ELECTRONIC & RADIO ENGINEER

Incorporating **WIRELESS ENGINEER**

In this issue

Three shillings
and sixpence

APRIL 1959 Vol 36 *new series* **No 4**


high efficiency
with low cost
Whatever the aerial or frequency there is a BICC cellular-polythene Downlead to suit every television requirement. These cables are made under strict process quality control. Yet they are low in cost, providing the most economical method of ensuring high quality performance. Publication No. 357 gives further technical details. May we send you a copy?
BICC T/V downleads

Miss Neasden says she does not know what a resistor is

she should ask us... we know quite a lot about them

wire wound resistors by

Electrothermal

STAND NO. 135

R.E.C.M.F. Exhibition, Grosvenor House, W.1.
April 6th to April 9th, 1959

ELECTROTHERMAL ENGINEERING LTD., 270 NEVILLE ROAD, LONDON, E.7.

Now!

d.c. to V.H.F.

The TYPE A79

a range of WIDE BAND ATTENUATORS

by **Advance** –to be sure!

FOUR MODELS

★ **TYPE A75** Range 99dB in 1dB steps. Input and Output Impedance 75 ohms. Price **£19** net in U.K.

★ **TYPE A76** Range 90dB in 10dB steps. Input and Output Impedance 75 ohms. Price **£8 10s.** net in U.K.

★ **TYPE A79** Range 99dB in 1dB steps. Input and Output Impedance 75 ohms. *In metal case for laboratory use.* Price **£22 10s.** net in U.K.

★ **TYPE A84** Range 99dB in 1dB steps. Input and Output Impedance 75 ohms. Price **£17 10s.** net in U.K.

TYPE A94 *Type A76 supplied less resistors for customers to fit their own network. Price* **£7 10s.** *net in U.K.*

To our comprehensive range of attenuators, we announce the addition of four new models.

These attenuators employ resistive ladder network and provide accurate attenuation over a wide frequency band.

They are small compact units, ideal for building into customers' own equipment or for bench use in the laboratory.

Full details are given in leaflet No. R57 gladly sent on request.

GD 68/B **Advance** COMPONENTS LIMITED · ROEBUCK ROAD · HAINAULT · ILFORD · ESSEX · TELEPHONE : HAINAULT 4444

Marconi

A.M. SIGNAL GENERATOR TF801D

The TF 801D is the latest addition to the Marconi family of precision a.m. generators. With a frequency range of 10 to 470 Mc/s, its salient features include superfine tuning with crystal checking, and oscillator l.t. regulation for maximum stability. Spurious f.m. is less than 0·001% of carrier frequency, and its high-quality 50-ohm output has a v.s.w.r. better than 1·2.
Carrier level is continuously variable from 0·1 μV to 1 volt and is stabilized by an automatic level control system. Sinewave a.m. up to 90% may be applied both internally and externally; pulse modulation may be applied externally in the p.r.f. range 50 c/s to 50 kc/s.

Full details will be gladly sent on request—please ask for leaflet V142.

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AM & FM SIGNAL GENERATORS · AUDIO & VIDEO OSCILLATORS
FREQUENCY METERS · VOLTMETERS · POWER METERS
DISTORTION METERS · FIELD STRENGTH METERS
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Please address enquiries to MARCONI INSTRUMENTS LTD. at your nearest office:

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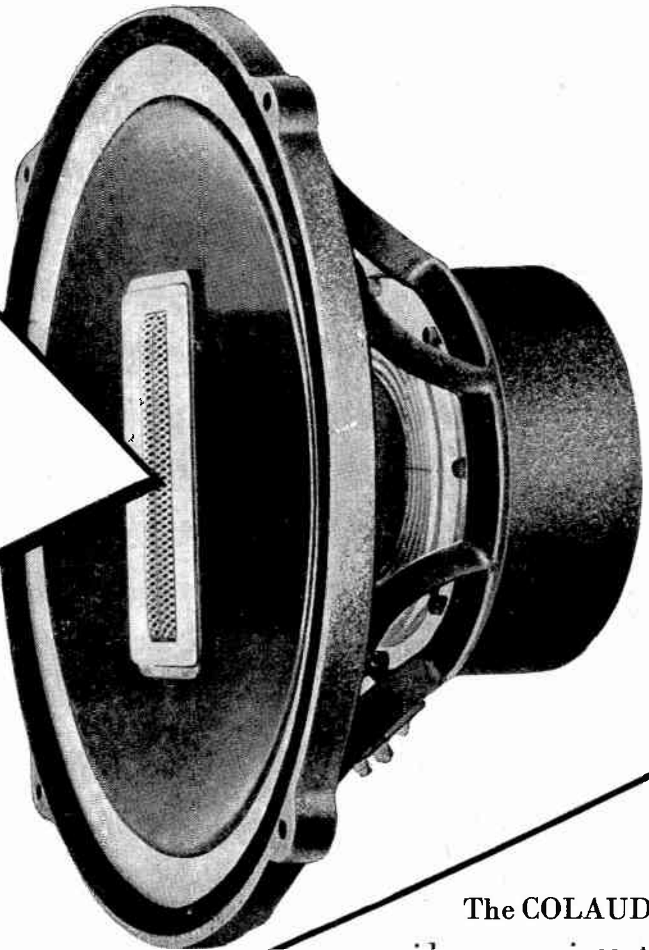
TC 142

. . . a new approach to better listening



The COLAUDIO provides a new incentive to listening, creates a new realism in reproduced sound, adds a new beauty to music and the finer nuances of speech. Combining a 15 in. direct radiator bass loudspeaker with two direct radiator, pressure-type high frequency reproducers in column form, the COLAUDIO is the culmination of over thirty years research, development and manufacture of loudspeakers for all purposes. Its perfection of tone can be truly appreciated only by an aural test—once heard, you will never be satisfied until you instal one in your own reproducing equipment

ESSENTIAL DATA

NOMINAL SIZE	15"
PEAK POWER HANDLING CAPACITY	25 watts.
VOICE COIL DIAMETER	3"
TOTAL FLUX	290,000 Maxwells
FREQUENCY RESPONSE	30-15,000 c/s
BASS RESONANCE	35 c/s
IMPEDANCE AT 400 c/s	15 ohms.

CELESTION

COLAUDIO

Rola Celestion Ltd. THAMES DITTON, SURREY, ENGLAND. Telephone: Emberbrook 3402/6

You can nearly always find a screw in the vast Unbrako range calculated to do just what you want better than any other screw.

But modern developments sometimes call for special screws not even standard to Unbrako.

When that happens our highly trained team of fastener-minded experts really get enthusiastic, responding to the challenge. They like to co-operate with you at the blueprint stage for preference, helping to design the perfect screw for the job, or they will simply make the screw to your specification, just about as well as a screw can be made.

