated to be 4–5; the current circulating in the orbit during the trapping period was calculated and the trapping efficiency assessed.

621.384.7 **2853**

Theory of an Electric High-Frequency Mass Spectrometer.—G. Falk & F. Schwering. (*Z. angew. Phys.*, June 1957, Vol. 9, No. 6, pp. 272–275.) A static and a high-frequency field are superimposed inside a cylindrical capacitor through which the ion beam travels, and an absorption-type spectrum is produced.

621.385 : 621.317.7 **2854**

An Ion-Optical Bench for the Study of Four-Pole Magnetic Lenses and of Magnetic Deflectors for Particle Accelerators.—A. Septier. (*C. R. Acad. Sci., Paris*, 21st Oct. 1957, Vol. 245, No. 17, pp. 1406–1409.) A purely electrostatic system operating with an electron beam or with ions of any mass is described. Aberrations in the magnetic system under test are detected using a fluorescent screen.

621.385.833 **2855**

Measuring with the Electron.—L. Marton. (*J. sci. industr. Res.*, Oct. 1957, Vol. 16A, No. 10, pp. 429–439.) A review of electron-optical techniques in microscopy, field mapping and interferometry.

621.385.833 **2856**

Addition to the Theory of Aberration of an Electron-Optical Focusing System with a Curvilinear Axis.—Yu. V. Vandakurov. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1850–1862.) From the trajectory of particles, equations are derived by means of which the second- and third-order geometrical aberration and first- and second-order chromatic aberration for arbitrary electric and magnetic fields are calculated. Formulae are also given for the dispersion of particles corresponding to their velocities and masses. Conditions are determined for which second-order geometrical aberration does not occur.

621.385.833 **2857**

Experimental Investigation of Aperture Aberrations of a System of Two Four-Pole Magnetic Lenses.—A. Septier. (*C. R. Acad. Sci., Paris*, 25th Nov. 1957, Vol. 245, No. 22, pp. 1905–1908.)

621.385.833 **2858**

Contrast in the Electron-Microscope Image.—R. C. Valentine. (*Nature, Lond.*, 22nd March 1958, Vol. 181, No. 4612, pp. 832–833.) Three methods of comparing contrast are discussed and evaluated numerically.

621.385.833 **2859**

Simultaneous Evaporation of Platinum and Carbon for Possible Use in High-Resolution Shadow-Casting for the Electron Microscope.—D. E. Bradley. (*Nature, Lond.*, 29th March 1958, Vol. 181, No. 4613, pp. 875–877.)

621.387.422 **2860**

The Discharge Process in Proportional Counter Tubes.—C. Keck. (*Z. angew. Phys.*, June 1957, Vol. 9, No. 6, pp. 286–292.)

621.387.462 **2861**

Semiconductor Device for γ-Ray Indication.—S. M. Ryvkin, A. P. Bogomazov, B. M. Konovalenko & O. A. Matveev. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1601–1602.) Best results were obtained with CdS and CdSe specimens, the conductivity of which changes sharply when irradiated.

621.387.462 : 546.289 **2862**

The Use at Low Temperatures of *n-p*-Type Germanium α-Particle Counters.—A. V. Aĭrapetyants, A. V. Kogan, N. M. Reĭnov, S. M. Ryvkin & I. A. Sokolov. (*Zh. tekh. Fiz.*, July 1957, Vol. 27. No. 7, pp. 1599–1600.)

621.398 : 621.391 **2863**

Some Problems of Telemetry in the Light of Communication Theory.—V. Pollak. (*Hochfrequenztech. u. Elektroakust.*, July 1957, Vol. 66, No. 1, pp. 19–25.) Communication theory is applied to the analysis of telemetry systems used in power transmission.

621.398 : 621.391 **2864**

Systems Engineering a P.D.M./F.M. Telemetry System.—F. J. Enge. (*Electronic Ind.*, March 1958, Vol. 17, No. 3, pp. 80–81 . . 128.) The basic principles involved in a p.w.m./f.m. telemetry system are discussed in relation to information theory.

PROPAGATION OF WAVES

621.396.11 : 551.510.52 **2865**

Tropospheric Scatter Propagation—a Summary of Recent Progress.—H. Staras. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 3–18.) A short history and an explanation in terms of physical concepts are given. In discussing bandwidth capabilities, correlation distances, aerial-to-medium coupling loss and angular diversity, published theoretical work is compared with recent experimental data.

621.396.11 : 551.510.52 **2866**

The Effect of Refraction on the Scatter Propagation of Ultra Short Waves in the Troposphere.—D. M. Vysokovskiĭ. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 30–36.) Approximate formulae are derived for determining the effect of such factors as the lowering of the height of the scattering region, the variation of the angle of scattering and the variation of the extent of the scattering region. The variation of power at the point of reception in the presence of refraction is calculated. The results obtained are compared with the theoretical and experimental data on the scattering of radio waves at turbulent inhomogeneities.

621.396.11 : 551.510.535 **2867**

Observations on Back-Scatter Echoes in Long-Distance Short-Wave Transmissions.—H. U. Widdel. (*Arch. elekt. Übertragung*, Nov. 1957, Vol. 11, No. 11, pp. 429–439.) The influence of aerial radiation patterns on the range and complexity of back-scatter echo observations was investigated using 1-ms pulses transmitted at a rate of 50/sec at frequencies of 5 980, 9 640, 15 275 and 17 845 kc/s. Two rhombics and one 250-m long-wire aerial 8 km from the transmitter were used for reception.

621.396.11 : 551.510.535 **2868**

The Role of Ionospheric-Layer Tilts in Long-Range High-Frequency Radio Propagation.—S. Stein. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 217–241.) The mechanism of reflection of h.f. radio waves from a nonspherical layer is studied by means of simplified models. Rays emitted from a ground transmitter at low angles of elevation and reflected from a suitably tilted F layer will miss the earth and illuminate a second area of the layer directly. If sufficient ionization exists in this second area, the ray will be propagated around the curve of the earth. A ray may be reflected several times and finally encounter a tilt so oriented that the surface of the earth is illuminated. Anticipated fading, m.u.f. and field-strength characteristics for these '^{n}F' modes are discussed. Many different ^{n}F modes and combinations of ^{n}F and conventional modes have been observed by back-scatter sounding between 12 Mc/s and 30 Mc/s, and they are described.

621.396.11.029.45 : 551.510.535 **2869**

An Extension to the Mode Theory of V.L.F. Ionospheric Propagation.—J. R. Wait. (*J. geophys. Res.*, March 1958, Vol. 63, No. 1, pp. 125–135.) The waveguide mode theory of v.l.f. propagation for a sharply-bounded homogeneous ionosphere is refined to include stratification at the lower edge of the ionosphere. The numerical results for a two-layer model are discussed in detail.

621.396.11.029.62 **2870**

Further Results of Tropospheric Drift Observations on the Question of the Mechanism of Long-Range Propagation at Metre Wavelengths.—L. Klinker. (*Z. Met.*, Oct./Nov. 1957, Vol. 11, Nos. 10/11, pp. 339–344.) 130 of the 160 measurements of tropospheric drift made in the course of one year's observations at Kühlungsborn (see 3279 of 1957) were compared with simultaneous aerological measurements made at Copenhagen. An analysis of the results appears to confirm earlier conclusions (see 587 of February).

RECEPTION

621.396.62 : 621.314.7 : 621.311.69 **2871**

Radio Waves power Transistor Circuits.—L. R. Crump. (*Electronics*, 9th May 1958, Vol. 31, No. 19, pp. 63–65.) Unpowered transistorized receivers are actuated by switching on a transmitter. The outputs from the receiving aerials need to be about 1 mW.

621.396.62.029.63/.64 : 621.396.822 **2872**

Noise Performance of a Three-Stage Microwave Receiver.—H. V. Shurmer. (*Electronic Radio Engr*, July 1958, Vol. 35, No. 7, pp. 271–274.) The analysis of the noise performance of microwave receivers is extended to the case where a crystal valve is preceded by a stage of r.f. amplification, such as a travelling-wave tube.

621.396.621.54 : 621.376.33 **2873**

Bandwidth, Number of Stages and Distortion Factor in V.H.F. F.M. Receivers.—E. G. Woschni. (*NachrTech.*, Oct. 1957, Vol. 7, No. 10, pp. 441–447.) The interdependence of selectivity, distortion factor, number of i.f. stages, and bandwidth is investigated and curves showing these relations are given.

621.396.812.3 **2874**

**Distribution of the Duration of Fades in Radio Transmission: Gaussian

Noise Model.—S. O. Rice. (*Bell Syst. tech. J.*, May 1958, Vol. 37, No. 3, pp. 581–635.) The fluctuations of a received radio signal due to fading are assumed to behave like the envelope of narrow-band Gaussian noise. Estimates of the distribution of the fade lengths for various depths of fades are given, and relations which may be useful in analysing fading data are derived. A similar problem involving the separation of the intercepts of the noise current itself, instead of its envelope, is also discussed.

621.396.82 2875

Experimental Investigation of the Limitation of Pulse Interference by Varying the Spectra and the Threshold of Limiting.—A. A. Gorbachev. (*Radiotekhnika, Mosk.*, June 1957, Vol. 12, No. 6, pp. 64–68.) In this method the signal and interference spectra are transformed by two linear converters to make use of certain features of the spectra, and the threshold of the limiter is made to follow automatically the level of the signal. Experiments have shown that good suppression is obtained of pulses of duration up to several milliseconds.