So, standard or special, you can always safely specify Unbrako, the people who offer the most comprehensive specialised screw service in the world.

To be on the safe side, better get in touch with Unbrako over any fastener question. Remember what they say — Unbrako screws cost less than trouble.

There are two Unbrako lists you should have in your library, and a postcard or telephone call will bring them by return. They form a detailed and comprehensive guide to the whole Unbrako range, standard and non-standard, and no progressive firm should be without them.

Unbrako screws cost less than trouble

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New oscillograph 1059

ADVANCED *TRUE DOUBLE-BEAM OSCILLOGRAPH

***** *True double-beam—i.e. both beams use a common x-axis and there is no beam switching.*

CATHODE-RAY TUBE
Cossor 4 in. (10 cm.) double-beam, p.d.a., type 93D with green fluorescence, operating with overall accelerating potential of 3 kV or 6 kV.

Y1 AMPLIFIER
1 c/s to 10 Mc/s (30% down).
Rise-time: 0.04 μsec.
Output deflection: 6 cm (4 cm at 10 Mc/s).
Sensitivity: calibrated 100 mV/cm to 10 V/cm.
Sensitivity control: in steps 3:1 and 10:1 with continuously variable intermediate control.
Input Attenuator impedance: 1.2 MΩ and 65 pF.

Y2 AMPLIFIER
Identical with Y1 amplifier.

SIGNAL DELAY
200 mμsec approximately. Not more than 10 mμsec differential between channels.

PRE-AMPLIFIER (2)
Gain 10. 5 c/s to 200 kc/s (30% down).
Input Resistance: 3 MΩ.
One for A1 amplifier, the other for A2 or X amplifier.

PROBES (OPTIONAL EXTRA)
Frequency-compensated "L" attenuator.
Input impedance: 6 MΩ and 15 pF.
Insertion loss: 10:1.

TIME-BASE
Triggered.
Range: 0.03 μsec/cm to 15 msec/cm in eleven steps. Triggered from positive or negative signals derived externally or from Y1 amplifier.
Sensitivity: pulse—1 cm. deflection or 2 V external. Sine wave—2 cm deflection or 2 V r.m.s. external at frequencies up to 5 Mc/s.
Expansion amplifier, continuously variable gain up to 5 times. Time-base output available at front panel on slow speed ranges.
Delayed time-base: continuously variable delay 2 μsec to 150 μsec. Delay jitter not greater than 1 part in 1,000. Sensitivity pulse —1 cm deflection or 2 V external.

X AMPLIFIER
10 c/s to 750 kc/s (30% down).
As time-base amplifier: continuously variable expansion up to 5 times.
As independent X amplifier: sensitivity variable from 1 V/cm to 100 V/cm in 5 ranges.

CALIBRATION
Voltage measurement: internal calibrating voltage (square wave) referred through sensitivity control of the amplifiers. Accuracy ±3%.
Time measurement: by directly calibrated X shift control (±5%) and/or by 20 mμsec (±3%) black-out pips (for accurate measurement of rise-time).

POWER SUPPLY
Mains: 100 V to 130 V and 200 V to 250 V.
Frequency: 50 c/s to 100 c/s.
Consumption: 550 W.
Internal supplies are stabilized where necessary.

SIZE AND WEIGHT

Height:	17½ in.	(43.2 cm).
Width:	12 in.	(30.5 cm).
Depth:	24¾ in.	(62.9 cm).
Weight:	80 lb.	(36.3 kg).

ACCESSORY
Camera Model 1428.

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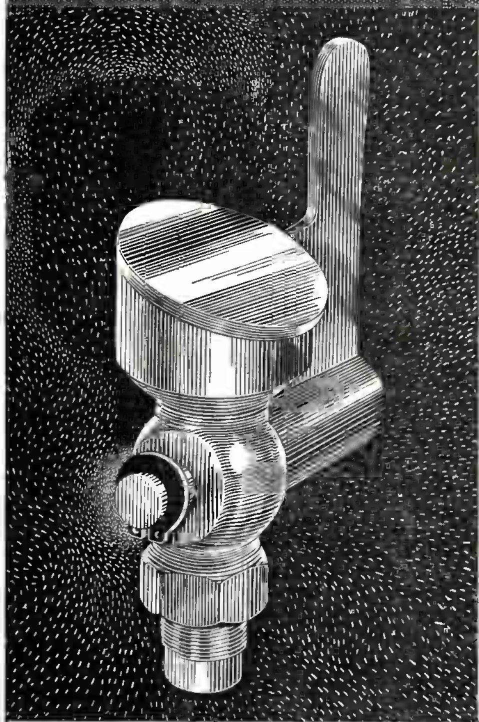
The Instrument Company of the Cossor Group

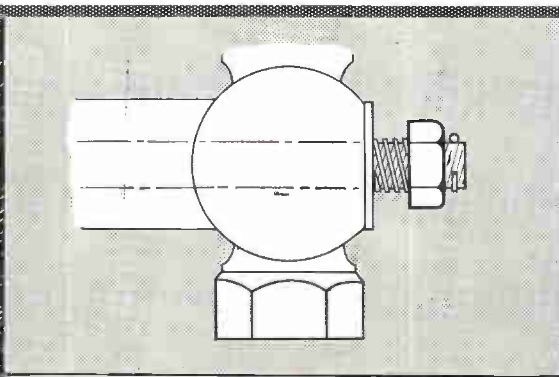
COSSOR HOUSE, P.O. BOX 64, HIGHBURY GROVE, LONDON, N.5.

Telephone: CANonbury 1234 (33 lines). Telegrams: Cossor, Norphone, London. Cables: Cossor, London. Code: Bentley's Second.

TAS/CI.11

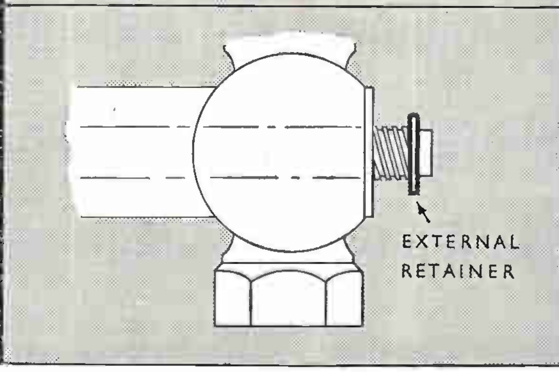
The logical advance in Retaining





OLD WAY

To provide a shoulder for the tensioning spring on this filler cup entailed an extra long, threaded shaft, a nut, a hole drilled to take a cotter pin and an altogether tedious assembly.



THE SALTER WAY

The spring is quickly positioned on a shorter, **PLAIN SHAFT** by a ***Grip Ring*** which is snapped into position for secure, frictional grip. No groove required.

save material–reduce assembly time–cut costs

When it's a question of assembling components in any engineering field, Salter Retainers are the answer. They replace nuts and bolts, screws, cotter pins, and eliminate expensive threading and machining operations. A large standard range is at your immediate disposal, and we should welcome the opportunity to assist in developing special retainers to solve your problems.

Send for the Salter Retainer catalogue—no designer is complete without it.

NEATER – MORE POSITIVE – PERMANENT RETAINING

SALTER "TRUARC" Regd. Trade Mark

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M-W.444

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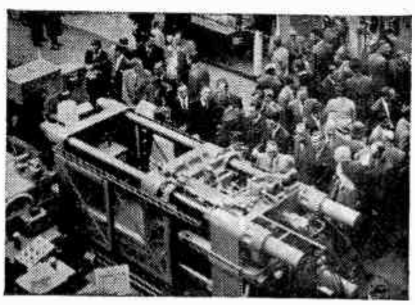
international plastics exhibition and convention

GRAND, NATIONAL AND EMPIRE HALLS OLYMPIA, 17-27 JUNE, 1959

The INTERNATIONAL PLASTICS EXHIBITION at Olympia, London, 17-27 June, will provide a complete up-to-the-minute picture of plastics progress in the industries of the world.
Thousands of exhibits from some 15 countries will combine to present the newest plastics materials, the most up-to-date equipment and a comprehensive range of industrial and domestic finished products. Here you can talk plastics chemistry, mechanics and merchandising with experts from many countries—see plastics machinery in action—gather new product ideas and new manufacturing methods—study new techniques. Do not miss this vitally important event.



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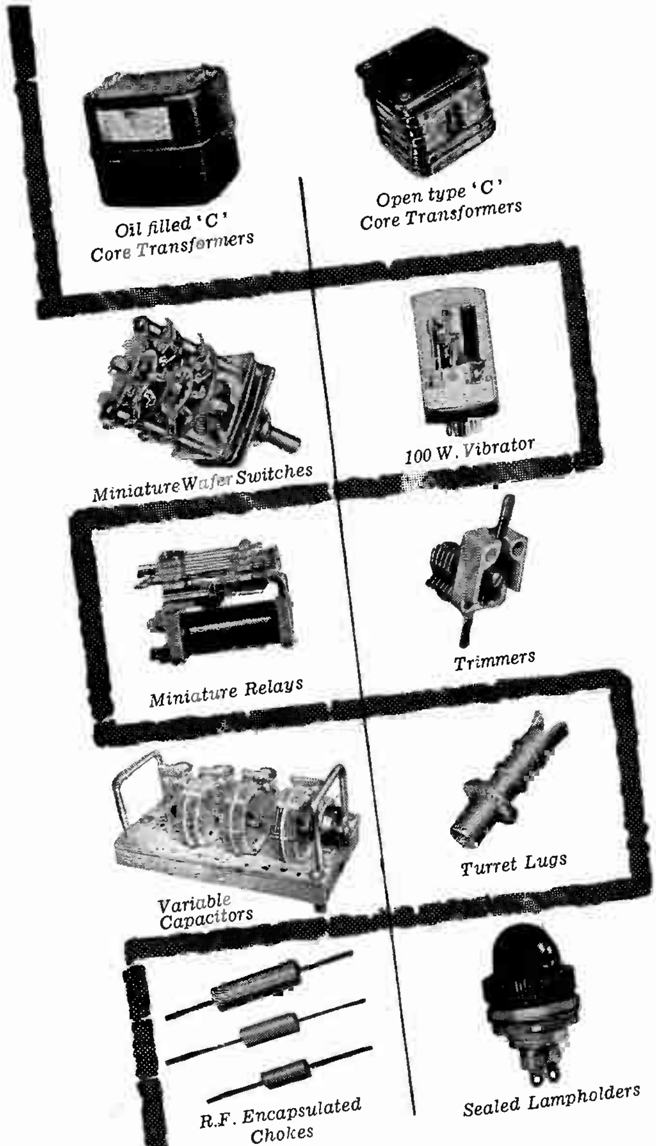

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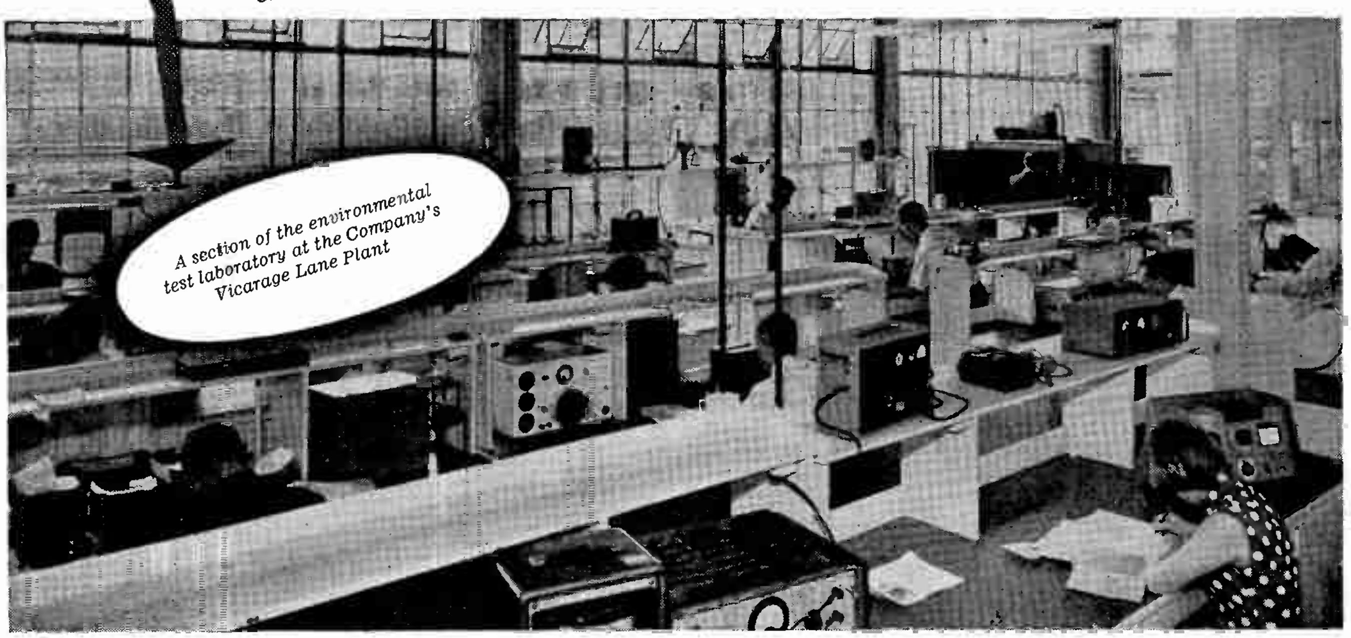
Plessey

In its various factories throughout the Country, Plessey produces an immense range of Type Approved electronic components. The Company supplies much of the material for vital communications equipment and major research projects, which can range from human voice synthesis to rockets.