STATIONS AND COMMUNICATION SYSTEMS

621.376 : 621.395.665.1 2876

Automatic Speech Amplitude Control.—L. R. Battersby. (*Electronics*, 23rd May 1958, Vol. 31, No. 21, pp. 71–73.) A simple differentiating network changes the energy distribution of speech so that input amplitude variations of up to 35 dB over the range 300–3 000 c/s are reduced to output variations of about 1 dB without serious loss of intelligibility. Field tests with portable f.m. transceivers show that a marked increase in a.f. output can then be obtained when the r.f. signal/noise ratio is poor.

621.391 2877

On the Relation between the Rate of Transmission of Information and the Freedom from Interference in a Communication System.—E. L. Blokh. (*Radiotekhnika, Mosk.*, June 1957, Vol. 12, No. 6, pp. 3–14.) The relation is investigated with the aid of geometrical methods for codes corresponding to the simplest and the densest distributions of signal points.

621.391 : 621.396.82 2878

Statistical Properties of Signals and Interference in Two-Channel Phase Systems.—V. V. Tsvetnov. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 12–29.) Statistical properties of sinusoidal signals and Gaussian interference in two-channel phase systems are discussed and the law of the distribution of amplitude and phase differences is derived. The statistical characteristics are considered for the following cases: (*a*) instantaneous phase error and uncorrelated two-channel interference; (*b*) relatively strong signals.

621.391 : 621.396.822 2879

Zeros of Gaussian Noise.—G. M. White. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 722–729.) Results are reported of an investigation by an experimental technique for determining the statistics of the times a random function crosses through its average value. Close agreement was found with values calculated theoretically; others not amenable to calculation gave results intuitively expected. The equipment is briefly described.

621.396.712 2880

The New Transmitting Station at Lourenço Marques in Mozambique.—M. Dick. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 420–427.) The 3·2–22-Mc/s transmitter described can also be used for transmitting at any frequency in the medium-wave band.

621.396.712 : 621.396.677.3 2881

Long-Distance H.F. Broadcasting.—Bennington. (See 2637.)

SUBSIDIARY APPARATUS

621.3.013.78 2882

The Reflection and Screening Effects of Metallic Enclosures in a Plane Electromagnetic Wave.—H. Kaden. (*Arch. elekt. Übertragung*, Oct. 1957, Vol. 11, No. 10, pp. 403–415.) Three cases of reflection and screening action are investigated for wavelengths of the order of the dimensions of the enclosure: (*a*) a hollow cylinder with axis parallel to the electric field, (*b*) a hollow cylinder with axis parallel to the magnetic field, and (*c*) a hollow sphere. The increase of the reflection coefficient with rising frequencies, screen attenuation, and resonance effects are discussed. See also 578 of 1957.

621.311.69 : 539.16 2883

Nuclear Batteries.—E. K. Aschmoneit. (*Elektronik*, Oct. 1957, Vol. 6, No. 10, pp. 287–290.) Outline of principles of operation and applications. See also 1916 of 1957 (Milliron).

621.311.69 : 621.396.62 : 621.314.7 2884

Radio Waves power Transistor Circuits.—Crump. (See 2871.)

621.314.63 : 537.311.33 2885

Voltage/Current Characteristics of Semiconductor Power Rectifiers.—E. I. Rashba & A. I. Nosar'. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, pp. 1431–1445.) The distribution of the concentration of carriers of *p-i-n* and *p-n-n* types is examined. Monomolecular and bimolecular recombinations are considered. The dependence of the injection coefficient on the current and the *V/I* characteristics is calculated and the voltage drop in the bulk of the semiconductor is investigated.

621.314.63 : 546.28 2886

Medium-Power Silicon Rectifiers.—R. J. Andres & E. L. Steele. (*Electronic Ind.*, March 1958, Vol. 17, No. 3, pp. 62–65.) The construction, manufacturing techniques and electrical characteristics of the diode Type MN-14 of *p-i-n* structure are described.

621.316.722.1 : 621.314.7 2887

Stabilized Power Supplies using Transistors.—E. Baldinger & W. Czaja. (*Z. angew. Math. Phys.*, 25th Jan. 1958, Vol. 9, No. 1, pp. 1–25.) The operation of simple stabilizing circuits is analysed and methods of improvement by pre-amplification or compensation are discussed. Practical circuits which have low temperature coefficients are described.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 : 535.623 : 061.3 2888

International Symposium, Physical Problems of Colour Television, Paris, July 1957.—(*J. opt. Soc. Amer.*, Jan. 1958, Vol. 48, No. 1, pp. 73–74.) Proceedings of this conference are to be published by *Acta Electronica*, 23 rue du Retrait, Paris 20ᵉ, price $10.

621.397.61 2889

Television Transmitters.—V. Milliquet. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 438–446.) An installation comprising a 1·5-kW vision transmitter and a 300-W f.m. sound transmitter is described.

621.397.611.2 2890

Mechanism of Electronic Commutation in Television Tubes with Energy Storage.—Ya. A. Ryftin. (*Zh. tekh. Fiz.*, Aug. 1957, Vol. 27, No. 8, pp. 1870–1885.) A comparison of experimental and theoretical results shows that the main characteristics of a storage-type camera tube can be improved by increasing the relative line shift. This can be achieved by increasing the size of the target or by decreasing the effective radius of the scanning spot. Results are presented graphically.

621.397.611.2 2891

Some New Structure-Type Targets for the Vidicon—an Analysis of their Operation.—S. A. Ochs & P. K. Weimer. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 49–61.) Two types of target are discussed, (*a*) a 'lateral-flow' structure with photocurrents flowing parallel to the target plane, and (*b*) a 'bridge-type' structure with each picture element providing an internal closed circuit thus establishing a charge pattern independently of the scanning beam. Experimental structure-type targets capable of several hundred lines resolution are described but are more difficult to make than conventional layer-type targets. Use can be made of sensitive photoconductors with resistivity too low for the standard vidicon.

621.397.62 2892

Assessment of the Quality of Horizontal Synchronization in Television Receivers.—H. Lutz. (*Arch. elekt. Übertragung*, Nov. 1957, Vol. 11, No. 11, pp. 461–470.) Parameters governing the operation of a phase-comparison system of horizontal synchronization are derived with reference to the operating point on the control-circuit characteristic. The noise characteristics of the system and its response to a sine-wave or pulse disturbance are calculated and a circuit for measuring these properties is described. A full evaluation of pull-in and locking properties is only possible if the nonlinear part of the characteristic of the control circuit is known. Methods of plotting the unstable portion of the characteristic and of analysing the pull-in process are described.

621.397.62 : 621.385.832 **2893**

A Television Picture Tube with Increased Effective Perveance for Cathode Modulation.—W. F. Niklas, C. S. Szegho & J. Wimpffen. (*J. Telev. Soc.*, Jan./March 1958, Vol. 8, No. 9, pp. 368–375.) The tube characteristics associated with grid and cathode modulation are examined. The requirements for a new picture tube designed for cathode modulation are listed. Special features of the electron gun are described.

621.397.62 : 621.396.662 **2894**

Performance of Television Turret Tuners.—K. H. Smith. (*J. Telev. Soc.*, Jan./March 1958, Vol. 8, No. 9, pp. 377–390.) A detailed description of experimental work on circuits and valve design for improving band-III noise performance and gain.

621.397.62 : 621.396.662 **2895**

A Comparison of Turret-Type and Switch-Type Television Tuners.—V. A. Jones. (*J. Brit. Instn Radio Engrs*, June 1958, Vol. 18, No. 6, pp. 346–347. Discussion, pp. 349–357.) A table comparing the characteristics of German, U.S. and British tuners is given.

621.397.62 : 621.396.662.002.2 **2896**

Mass-Production Techniques for Television Tuners.—P. C. Ganderton. (*J. Brit. Instn Radio Engrs*, June 1958, Vol. 18, No. 6, pp. 331–340. Discussion, pp. 349–357.) Design details for Band I, II and III turret-type tuners, to ensure uniformity of performance and freedom from temperature effects. A system using 'wobbulated' signals for coil alignment is described.

621.397.62 : 621.396.662.002.2 **2897**

Some Aspects of Television Tuner Production.—S. H. Perry. (*J. Brit. Instn Radio Engrs*, June 1958, Vol. 18, No. 6, pp. 341–345. Discussion, pp. 349–357.) A statistical method for analysing production variations in a particular type of tuner is described, together with manufacturing methods used to achieve the required accuracy and standardized performance.

621.397.62 : 621.396.665 **2898**

A.G.C. Circuits in Television Receivers.—S. N. Doherty & P. L. Mothersole. (*J. Telev. Soc.*, Jan./March 1958, Vol. 8, No. 9, pp. 350–367.) The problem of obtaining a wide range of a.g.c. is discussed and a range of 80 dB is shown to be possible. The development is described of a simple high-gain gate circuit which is free from the faults of existing circuits.

621.397.62 : 621.396.665 **2899**

Automatic-Gain-Control Circuits in Television Receivers for Negative-Modulation Systems.—P. L. Mothersole. (*J. Brit. Instn Radio Engrs*, May 1958, Vol. 18, No. 5, pp. 307–316.) A simple gate circuit is described in which the a.g.c potential is dependent on the black level of the signal and is completely independent of the line timebase frequency. The circuit also has the advantages of (*a*) direct coupling, (*b*) high input impedance, (*c*) relatively few valves, and (*d*) synchronizing-pulse cancellation.

TRANSMISSION

621.396.61 **2900**

Peculiar Features in the Design of Radio Transmitters in which the Anode Circuits are Fed from Sources with a High Internal Impedance.—N. I. Shteĭn. (*Radiotekhnika, Mosk.*, May 1957, Vol. 12, No. 5, pp. 48–53.) The optimum utilization of d.c. sources with a relatively high internal impedance, such as rotary converters and thermionic rectifiers, is discussed. Recommendations are given for the design of valve oscillators to ensure the highest output.

621.396.61 : 621.396.822.1 **2901**

Combination Frequencies and Crosstalk in Transmitters for Class-C Operation.—L. Leng. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 433–437.) Theoretical investigation of the conditions under which crosstalk arises in stations where several transmitters are housed together.

621.396.71 **2902**

New Brown Boveri Transmitters all over the World.—S. Pedersen. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 400–409.)

621.396.712 **2903**

Modern Regional Broadcast Transmitters.—W. Klein. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 416–420.) Description of a 25-kW medium-wave transmitter.

621.396.712 : 621.376.3 **2904**

Frequency-Modulated V.H.F. Broadcast Transmitters.—K. Lutz. (*Brown Boveri Rev.*, Oct. 1957, Vol. 44, No. 10, pp. 409–415.) Brief description of a new series of transmitters available for 1, 3 and 10 kW outputs.

VALVES AND THERMIONICS

621.314.63 **2905**

The Potential of Semiconductor Diodes in High-Frequency Communications.—A. Uhlir, Jr. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1099–1115.) The construction, properties and applications of three types of diode are described. Graded *p-n* junctions made by solid-state diffusion are low-loss nonlinear capacitors which can be used for low-noise amplifiers, amplifying frequency converters, harmonic and subharmonic generators, switches, limiters and voltage-tuned passive circuits. Point-contact diodes (or *p-n* junctions at lower frequencies) are nonlinear resistors, used as microwave rectifiers. *p-i-n* diodes have a high-frequency resistance which is dependent on the direct current, and so can be used as wide-band microwave switches or attenuators.

621.314.63 **2906**

The Evolution of the Theory for the Voltage/Current Characteristic of *P-N* Junctions.—J. L. Moll. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1076–1082.) The rectifying action is controlled essentially by the equilibrium densities, diffusion constants and recombination times of minority carriers. Low-level behaviour of Ge junctions at room temperature is described theoretically on the basis of diffusion and recombination rates on either side of the barrier region. For Si, effects of carrier recombination and generation in the barrier region must be included. At high current densities, junctions depart from the ideal low-level rectifier law because of effects associated with majority-carrier modulation.

621.314.63 : 537.311.33 **2907**

The Time-Lag of Semiconductor Diodes in Pulse Operation and its Physical Interpretation.—W. Heinlein. (*Arch. elekt. Übertragung*, Oct. 1957, Vol. 11, No. 10, pp. 387–396.) An interpretation of the delay mechanism operative in pulse-controlled *p-n* junctions is confirmed by measurements on Ge crystal diodes with large diffusion regions.

621.314.63+621.385.029.6] **2908**
: 621.317.799

Diode and Klystron Test Set for the 3·2-cm Region.—Otto. (See 2842.)

621.314.63+621.314.7].002.2 **2909**

Outdiffusion as a Technique for the Production of Diodes and Transistors.—J. Halpern & R. H. Rediker. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1068–1076.) The diffusion of Sb out of Ge has been studied, and the use of this process is shown to be an easy way of producing *p-n* junctions in compensated *n*-type Ge. High-speed narrow-base diodes and high-speed *n-p-n* Ge transistors have been made by outdiffusion.

621.314.7+621.314.63 **2910**

Lumped Models of Transistors and Diodes.—J. G. Linvill. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1141–1152.) A lumped approximation is made at the beginning of analysis of a distributed system. This generally simplifies analysis, permits consideration of phenomena prohibitive to analysis on a differential basis and provides a close tie with the physical aspects involved. Lumped models are shown which can be used to approximate the properties of transistors and diodes over a wide range of conditions and applications.

621.314.7 **2911**

Transistors.—(*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 952–1300.) This issue commemorates the tenth anniversary of the invention of the transistor. Abstracts of some of the papers are given individually; titles of others are as follows:

(*a*) The Technological Impact of Transistors.—J. A. Morton & W. J. Pietenpol (pp. 955–959).

(*b*) Review of Other Semiconductor Devices.—S. J. Angello (pp. 968–973).

(*c*) Analogue Solution of Space-Charge Regions in Semiconductors.—L. J. Giacoletto (pp. 1083–1085).

(*d*) Germanium and Silicon Rectifiers.—H. W. Henkels (pp. 1086–1098).

(*e*) New Concepts in Microwave Mixer Diodes.—G. C. Messenger (pp. 1116–1121).

(*f*) Advances in the Understanding of the *P-N* Junction Triode.—R. L. Pritchard (pp. 1130–1141). 98 references.

(*g*) Construction and Electrical Properties of a Germanium Alloy-Diffused Transistor.—P. J. W. Jochems, O. W. Memelink & L. J. Tummers (pp. 1161–1165).

(*h*) Technology of Micro-alloy Diffused Transistors.—C. G. Thornton & J. B. Angell (pp. 1166–1176).

(*i*) Power Transistors.—M. A. Clark (pp. 1185–1204).

(*j*) Measurement of Transistor Thermal Resistance.—B. Reich (pp. 1204–1207).

(*k*) Measurement of Internal Temperature Rise of Transistors.—J. T. Nelson & J. E. Iwersen (pp. 1207–1208).

(*l*) A Five-Watt Ten-Megacycle Transistor.—J. T. Nelson, J. E. Iwersen & F. Keywell (pp. 1209–1215).

(*m*) The Effective Emitter Area of Power Transistors.—R. Emeis, A. Herlet & E. Spenke (pp. 1220–1229).

(*n*) Multiterminal *P-N-P-N* Switches.—R. W. Aldrich & N. Holonyak, Jr (pp. 1236–1239).

(*o*) The Application of Transistors to Computers.—R. A. Henle & J. L. Walsh (pp. 1240–1254).

(*p*) Application of Transistors in Communications Equipment.—D. D. Holmes (pp. 1255–1260).

(*q*) Transistor Monostable Multivibrators for Pulse Generation.—J. J. Suran (pp. 1260–1271).

(*r*) A Design Basis for Junction-Transistor Oscillator Circuits.—D. F. Page (pp. 1271–1280).

621.314.7 **2912**

Two-Dimensional Current Flow in Junction Transistors at High Frequencies.—R. L. Pritchard. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1152–1160.) At h.f. the distributed nature of the base region must be taken into account. In the usual equivalent circuit, the ohmic base resistance must be replaced in general by a complex base impedance. The effect of this modification upon circuit performance is discussed and transistor design considerations are outlined.

621.314.7 **2913**

Junction Transistor Short-Circuit Current Gain and Phase Determination.—D. E. Thomas & J. L. Moll. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1177–1184.) Analysis is given showing that the complete common-base and common-emitter current-gain magnitude and phase characteristics as a function of frequency can be determined from three amplitude measurements, namely the low-frequency current gain, the magnitude of the common-emitter current gain at a single frequency in the common-emitter cut-off region, and the common-base cut-off frequency. Equations are developed for determining the common-emitter and common-base short-circuit current gains from these three measurements.

621.314.7 **2914**

The Intrinsic-Barrier Transistor—How it Works.—J. M. Early. (*Bell Lab. Rec.*, March 1958, Vol. 36, No. 3, pp. 86–90.) Examination of the structural limitations of a three-layer transistor leads to an estimate of the ultimate capabilities of alloy and grown-junction types and to the development of a new type of transistor triode, *p-n-i-p*, in which an 'intrinsic' or neutral layer between the base and collector layers has permitted transistor operation at higher voltages and frequencies.

621.314.7 + 621.314.63] : 538.63 **2915**

Transistors and Diodes in Strong Magnetic Fields.—H. A. Kampf. (*Electronic Ind.*, March 1958, Vol. 17, No. 3, pp. 71–73.) The results of tests at field strengths up to 10 kG show that semiconductor devices give a more reliable performance in a magnetic-field environment than thermionic valves.

621.314.7 : 546.28 **2916**

The Blocking Capability of Alloyed Silicon Power Transistors.—R. Emeis & A. Herlet. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1216–1220.) Experimental results on a series of Si *n-p-n* alloyed transistors show good agreement with theory. Details of the *I/V* characteristic are discussed. Three base resistivity regions are distinguished: (*a*) pure breakdown, (*b*) approximately simultaneous occurrence of breakdown and punch-through, and (*c*) pure punch-through.