Modern electronics technique demands a degree of precision and accuracy rivalled in few other fields—with brains and manufacturing facilities to match. All this is aimed at utter reliability; no component must fail in service. Plessey are proud to be making such a contribution to an exacting science.

Plessey Type Approved components are available to Industry in many forms. Some are illustrated here. To give more detailed information a series of publications exists, and will gladly be sent on receipt of a letter stating your interests.



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CP2

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(Regd. Trade Mark)

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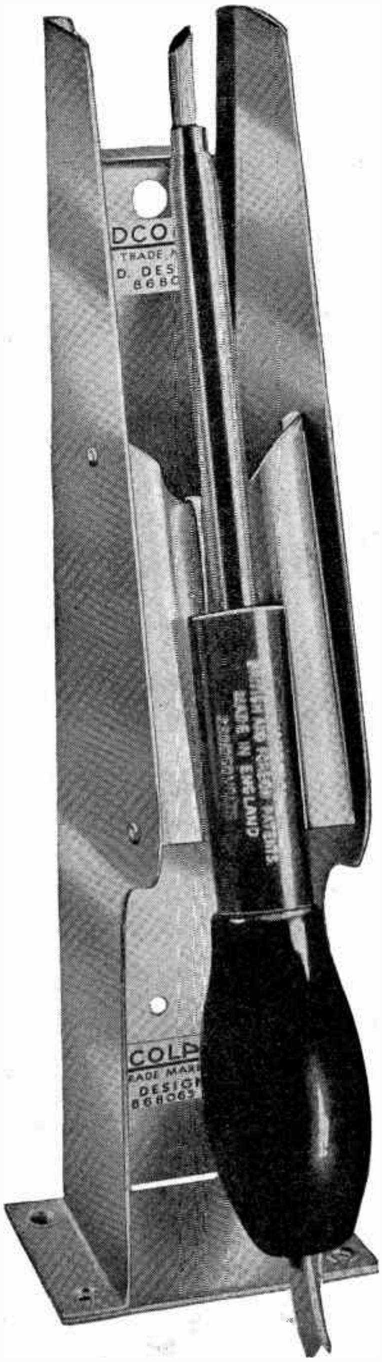
and Equipment



Comprehensive Range of Models

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Solder Dipping Pots

Supplied in ALL VOLT RANGES

A PRODUCT FOR PRODUCTION

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List No. 68

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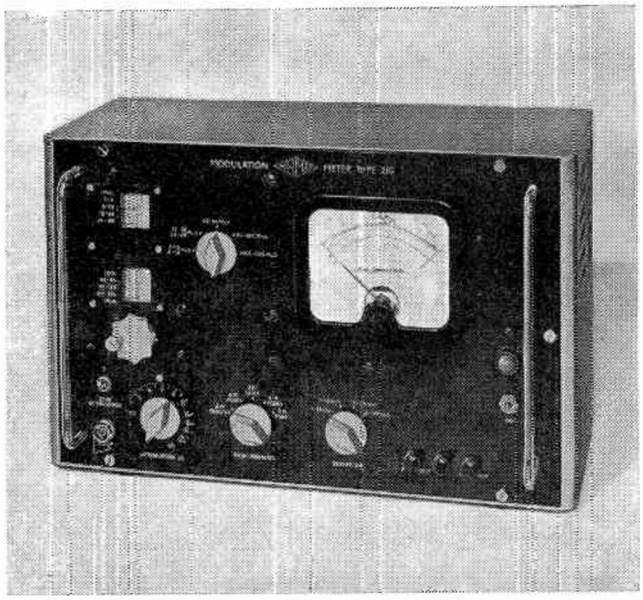
Measuring Modulation?



AIRMEC

MODULATION METER

measures

- *percentage modulation of amplitude Modulated Signals*
- *peak deviation of frequency Modulated Signals*

THE MODULATION METER Type 210
may be used to measure the percentage modulation of amplitude modulated signals and the peak deviation of frequency modulated signals in the carrier frequency range 2.25 Mc/s to 300 Mc/s.

Outputs at the intermediate frequency of 750 kc/s and at low frequency are available from terminals on the front panel. These outputs enable the modulated envelope of the input signal and the demodulated signals to be observed on an oscilloscope.

The limiting action of the instrument is so effective that it can be used to measure spurious frequency modulation occurring on amplitude modulated signals. Furthermore, changes of mean carrier level when amplitude modulation is applied can be measured to an accuracy of better than ±1%.