621.314.7 : 546.28 : 621.318.57 **2917**

The Electrical Characteristics of Silicon *P-N-P-N* Triodes.—I. M. Mackintosh. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1229–1235.) The *p-n-p-n* triode shows switching properties analogous to the conventional thyratron. A general analysis of four-region structures is given and applied specifically to the *p-n-p-n* triode. Much of the detailed behaviour of the device can be explained in terms of this analysis, and theoretical curves are given which are in good agreement with experimental results.

621.314.7 : 546.289 **2918**

Large-Area Germanium Power Transistors.—B. N. Slade & J. Printon. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 98–108.) Both *p-n-p* and *n-p-n* experimental alloy-junction power transistors have been developed to operate at collector currents of 10 A by increasing the junction area and using ring-type emitters. Collector-to-base current ratios range up to 200 at 1A and to 60 at 10 A. Thermal resistances are about 1–2°C/W.

621.314.7 : 546.289 **2919**

A High-Frequency Germanium Drift Transistor by Post-Alloy Diffusion.—J. S. Lamming. (*J. Electronics Control*, March 1958, Vol. 4, No. 3, pp. 227–236.) Two or more impurities of different conductivity type are alloyed into *n*-type Ge to produce an alloyed *p-n* junction which is then modified by diffusion to produce a diffused *p-n* junction, the graded *n*-type region being suitable for the base and the *p*-type region for the emitter of a drift transistor. Alpha cut-off frequencies in excess of 200 Mc/s with collector capacitances of 1·5 pF and base resistances of 50 Ω have been achieved by this method.

621.314.7 : 621.372.57 **2920**

The Advantage of Considering a Transistor as an Active Quadripole.—A. Pinciroli & S. Fubini. (*Ricerca sci.*, Jan. 1958, Vol. 28, No. 1, pp. 152–159.) The matrix analysis of valve networks used in 2744 of 1949 (Pinciroli & Taraboletti) is similarly applied to transistor circuits, which simplifies the solution of network problems.

621.314.7 : 621.396.822 **2921**

Noise in Junction Transistors.—A. van der Ziel. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1019–1038.) A survey of shot noise and flicker noise in junction diodes and transistors. The shot effect theory is treated in detail, both from the collective and corpuscular points of view. Fonger's theory of flicker noise is given. Experimental noise measurements are reviewed in relation to the theories described. 61 references.

621.314.7 : 681.142 **2922**

Silicon-Germanium Transistors.—J. J. Bowe. (*Electronic Equipm.*, Sept. 1957, Vol. 5, No. 9, pp. 26–27.) Temperature characteristics of point-contact transistors for computer applications are considered. Si-Ge alloy is shown by tests to be preferable to pure Ge.

621.314.7.004.15 **2923**

Evaluation of Transistor Life Data.—J. D. Johnson & B. VanSwearingen. (*Trans. Inst. Radio Engrs*, Aug. 1957, No. PGRQC-11, pp. 15–26. Abstract, *Proc. Inst. Radio Engrs*, Oct. 1957, Vol. 45, No. 10, p. 1432.)

621.383.27 **2924**

Temperature Dependence of Photomultiplier Gain.—F. E. Kinard. (*Nucleonics*, April 1957, Vol. 15, No. 4, pp. 92–97.) The gain stability of photomultipliers in the range −20° to +60°C was measured by recording output pulse height as a function of temperature when a light pulse of constant amplitude was applied.

621.383.27 **2925**

A Photomultiplier with Controllable Cathode Area.—W. Hartmann. (*Ann.*

Phys., *Lpz.*, 15th July 1957, Vol. 20, Nos. 1–6, pp. 247–249.) An improvement in signal/noise ratio can be achieved by reducing the effective area of the photocathode by means of a control electrode to the minimum necessary for a given application.

621.383.4 : 546.289 **2926**

Narrow-Base Germanium Photodiodes.—D. E. Sawyer & R. H. Rediker. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1122–1130.) The operation of Ge photodiodes at room temperature both as reverse-biased and photovoltaic detectors is analysed. General expressions are derived for the steady-state and the time-varying detector signal components. Equivalent circuits for both reverse-biased and photovoltaic operation are obtained as well as the noise equivalent circuit for reverse-bias operation. Advantages of a reduction in base width, and narrow-base photodiode design are discussed.

621.383.4 : 546.48.231 : 535.371.07 **2927**

A Hysteresis Effect in Cadmium Selenide and its Use in a Solid-State Image Storage Device.—F. H. Nicoll. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 77–85.)

621.383.4 : 546.48.241 **2928**

Cadmium Telluride Photovoltaic Cells.—G. A. Lomakina, Yu. A. Vodakov, G. P. Naumov & Yu. P. Maslakovets. (*Zh. tekh. Fiz.*, July 1957, Vol. 27, No. 7, p. 1594.) Samples having a specific conductivity of approximately $40\ \Omega^{-1}.\text{cm}^{-1}$ and a thermo-e.m.f. of $200\ \mu$V/deg were investigated. When exposed to sunlight, these photocells gave an e.m.f. > 500 mV with an efficiency of 2%.

621.383.4 : 621.397.611.2 **2929**

Differential Method of Lag Compensation in Photoconductive Devices.—H. Borkan & P. K. Weimer. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 62–76.) By taking the difference of the signals from two photoconductive elements having unlike transient responses, a resultant signal can be obtained having a faster response than either element alone. Measurements of lag compensation using a pair of commercial photocells and a pair of photoconductive camera tubes have been made. The application of the method to a single camera tube, designed to yield a lag-corrected signal directly, is discussed.

621.385.029.6 **2930**

Calculation of Resonance Frequencies of a Smooth-Anode Cylindrical Magnetron in the Brillouin State.—J. Coste. (*C. R. Acad. Sci., Paris*, 21st Oct. 1957, Vol. 245, No. 17, pp. 1404–1406.) The perturbation method applied earlier [1903 of 1956 (Coste & Delcroix)] is outlined, and results of calculation are considered with reference to the analysis of Harris (3293 of 1952). Restriction of the Brillouin state to a distance from the cathode of twice the cathode radius, and an alteration of sign in a formula of Harris, lead to real values for the resonance frequencies.

621.385.029.6 **2931**

An Electrostatically Focused Travelling-Wave-Tube Amplifier.—K. K. N. Chang. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 86–97.) The principle of biperiodic focusing (640 of February) has been applied by using a pair of concentric bifilar helices and an annular gun. Design curves for choice of optimum geometry of the tube for a given beam perveance using a minimum focusing field are given. Experimental d.c. tests have shown a current transmission of 97 % with a beam perveance of about 2×10^{-6} A/V$^{3/2}$. A 10 dB gain has been observed at 2 970 Mc/s.

621.385.029.6 : 537.533 **2932**

Note on a Method of Investigation by Analogy of Interaction in Valves with Crossed Electric and Magnetic Fields.—B. Epsztein. (*C. R. Acad. Sci., Paris*, 18th Nov. 1957, Vol. 245, No. 21, pp. 1790–1793.) The interaction between h.f. waves and an electron beam in crossed-field valves is investigated by means of a c.r. tube in which analogous effects occur. See also 3703 of 1957.

621.385.029.6+621.384.6] : 621.372.2 **2933**

The Propagation of Slow Waves.—J. Dain. (*Electronic Engng*, June 1958, Vol. 30, No. 364, pp. 388–393.) A brief history of the use of continuous interaction between electrons and a travelling wave is given. The basic properties of the field patterns of travelling e.m. waves as applied to particle accelerators and microwave valves are discussed and the techniques for measuring the main parameters are outlined. Consideration is given to simple and multiple periodic structures with particular reference to cross-wound helices and corrugated waveguides.

621.385.029.63 : 537.533 : 621.375.9 **2934**

Parametric Amplification of the Fast Electron Wave.—R. Adler. (*Proc. Inst. Radio Engrs*, June 1958, Vol. 46, No. 6, pp. 1300–1301.) A proposal for applying parametric amplification to the electron coupler described by Cuccia (2975 of 1949), by feeding a pumping signal to an electrode system so as to produce an inhomogeneous transverse field across the electron stream.

621.385.1 **2935**

The Schottky-Langmuir Law of Discharges in a New Reference System for Small Electrode Distances.—K. Mie. (*Hochfrequenztech. u. Elektroakust.*, July 1957, Vol. 66, No. 1, pp. 1–11.) A new approximation to the Langmuir solution is derived which is applicable to valves with small electrode distances, operating under conditions very close to the Boltzmann limit.

621.385.1 **2936**

Considerations Affecting the Rise and Decay of Cathode Currents in Receiving Tubes.—E. R. Schrader. (*RCA Rev.*, March 1958, Vol. 19, No. 1, pp. 109–127.) Curves, resulting from the application or removal of heater power, are analysed to determine their dependence on various valve properties. Current decay is more reproducible and permits better resolution. In some types of valve changes in cathode activation appear as shifts of the temperature-limited region and sometimes as transient changes in the space-charge-limited region. Experiments with a diode having a rotating anode are used to illustrate the effects of cathode poisoning and reactivation.

621.385.1 : 621.317.799 **2937**

High-Speed Tester Checks Tubes in Groups.—Gordon. (See 2843.)

621.385.3/.5+621.314.7] : 621.396.822 **2938**

Experimental Investigation of Low-Frequency Noise in Valves and Transistors.—B. V. Abramov & V. I. Tikhonov. (*Radiotekhnika, Mosk.*, June 1957, Vol. 12, No. 6, pp. 45–51.) Measurements of the spectral intensity of noise are described and tables are compiled showing results obtained for various operating conditions.

621.385.5 : 621.396.822 **2939**

The Noise Characteristics of the Valve Type EF86 in the Low-Frequency Region.—M. Jansen & H. Lembke. (*NachrTech.*, Nov. 1957, Vol. 7, No. 11, pp. 519–523.) Report on tests made on triode-connected valves to obtain the noise spectrum in the range 1–10^7 c/s.

621.387 **2940**

Lateral-Current Control Mechanism for Cold-Cathode Gas Discharges.—D. J. Belknap & L. R. Crump. (*J. appl. Phys.*, April 1958, Vol. 29, No. 4, pp. 737–738.)

621.387 : 621.316.722 **2941**

Noise and Impedance Measurements in Voltage-Regulator Tubes.—A. van der Ziel & E. R. Chenette. (*Physica*, Oct. 1957, Vol. 23, No. 10, pp. 943–952.) Results of a series of measurements over a frequency range 1 c/s–30 Mc/s indicate that at low frequencies both admittance and noise are influenced by the same current multiplication process.

MISCELLANEOUS

621.3.004.15 **2942**

On the Measurement of Component Reliability.—I. K. Munson. (*Trans. Inst. Radio Engrs*, Aug. 1957, No. PGRQC-11, pp. 27–33.) In reliability tests at elevated temperature care must be exercised to ensure that the rise in temperature does not cure and hence conceal a defect in the component e.g. by annealing soldered parts.

413.164 : [621.38+621.372.8 **2943**

Elsevier's Dictionary of Electronics and Waveguides in Six Languages. [Book Review]—W. E. Clason. Publishers: Elsevier, Amsterdam, and Clever-Hume Press, London, 1957, 628 pp., 90 s. (*Nature, Lond.*, 22nd March 1958, Vol. 181, No. 4612, p. 801.) The languages included are English, French, Spanish, Italian, Dutch and German.

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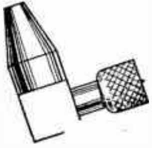

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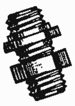

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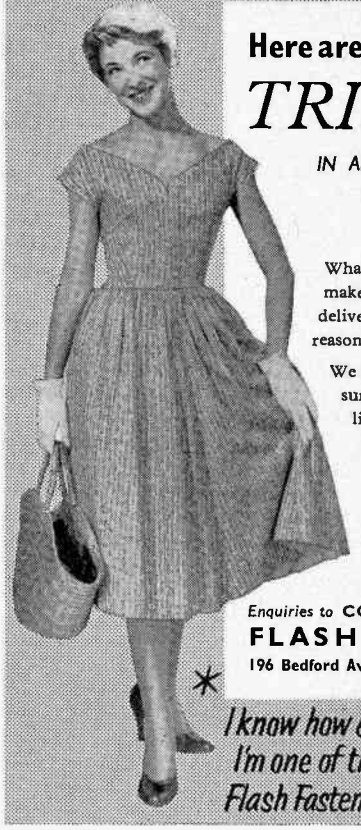

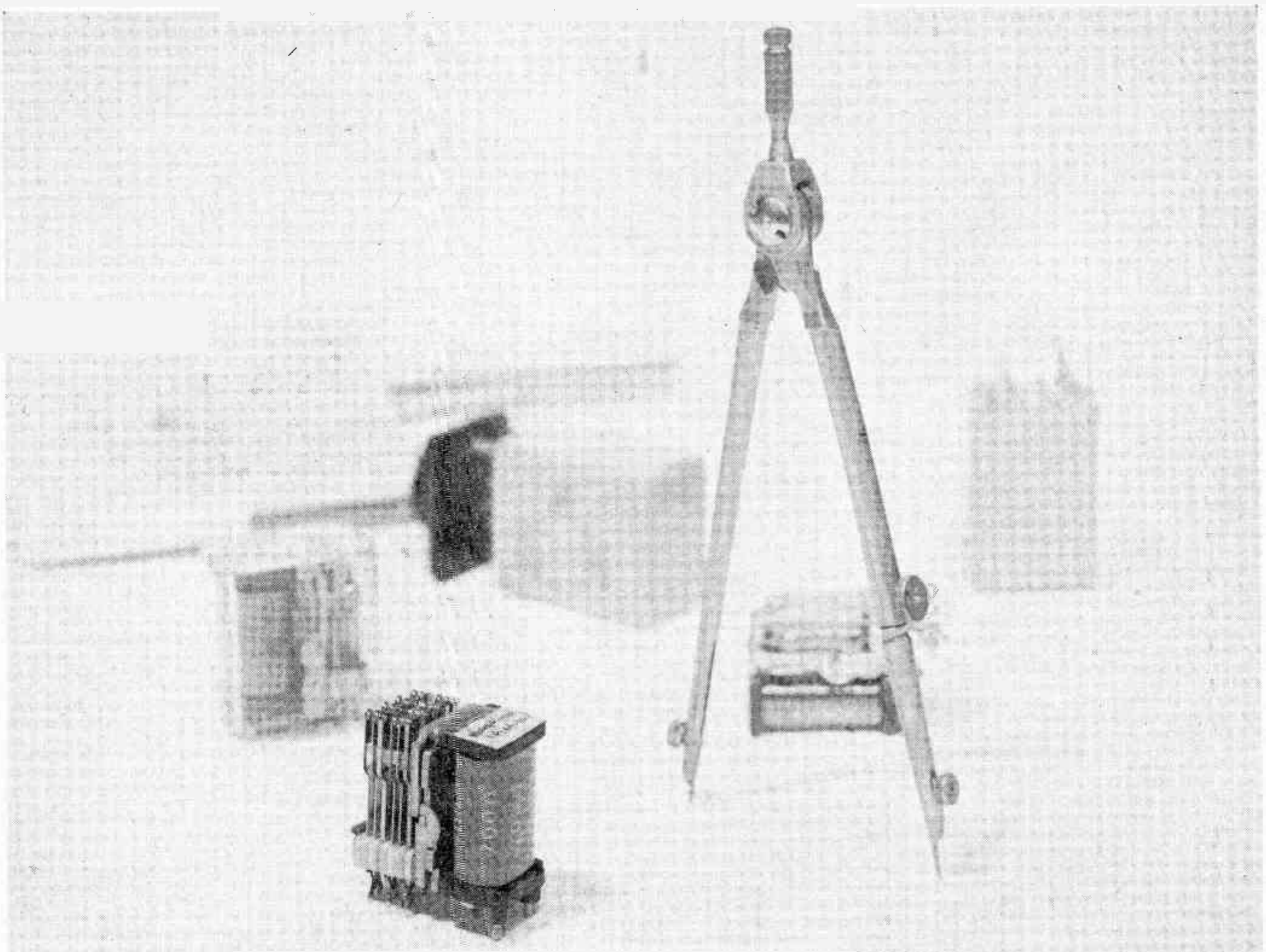

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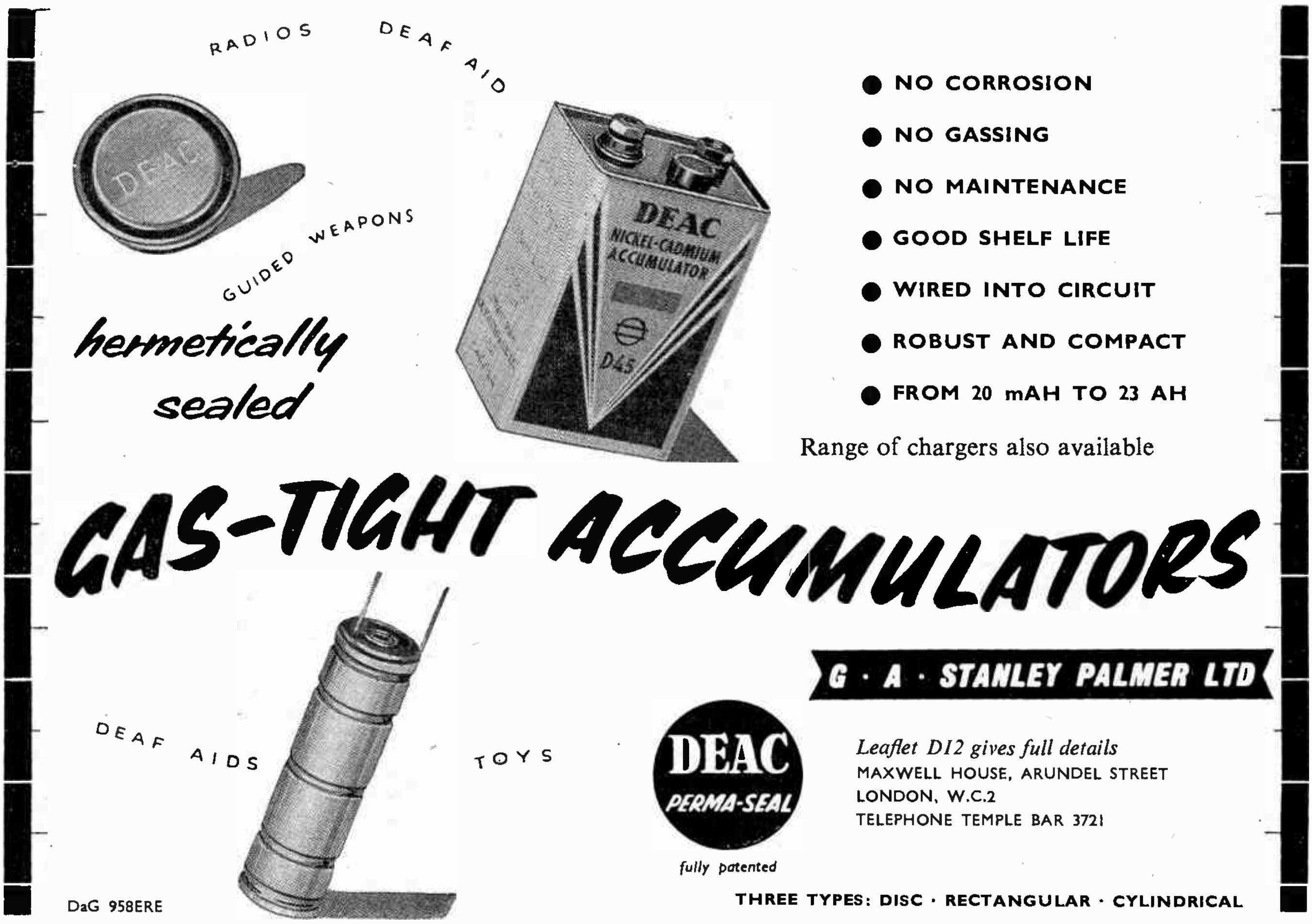