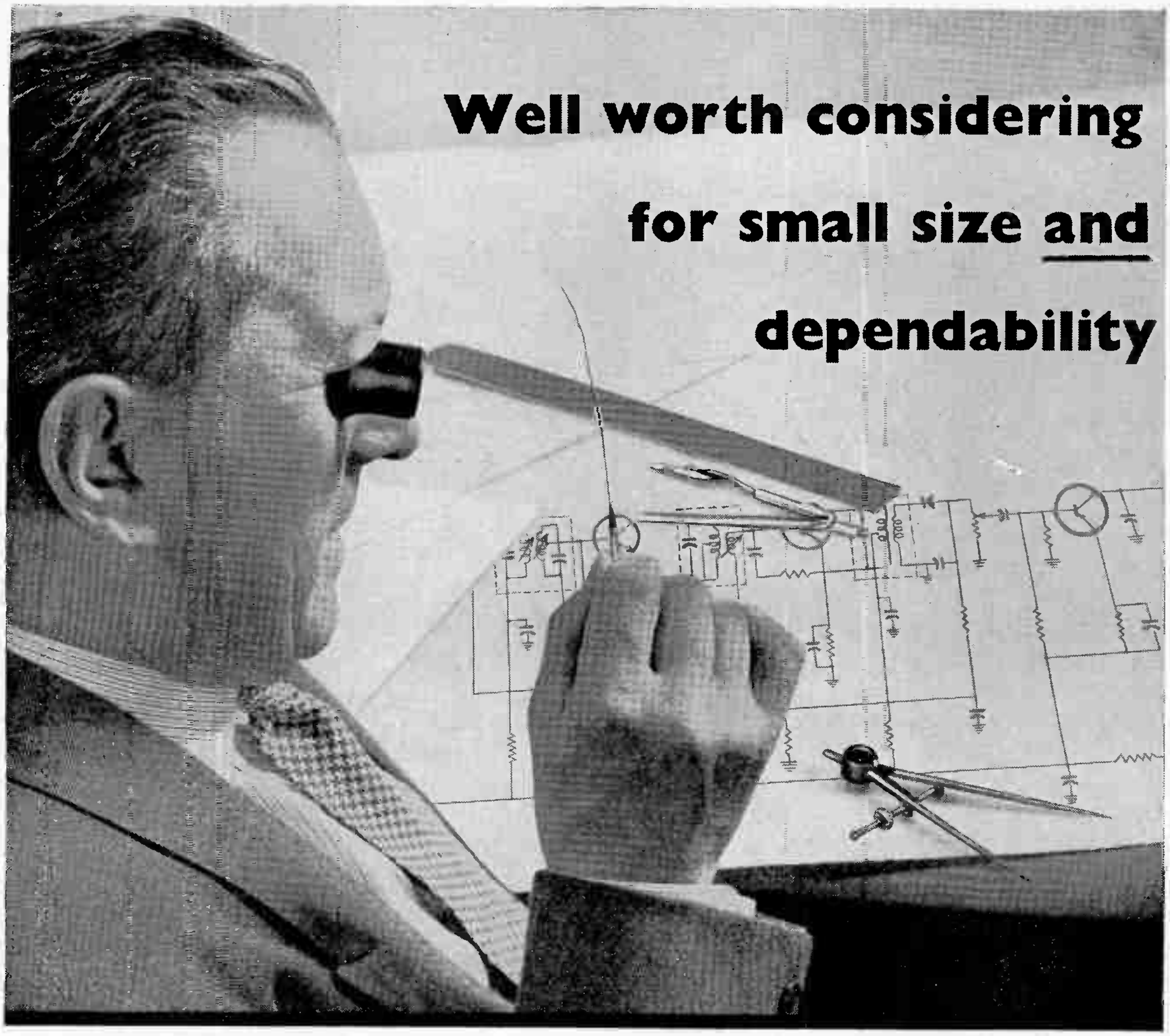
Simplicity of Operation
One of the most attractive features of the instrument is its simplicity of operation. The tuning control is adjusted until a meter reading is obtained, and the input attenuator adjusted for full-scale deflection. It is then only necessary to switch to the A.M. or one of the F.M. positions to obtain a direct reading of modulation.

Statistics of Airmec Modulation Meter
★ Frequency range . . . 2.25 to 300 Mc/s in 7 bands (up to 600 Mc/s with reduced sensitivity)
★ Input level . . . A.M. 7-700 mV
F.M. 7 mV-10 V (3 kc/s-100 kc/s)
50 mV-10 V (0-3 kc/s)
★ Modulation Frequency range . . . 30 c/s-15 kc/s
★ A.M. Range . . . 0-100% ±3%
★ F.M. Range 0-100 kc/s ±5%

DELIVERY—EX STOCK

Airmec | **AM/FM Modulation Meter Type 210**

AIRMEC LIMITED · HIGH WYCOMBE · BUCKS.
Telephone: High Wycombe 2060



The new type BTH Germanium Point Contact Rectifiers—

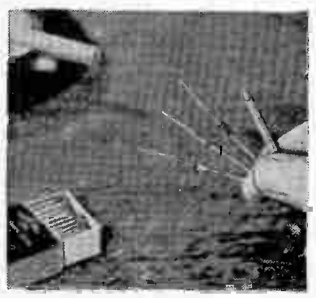
Only ¼ in. long, yet their miniature size is combined with high performance and complete dependability! They offer the following outstanding characteristics:

- HIGH TEMPERATURE STABILITY
- ABILITY TO WITHSTAND TROPICAL CONDITIONS
- SMALLER DIMENSIONS
- VERY LONG LIFE

RATINGS: CONTINUOUS OPERATION AT 25°C. (77°F.)

TYPE	PEAK INVERSE VOLTAGE† V	MAX. INPUT CURRENT mA	MAX. RESISTANCE at + 1 volt ohms	MIN. RESISTANCE at − 50 volts kilohms
CV 448*	80	30	333	500
CG41–H	65	30	250	50
CG42–H	100	30	500	1,000
CG44–H	80	30	333	500
CG50–H	100	30	500	200

*Type CV 448 has been granted 'type approval'. †Corresponds to 1·2 mA inverse current.



BRITISH THOMSON-HOUSTON

THE BRITISH THOMSON-HOUSTON CO·LTD·LINCOLN·ENGLAND

an A.E.I. Company

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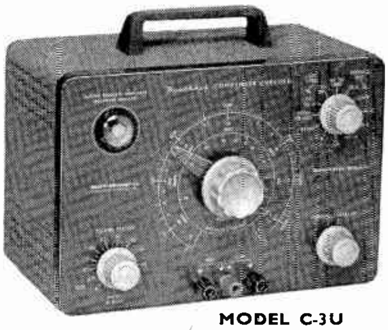
NEW Heathkit **'BUILD-IT YOURSELF' KITS**

CAPACITOR/RESISTOR TEST BRIDGE



MODEL C-3U

£7.19.6 Delivered free in U.K.
Deferred terms available

This is a most useful instrument in any laboratory or workshop. It is a self-contained AC-powered bridge which quickly measures resistors and a very wide range of all types of capacitors—determining capacity, power factor and indicating the degree of leakage. A Dual-Sensitive Magic Eye is used as the indicator of bridge balance. A unique safety-factor is the use of a special spring-return test switch. This automatically discharges the condenser after test so as to eliminate completely the risk of D.C. shock to the operator.

MEASUREMENT RANGES:
Capacity: 0.00001 μF (10 pF) to 1000 μF, in 4 ranges.
Resistance: 100 Ω to 5 MΩ, in 2 ranges.
Polarising Voltages: 5, 150, 250, 350 (D.C.).

OTHER HEATHKIT MODELS SUPPLIED AS KITS, delivered free in U.K.—(deferred terms if desired):

S-88 HI-FI STEREO AMPLIFIER KIT, 16 watts output (8 per channel). 0.1% distortion at 6 W/channel. Inputs for Stereo or Monaural—gram., radio and tape recorder. Many special features, including ganged controls—Ultra-linear push-pull output, with multi-loop N.F.B. Beautifully styled metal case, 2-tone grey. **£25.5.6**

S-33 POPULAR-PRICED HI-FI STEREO AMPLIFIER KIT. 6 watts output, (3 per channel). 0.3% distortion at 2.5 W/channel. 20 dB N.F.B. **£11.8.0**

USP-1 HI-FI STEREO PRE-AMPLIFIER. Ideal for low-output Hi-Fi P/U's. Switched inputs,—tape, gram, mic.: stereo/monaural. **£5.19.6**

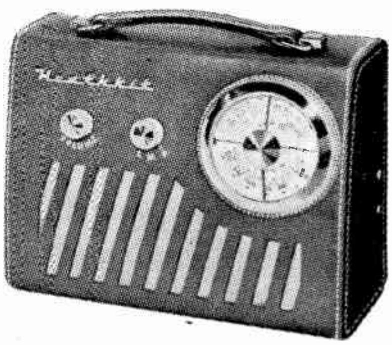
SSU-1 HI-FI SPEAKER SYSTEM KIT. A low-cost twin speaker system for Stereo or Monaural amplifiers. Ideal for small or medium-size rooms. Frequency range, 40-16,000 c/s. 8" high-flux bass unit and 4" wide-diffusion treble. Ductive reflex cabinet, supplied "in the white", pre-cut and sanded, for final finish to choice. **£10.9.0.** (Legs optional, £1.7.0 extra).