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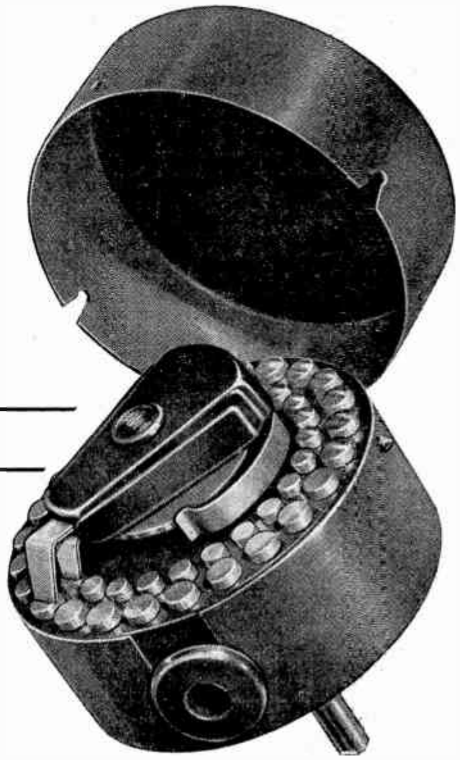

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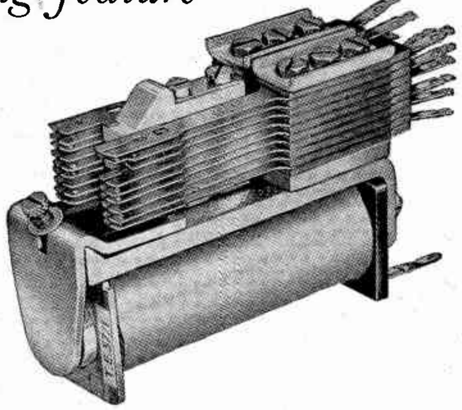

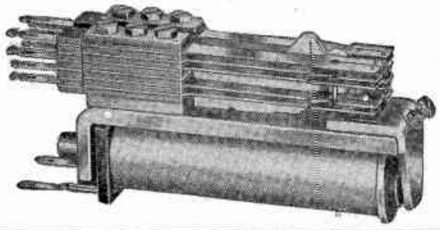

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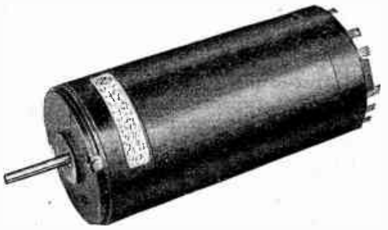

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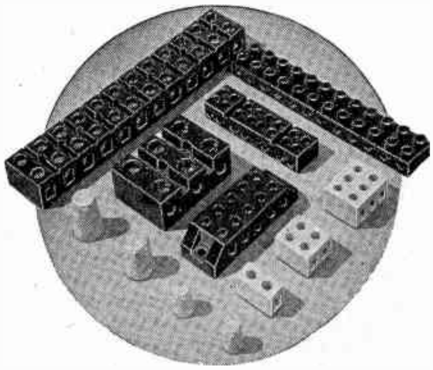

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ELECTRONIC & RADIO ENGINEER

CLASSIFIED ADVERTISEMENTS

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Semi-Display *advertisements with centralized lines* £1/15/- *per single column inch—no additional charge for Box Numbers. The Publishers are not responsible for clerical or printers' errors although every care is taken to avoid mistakes.*

Press date for the October 1958 issue is first post 24th September 1958

SITUATIONS VACANT

FEDERAL BROADCASTING CORPORATION OF RHODESIA AND NYASALAND

APPLICATIONS are invited for the following posts. Appointments on probation for one year in the first instance with prospects of permanent employment. Contributory pension scheme. Free passage on first appointment and assistance towards family passages. Liberal leave on full salary. Holiday Benefit Fund.

Maintenance Engineer—Transmitters (M2C/50191/EO)
SALARY scale £1,200 rising to £1,600 a year. Candidates, not over 45, should possess at least C. and G. Final Cert. in Telecomms. or equivalent. They must have a sound theoretical and practical knowledge of broadcast radio transmitters and ancillary plant; radio frequency feeders; aerials and arrays, and methods of impedance matching. They must be capable of organising routine maintenance of technical equipment, supervising staff, store-keeping, etc. Knowledge of workshop practice and ability to drive car desirable.

Recording Engineer (M2C/50192/EO)
SALARY £1,200 rising to £1,600 a year. Candidates not over 45, should possess at least C. and G. Final Cert. in Telecomms. or equivalent. They should have had at least five years' recent specialised and practical recording experience with a reputable recording engineering organisation. They should be familiar with modern methods of magnetic tape and disc recording and with the maintenance, adjustment, and testing of studio and portable recording machines. They should also have a good working knowledge of studio equipment, balance technique, outside broadcasts and mobile recording. Knowledge of workshop practice desirable and ability to drive a car essential.

Technicians
SALARY scale £800 rising to £1,400 a year. Candidates, not over 35, should possess at least C. and G. Inter. Cert. with Radio I or equivalent. Vacancies in following categories:

Technician—Transmitters (M2C/50194/EO)
CANDIDATES must have a sound theoretical and practical knowledge of broadcast M.F. and H.F. transmitters and ancillary equipment. They should have had at least three years' practical experience on the maintenance and operation of such transmitters with a reputable broadcasting or telecomms. organisation. Knowledge of feeders, aerials and methods of construction, erection and matching, an advantage. Practical knowledge of simple constructional workshop methods desirable.

Technician—Studio (M2C/50203/EO)
CANDIDATES must have a sound theoretical and practical knowledge of studio programme input equipment and ancillary apparatus, with its operation, testing and maintenance. They should have had at least three years' practical experience directly associated with such equipment preferably with a reputable broadcasting organisation. Knowledge of modern constructional and technical wiring methods and ability in light workshop practice highly desirable.

Technician—Installation (M2C/50206/EO)
CANDIDATES, preferably single, should have a good general knowledge of broadcasting engineering plant. They should have specialised knowledge of modern practice in the installation of such plant and should be capable and versatile in the use of tools and workshop equipment.

Technician—Recording (M2C/50208/EO)
CANDIDATES should have a good general knowledge of broadcasting programme input equipment with specialised knowledge of recording methods and practice. They must have had at least two years' practical experience in tape and disc recording with an organisation of repute and must be familiar with the testing and maintenance of recording machines and with methods of mobile and portable recording. Ability in technical wiring and workshop practice desirable. Candidates should be able to drive a recording van and hold a valid driving licence.

WRITE to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote the reference number shown against the post applied for. [1232

SITUATIONS VACANT

SENIOR SCIENTIFIC OFFICERS

required by

The Atomic Weapons Research Establishment, Aldermaston, Berkshire

(for duty at Orfordness, Suffolk)

POST 1. To be directly responsible to the Officer-in-Charge for the general direction and management of a team engaged on a programme of advanced development and complex field trials. Good organising ability and interest in technical management are essential for this post.

THE team is responsible for the design of airborne telemetry equipment and associated ground stations, the organising and carrying out of complex field and flight trials and the evaluation of the results. There is ample scope for original design.

POST 2. To take charge of a section of the above team engaged in advanced electronic instrumentation development and to be responsible for the detailed technical direction of this work.

FOR both posts a 1st or 2nd Class Honours Degree or equivalent is required and several years' experienec in electronics or communication equipment development including pulse and V.H.F. techniques. Experience in airborne radar and G.W. equipment development will be considered a distinct advantage.

SALARY: £1,215–£1,425 p.a.

CONTRIBUTORY Superannuation Scheme. A house or assistance towards legal expenses on house purchase will be available for married officers living beyond daily travelling distance.

POSTCARDS for application forms to the Senior Recruitment Officer at A.W.R.E., Aldermaston, Berks. Please quote ref. 1969/46. [1227

GUIDED Weapons. Physicists and Engineers required by Ministry of Supply at Royal Aircraft Establishment, Armament Research and Development Establishment and Royal Radar Establishment, for research on guided weapons (including ballistic missile) and supervision of development in industry. 1st or 2nd class hons. degree or equivalent required for Scientific Officer (£595–£1,050) and three years' post-graduate experience in addition for Senior Scientific Officer (£1,130–£1,330). Women's rates the same by 1961. Superannuation under F.S.S.U. Opportunity for those under 32 to compete for established posts. Forms from M.L.N.S. Technical and Scientific Register (K), 26 King Street, London, S.W.1 (quote A.291/8A). Closing date 24th September, 1958. [1234

MINISTRY OF SUPPLY RESEARCH AND DEVELOPMENT ESTABLISHMENTS at Farnborough (Hants), Bedford, Aberporth, Llanbedr (Wales), Sevenoaks (Kent), Waltham Abbey (Essex), Harefield (Middx.), Risley (Lancs) require (*a*) Assistant Experimental Officers (minimum age 18 and (*b*) Experimental Officers (men and women) for experimental work in Physics, Electronics, Electrical Engineering, Chemistry or Mathematics. Qualifications: H.S.C., G.C.E. (AL), H.N.C., pass degree or equivalent. Salaries (men) (*a*) £370 (age 18)–£685 (age 26)–£850; (*b*) £920–£1,190. Applicants should state preferred fields of work and location. Forms from M.L.N.S. Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting A.233/8A. [1221

MINISTRY OF SUPPLY RESEARCH AND DEVELOPMENT ESTABLISHMENTS, mainly in southern half of England, require (*a*) Senior Scientific Officers (minimum age 26) and (*b*) Scientific Officers for work in physics, electronics, electrical or mechanical engineering, applied mathematics, aeronynamics, chemistry or metallurgy. 1st or 2nd class hons. degree or equivalent required and for S.S.O. at least three years' post-graduate experience. Starting salary in range (*a*) £1,130–£1,330, (*b*) £595–£1,050 (male, in provinces). Rates for women reaching equality in 1961. Superannuation under F.S.S.U. Opportunity for those under 32 to compete for established posts. Candidates should indicate fields of work in which interested. At National Gas Turbine Establishment, Pyestock, Hants, houses available for letting to married staff, and opportunities for new graduates to have workshop training. Forms from M.L.N.S. Technical and Scientific Register (K), 26 King Street, London, S.W.1. (Quote A.300/8A). [1235

SITUATIONS VACANT

TECHNICIANS

required by

The Atomic Weapons Research Establishment, Aldermaston, Berkshire

(for duty at Orfordness, Suffolk)

TO be a member of a team engaged on complex field trials involving the use of telemetry devices. Duties will include the operation and maintenance of a wide range of radio, radar and electronic instruments and equipment.

ATTENDANCE at overseas trials will be required from time to time on a rotational basis.

OPPORTUNITIES may occur for some officers to work on the development and production of "one off" experimental devices and equipment.

A recognised apprenticeship or equivalent training in electronics is essential. Previous experience in field trials or in the technical branches of the fighting services as well as several years' experience in electronics, an advantage.

SALARY: £755 (at age 28)—£875 p.a.

CONTRIBUTORY Superannuation Scheme. A house or assistance towards legal expenses on house purchase will be available for married officers living beyond daily travelling distance.

POSTCARDS for application forms to the Senior Recruitment Officer at A.W.R.E., Aldermaston, Berks. Please quote ref. 1976/46. [1230

ELECTRONICS ENGINEER

COURTAULDS LIMITED invites applications from graduate Electronics Engineers. The successful applicant will be about 25 years of age and will be familiar with digital computer techniques or have had experience in the development of industrial electronic measuring and control systems. The work he will do will be unusually varied and interesting.

CANDIDATES should write for a detailed form of application to the Director of Personnel, Courtaulds Limited, 16 St. Martins-le-Grand, London, E.C.1, quoting reference number C.99. [1236

ELECTRICAL Engineer required to be responsible for installation of high voltage radiotherapeutic equipment at Hospital's Branch, Sutton. Equipment will include linear accelerator, and probably Van de Graaff generator. Appointment for three years in first instance. Salary in accordance with Whitley Council scale for Chief Physics Technician, £785 to £1,005, plus London Weighting. Applications, giving names and addresses of three referees to be sent to the House Governor, Royal Marsden Hospital, Fulham Road, S.W.3. [1237

MINISTRY OF TRANSPORT AND CIVIL AVIATION requires Electrical Engineers (Assistant Signals Officers) for aviation telecommunications and electronic navigational aids. Minimum age 23, 1st or 2nd Class degree in Physics or Engineering, or A.M.I.E.E. or A.F.R.Ae.S. (candidates with Parts I, II and III of A.M.I.E.E. or Parts I and II of A.F.R.Ae.S. or equivalent, or of very high professional attainment without these qualifications considered). Salary £665 (age 23) to £1,085 (age 34). Minimum £1,250. Slightly lower outside London and for women. Five-day week. Further details and forms from M.L.N.S., Technical and Scientific Register (K), 26 King Street, London, S.W.1, quoting D.129/8A. [1222

INSPECTOR of Lights required by Sarawak Government Marine Department either on probation to pensionable establishment or on contract with gratuity of 12½ per cent of final monthly salary for each completed month of service. Salary scale (including Inducement pay) equivalent to £1,176 to £1,932 a year. Child allowance between £72 and £123 a year. Education allowance up to £280 a year. Outfit allowance £60. Free passages. Liberal leave on full salary. Candidates, preferably under 35, should be A.M.I.E.E. or A.M.I.Mech.E. and have a good all-round knowledge of electricity especially in connection with the operation of lighthouse flashers (Chance/Londex), Reed and Klaxon Motors, Echo Sounding Gear, otherwise straightforward small unit diesel electric generation, current conversion, etc., a knowledge and experience of small diesel engines, working knowledge V.H.F. radio equipment and have a good head for heights up to 130 ft. Write to the Crown Agents, 4 Millbank, London, S.W.1. State age, name in block letters, full qualifications and experience and quote M2A/50186/EO. [1231

SITUATIONS VACANT

ATOMIC ENERGY RESEARCH ESTABLISHMENT

Harwell

A highly skilled and adequately trained Electronic Technician is required to be responsible for a team installing, commissioning and maintaining electronic equipment associated with the 7 GeV Proton Synchrotron. Practical and supervisory experience with high and low power R.F. equipment, controls and monitoring equipment are required. H.N.C. or C. and G. Final certificates advantageous.

SALARY: £1,050–£1,270.

HOUSING; superannuation; excellent staff conditions.

SEND postcard to the Group Recruitment Officer (1178/46, U.K.A.E.A. Atomic Energy Research Establishment, Didcot, Berks., for application form and details. [1229

SITUATIONS VACANT

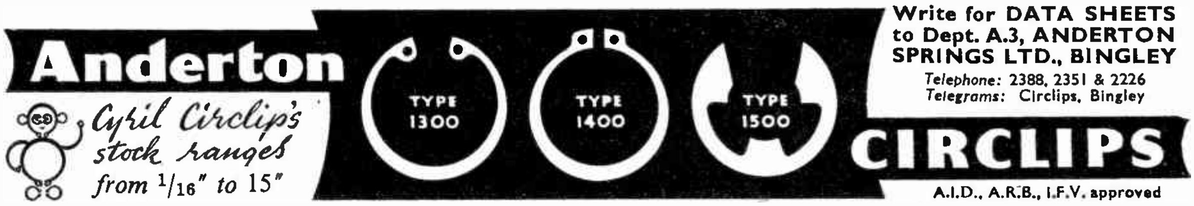
APPLICATIONS are invited for three appointments of Demonstrator in the Department of Aircraft Electrical Engineering. Responsibilities include everyday running of teaching and research laboratories, development of new equipment, supervision and checking of students' work, and assisting with research. The laboratories are those devoted to (*a*) radio and radar, (*b*) control systems and electronic computers and (*c*) E.M. machines and power distribution, and are used by students of the two-year course in light current electrical engineering and the one-year advanced course in guided weapons. Candidates should hold a University degree or equivalent professional qualification and should preferably have had previous experience in a suitable field. Salary, dependent on previous experience, in range £600 to £1,000 per annum, with superannuation under F.S.S.U. and family allowance. Successful candidate might later be considered for promotion to Lecturer. Applications giving full particulars and quoting the names of three referees, to the Recorder, The College of Aeronautics, Cranfield, Bletchley, Bucks. [1228

BUSINESS OPPORTUNITIES

CANADIAN Sales for you. We successfully represent to the Government of Canada, manufacturers in United States and Canada, covering electronics and allied industries. Department of Defence Production is our largest customer. We are an incorporated company with twenty-five years' experience in the electronic industry in Canada. References will be made available. We have an important sales story to tell British manufacturers at this time. Box 6890. [1238

BOOKS WANTED

WANTED. *Wireless Engineer*, January–July, 1952. Index 1953. November–December, 1954. February–June–December, Index, 1955. January–December, Index, 1956. *Electronic & Radio Engineer*, January–February, 1957. Offers to R. Morris, 16 Magdala Road, Nottingham. [1233



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EDISWAN
MAZDA

Transistor news

Ediswan Mazda transistors are currently used in a wide range of electronic equipment. Get the facts about them at your finger tips—if you have not already applied for a complete set of Ediswan Mazda semiconductor Technical Data Sheets, we suggest that you do so now on your company's letter heading.