DX-40U 'HAM' TRANSMITTER KIT. Covers all Amateur bands, 80-10m. Power Inputs 75 Watts CW, 60 W. pk. c/c. phone. 40 W. to Aerial. Provision for V.F.O. Pi network output coupling. 3 switch-selected crystal positions. 6146 final amp. Filters minimise TVI. **£29.10.0**

VF-1U VARIABLE FREQUENCY OSCILLATOR KIT. Doubly screened, stabilised, covers 160, 80, 40, 20, 15 and 10 metres. Ideal for use with the Heathkit DX-40U (or similiar) transmitter. **£10.12.0**

UJR-1 YOUNG EXPERIMENTER'S KIT. An ideal introduction to radio. Any youngster will enjoy building this unusually sensitive and selective, dual-wave transistor radio, with 'High-Q' Ferrite coils. **£2.19.6**
UJR-15 Additional Amplifier Stage for this kit. **16s. 6d.**

HEATHKIT MODEL UXR-1 KIT
6-Transistor, Dual-Wave Portable

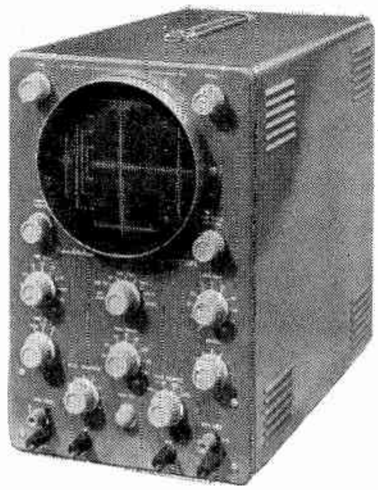


MODEL UXR-1

£17.17.0 (inc. £5.11.0 P.T.) Delivered free in U.K.
Deferred terms available

The very attractive solid leather case of this elegantly styled portable radio set, and its exceptionally good performance, excite admiration everywhere. Its 7" x 4" high-flux speaker gives an unusually rich and clear reproduction, although the set is small and light enough to be carried anywhere. A High-Q Ferrite rod aerial gives good selectivity and sensitivity. Being completely self-contained this is an ideal second set for the home, and most economical to run; a 3s. 6d. battery lasts about 6 months with average use. Not taking its power from the mains supply it is particularly safe in operation. The UXR-1 performs well in a car, too.
Printed-circuit board and pre-aligned I.F.T's permit assembly in 4 to 6 hours and you then have a delightful portable in the 25-30 guinea class.

Assemble the model O-12U 5" Oscilloscope Kit and you'll have a large-screen laboratory instrument at utility 'scope price. Exceptionally wide-range time-base (10 c/s to over 500 kc/s.). 'Y' frequency range 3 c/s to over 5 Mc/s, sensitivity 10 mV. RMS/cm. 'X' sensitivity 0.12 V. RMS/cm. Electronically stabilized and fused power supply. Output for Wobbulator. Voltage Calibrator (1 volt pk. to pk. source). Instantaneous shift, drift free. Automatic lock-in synch. Rise time 0.08 μsecs. or less. 'Z' modn. facilities. 2 gold-plated printed-circuit boards and pre-cut cabled wiring harness greatly facilitate assembly. Easy-to-follow 48-page Instruction Manual gives 'Step-by-Step' procedure. Ideal for Educational establishments, laboratories and service.



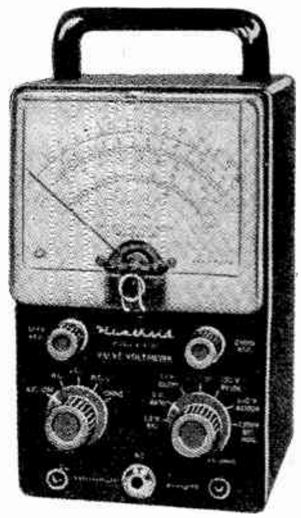
MODEL O-12U

£34.15.0 Delivered free in U.K.
Deferred terms available

EVERY HEATHKIT MODEL WILL ACHIEVE OR BETTER ITS ADVERTISED PERFORMANCE, IF BUILT IN ACCORDANCE WITH THE INSTRUCTION MANUAL 'STEP-BY-STEP' PROCEDURE

Because of the accuracy, reliability and sheer value of the model V-7A it is outselling all other VVM's. The precision and quality of its first-class components cannot be duplicated elsewhere at this price. Indication is by a large 4½" 200μA meter clearly calibrated for all ranges. The voltage divider uses 1% precision resistors. A gold-plated printed-circuit board simplifies the assembly, saves time and eliminates the possibility of wiring errors. It also ensures duplication of laboratory performance.

This multi-function VVM measures AC Volts (RMS and pk-to-pk.) DC Volts and Resistance. The 7, AC (RMS) and DC ranges are 1.5, 5, 15, 50, 150, 500 and 1,500. The 7, AC pk. to pk. Voltage ranges are 0-4, 14, 40, 140, 400, 1,400 and 4,000. Input impedance is 11 MΩ. Seven Ohm-meter ranges have multiplying factors of X1, X10, X100, X1000, X10K, X100K and X1MΩ. Centre-scale resistance readings are 10, 100, 1,000, 10K Ohms, 1 MΩ and 10 MΩ. A centre-zero dB scale is provided also for measuring audio amplifier performance. Test leads, prods and battery are included in the kit.



MODEL V-7A (KIT) VALVE VOLTMETER

£13.0.0 Delivered free in U.K.
Deferred terms available

ENGINEERING EXECUTIVES: FOR MANY OF YOUR REQUIREMENTS IT WILL PAY YOU TO BUY HEATHKIT TEST INSTRUMENTS IN KIT FORM AND HAVE THEM ASSEMBLED BY YOUR JUNIOR ENGINEERS.
THE INSTRUCTION MANUAL FOR EACH KIT GIVES CLEAR 'STEP-BY-STEP' PROCEDURE FOR ITS ASSEMBLY, ALIGNMENT AND OPERATION.
ONLY FIRST-CLASS COMPONENTS, MATERIALS AND PROCESSING ARE USED THROUGHOUT.