A SYMMETRICAL TRANSISTOR FOR SWITCHING CIRCUITS AND MODULATORS

The new Ediswan Mazda XS101 transistor has two identical 'P' type germanium electrodes; with appropriate bias conditions either will act as emitter with the other as collector. The average characteristics for the two conditions are identical.

**The new XS101 transistors can now be supplied, against order, for evaluation purposes.*

SYMMETRICAL TRANSISTOR TYPE XS 101

TENTATIVE RATINGS. Absolute values for 45°C. ambient

Maximum mean or peak collector/emitter voltage (with base maintained at least 1 v. positive with respect to the positive end of the emitter supply battery)	V	—20
Maximum mean or peak collector/emitter voltage (conducting)	V	—12
Maximum mean or peak collector to base voltage	V	—21
Maximum collector dissipation	mW	90
Maximum junction temperature	°C	75
Thermal resistance in free air	°C/mW	0.33

TENTATIVE CHARACTERISTICS at 25°C.

*Common base cut-off frequency (minimum)	Mc/s	2.5
*Average Current Amplification. Common Emitter (Degree of asymmetry, 1.5 to 1)	β	20

**Small signal values at* $V_c = -5\,V.$, $I_c = -1\,mA.$

INCREASED RATINGS FOR EDISWAN MAZDA TRANSISTORS

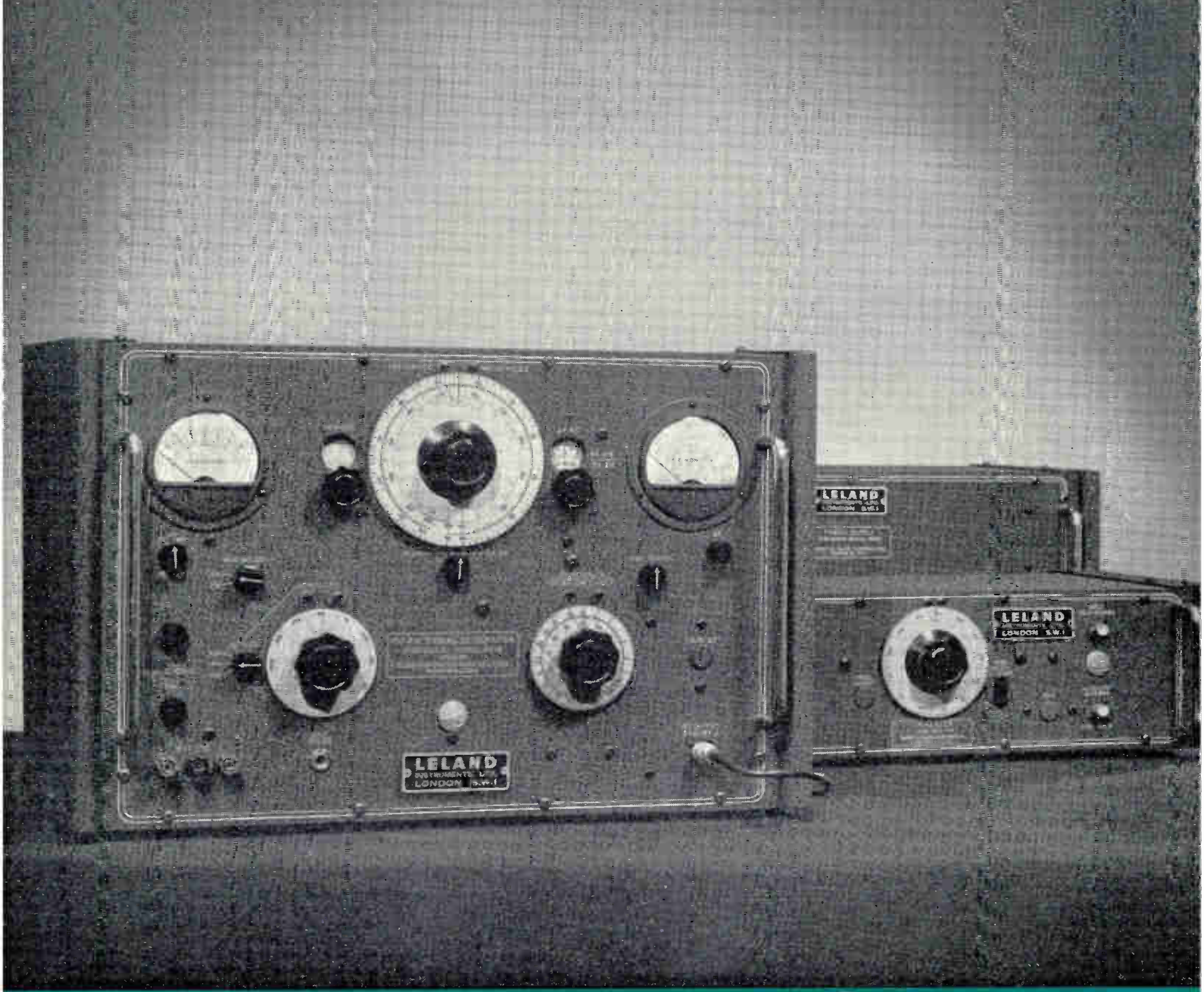
The table on the right shows the increased ratings (absolute at 45°C.) which now apply.

TRANSISTOR TYPE XC 101	
Maximum peak or mean collector/emitter voltage (common emitter circuit)	—16 v.
Maximum peak collector to emitter voltage with base driven to cut off (common emitter circuit) with external base/emitter circuit resistance less than 500 ohms	—35 v.
Maximum peak or mean collector/base voltage (common base circuit)	—35 v.
Maximum junction temperature	75°C.
TRANSISTOR TYPES XB 102 AND XB 103	
Maximum peak or mean collector/emitter voltage (common emitter circuit)	—16 v.
Maximum peak collector to emitter voltage with base driven to cut off (common emitter circuit) with external base/emitter circuit resistance less than 500 ohms	—35 v.
Maximum peak or mean collector/base voltage (common base circuit)	—35 v.
TRANSISTOR TYPES XA 101 AND XA 102	
Maximum peak or mean collector/emitter voltage (common emitter circuit)	—16 v.
Maximum peak or mean collector/base voltage (common base circuit)	—20 v.

SIEMENS EDISON SWAN LIMITED, *An A.E.I. Company*
Valve and CRT Division, 155 Charing Cross Road, London W.C.1.

CRC 15/16

World Radio History



F.M. *The latest* SIGNAL GENERATOR *by Boonton* MODEL 202-E

Developed to meet the needs of engineers engaged in the design of F.M. and television receivers for operation within the frequency range 54 to 216 megacycles, this instrument embodies the results of twenty years' experience in F.M. generators, gained in the production of the well-known Boonton 150 and 202-B series.

★ FREQUENCY: 54-216 Mc/s may be extended down to 100 Kc/s by using the Univerter Type 207-E below.

★ R.F. OUTPUT VOLTAGE: Maximum open circuit voltage at front panel jack is approx. 0.4 volts. With output cable attached 0.2 volts nominal. Output impedance 50 ohms resistive at front panel jack. Minimum output 0.1 microvolts.

★ FREQUENCY MODULATION: Three deviation ranges, 0-24 Kc/s, 0-80 Kc/s, and 0-240 Kc/s, each continuously adjustable. FM distortion at 75 Kc/s is less than 2% and at 240 Kc/s less than 10%.

★ FIDELITY CHARACTERISTICS: Deviation sensitivity of the Frequency Modulation system as a function of frequency is flat within ±1 Db from 30 c/s to 200 Kc/s.

★ AMPLITUDE MODULATION: Internal AM is available from zero to 50% with meter calibration points at 30% and 50%. External modulation may be used over the range 0-50%. A front panel jack connects to the screen of the final stage for pulse and square wave modulation.

★ SPECIAL FEATURES: Incremental frequency range: The Δ F switch permits tuning in increments of ±5, ±10, ±15, ±20, ±25, ±30, ±50, ±60 Kc/s in the 108 to 216 Mc/s range—half these values in the 54 to 108 Mc/s range. A fine tuning control permits continuous tuning over a range of approximately ±20 Kc/s in the 108 to 216 Mc/s range, and ±10 Kc/s in the 54 to 108 Mc/s range.

Boonton instruments are available subject to import licence, from the SOLE CONCESSIONAIRES for the United Kingdom—

LELAND INSTRUMENTS LTD.

Electronic Engineers & Consultants

TELEPHONE: VICTORIA 3243 (FIVE LINES)

22 & 23 MILLBANK · WESTMINSTER · LONDON S.W.1