Further information regarding any of these Heathkit models will gladly be sent on request

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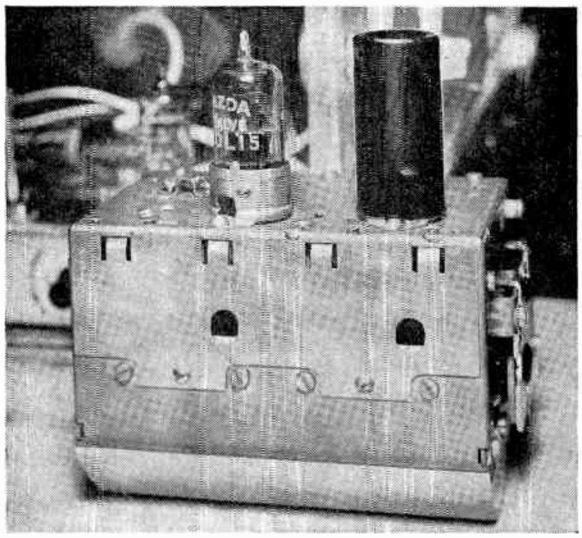
A member of the Daystrom Group, manufacturers of

THE LARGEST-SELLING ELECTRONIC KITS IN THE WORLD

Introducing a New High Performance Cascode Amplifier

For the information of set designers we are publishing details of several of our valves for TV circuits. If you are a TV manufacturer we shall be pleased to supply full technical details of any of these valves, together with valves for testing, on receipt of your enquiry. Reprints of advertisements dealing with other valves, mainly in our 'First Preference' Range of individual 0.3 amp heater valves, are available on request.
The valve dealt with in this advertisement is the Type 30L15, a new high performance cascode amplifier.



EDISWAN MAZDA 30L15

The 30L15 is a V.H.F. twin triode specially designed for use as a cascode amplifier in television tuners where a high gain and good noise performance are required. The 30L15 will provide a gain improvement of 4 to 5 dB and a noise factor improvement of 2 dB as compared with the 30L1.

Heater Current (amps)	I_h	0.3
Heater Voltage (volts)	V_h	7.0

MAXIMUM DESIGN CENTRE RATINGS

Anode Dissipation, either section (watts)	$P_{a(max)}$	2.0
Cathode Current, per section (mA) ...	$I_{k(max)}$	16.0
Anode Voltage (volts)	$V_{a(max)}$	250
Negative Grid Voltage (volts)	$V_{g(max)}$	−50
Grid to Cathode Resistance, section 1 (kΩ)	$R_{g'-k'(max)}$	500
Grid to Cathode Resistance, section 2 (kΩ)	$R_{g''-k''(max)}$	22*
Effective Grid to Earth Resistance, section 2 (kΩ)	$R_{g''-E(max)}$	150**

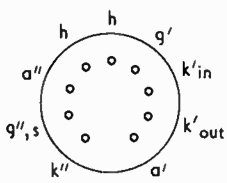
* *Grid current bias.*
** *With potentiometer bias from anode supply.*

INTER-ELECTRODE CAPACITANCES (pF)†

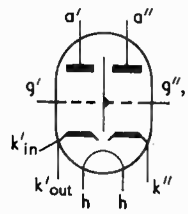
Grid 1 to Anode 1	$c_{g'-a'}$	1.5
Anode 1 to Cathode 1, Heater, Shield ...	$c_{a'-k',h,s}$	1.9
Grid 1 to Cathode 1, Heater, Shield ...	$c_{g'-k',h,s}$	3.1
Anode 2 to Cathode 1, Heater, Shield ...	$c_{a''-k',h,s}$	3.6
Grid 1 to Anode 2	$c_{g'-a''}$	0.0058

† *Inter-electrode capacity with holder capacity balanced out but with cylindrical screen can.*

BASE:
NOVAL—B9A



View of free end.



MAXIMUM DIMENSIONS

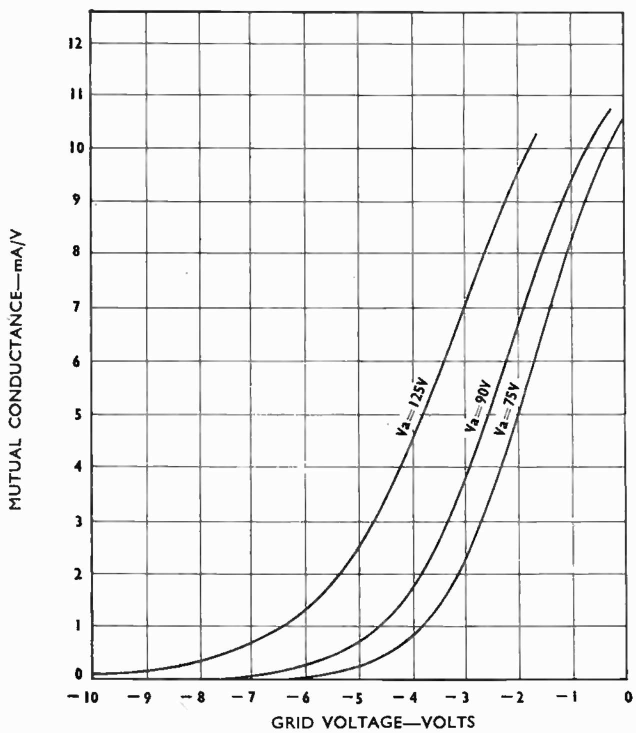
Overall Length (mm) 56 Seated Height (mm) 49
Diameter (mm) 22.2

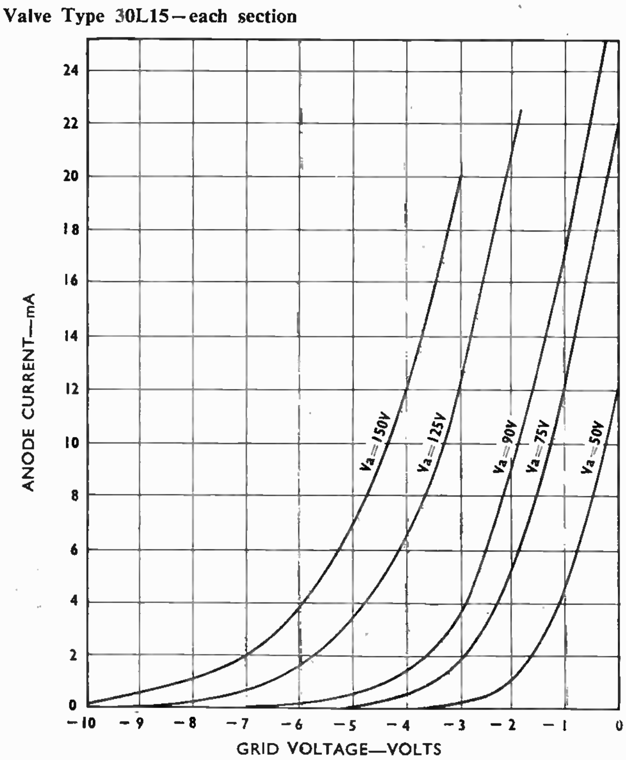
TYPICAL CASCODE OPERATION

Conditions given for circuit with the second section using potentiometer bias from the anode supply, making $V_{g''-E} = 99$ V.

Anode supply voltage (volts)	$V_{a(b)}$	200
Anode decoupling resistor, section 2 (kΩ)	$R_{a''}$	2.2
Anode current (mA)	I_a	15.3
Self bias resistor, section 1(Ω)	$R_{g'}$	100
Mutual conductance, section 1 (mA/V)...	g_m	8.8
Combined mutual conductance (mA/V)		8.5
A.G.C. Voltage (volts) to give $\Delta I_a \Delta V_{g'}$ = 0.1 mA/V	$V_{g''}$	−7
Input capacity working (pF)	$C_{in(w)}$	6
Change in input capacity by biasing to cut-off (pF)	ΔC	1.2

Characteristic Curves of Ediswan Mazda Valve Type 30L15—each section





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ELECTRONIC & RADIO ENGINEER

incorporating WIRELESS ENGINEER

HUGH S. POCOCK M.I.E.E. *Managing Editor*
W. T. COCKING M.I.E.E. *Editor*

CONTENTS

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ELECTRONIC &
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VOLUME 36 NUMBER 4

APRIL 1959 *incorporating* WIRELESS ENGINEER

'Carriers'

OUR readers have probably noticed that we are interested in clarity of expression in technical literature. Perhaps we are so interested because we so often find that difficulties of understanding turn out to be caused by some obscurity of expression or, even, to an actual misuse of words.

We have recently noticed a particular example of this which we think is worth commenting upon in some detail. We refer to the use of the term 'carrier' in the semiconductor field. We have been looking at the books. One starts off by defining, correctly in our view, a *charge carrier* as an electron or hole but, later, uses 'current carrier' as if it were synonymous. Another starts with 'current carrier' and appears to mean by it just 'charge carrier'! Still another has 'carrier', all by itself, and never says what it means.

Now, in fact, an electron or a hole carries a charge and it is the movement of the charge which constitutes a current. It is correct to speak of electrons and holes as charge carriers but not as current carriers. A current carrier is surely a material body, conducting or semiconducting material itself.

If 'carrier' is used to refer to current as well as to charge, there can be much confusion. When we say that electrons are the majority carriers, do we mean that the electron current is greater than the hole current or that there are more electrons than holes involved in the constitution of the current? The two things are not the same for, in general, electrons move faster than holes. We can imagine conditions in which there are more holes than electrons taking part in the current and yet, because of the greater velocity of the electrons, the electron movement constitutes the greater part of the current. If 'carrier' refers to charge, we should then say that the electrons are minority carriers but, if it refers to current, they are the majority carriers!

This well illustrates the kind of confusion which can arise in the minds of a beginner and which could so easily be avoided by just a little more care on the part of the writer.

Cathode Compensation

USE WITH PENTODE VIDEO STAGE

By H. D. Kitchin,*

In recent years, the use of cathode compensation in the video stage of a television receiver (see Fig. 1) has become an almost standard feature of British television receiver design. There are many reasons for the use of this type of high-frequency compensation, and it is proposed first of all to enumerate some of the most important of these.

A fundamental factor in the operation of the video stage is the choice of polarity of the video drive to the c.r.t. Of the two possibilities (i.e., negative- or positive-going video), negative-going video has been found to give the best balance between economy and performance. Some of the reasons for this are:

(a) The synchronizing-pulse separator can be of the simple grid-limiting type employing d.c. restoration at the (positive) tips of the synchronizing pulses and using the grid and cathode of the separator valve as a diode. The large signal available at the output of the video stage ensures very good elimination of the video signal down to a low level of contrast.

(b) The negative d.c. voltage developed as a result of the d.c. restoring action at the sync separator grid [as mentioned in (a)] is of a suitable magnitude to be used to provide a simple and cheap mean-level a.g.c. system.

(c) Failure of the video output valve does not result in the flow of heavy c.r.t. grid current, as it does with a positive video drive, when directly coupled to the c.r.t.

If we are to use a negative-going video drive then the input to the video stage (or video output stage if more than one stage is used) must be positive-going. Usually, this input will be obtained directly from the detector stage and it is desirable to retain the full d.c. component[1]. With the British system of positive modulation, the required positive-going video input is obtained from a rectifier giving a positive-going output. Therefore, if direct coupling is employed, some form of negative bias is required to avoid the flow of grid current and allow the valve to operate over the most useful part of its characteristic. There are only two possible methods of obtaining the bias: by means of resistance in the cathode circuit, or from some external source.

* Mains Radio Gramophones Ltd.

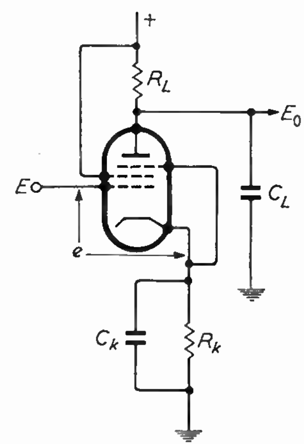


Fig. 1. *Practical circuit of a cathode-compensated pentode video amplifier.*

The first method has many advantages such as simplicity, cheapness and stabilization of the valve operating point. Unfortunately, it has one major disadvantage, the reduction in gain due to the negative feedback across the cathode impedance.

Although the second method (whereby bias is obtained from an external source) will eliminate this undesirable loss in gain, there are many difficulties to overcome. The main difficulty is in finding a suitable source of bias in a television receiver which has good regulation and a low internal impedance. These are essential requirements because, in the first case, any variation in the bias voltage is fully amplified by the video stage and will be apparent as a change in the brightness of the picture and, in the second, a high internal impedance will enable an appreciable shift in the operating point to occur if grid current flows in the video stage on interference pulses. If a long time-constant is associated with the high internal impedance, as is usually the case, there will be a reduction in brightness for a period after an interference pulse. Taking these and other factors into account, the use of cathode-resistance bias is very desirable and we must next consider how to avoid the loss in gain due to degeneration.

The simplest approach is to use a resistor bypassed with a large capacitance, following common a.f. practice. When dealing with the video waveform, however, a very large capacitance is required (e.g., 500 μF) if defects in the low-frequency response are to be avoided. Also, the partial removal of the d.c. component, due to degeneration still being present at d.c., will cause the operating point to shift as the mean level of the video signal changes. If the shift is sufficient to take the synchronizing pulses into the curved region of the valve characteristic near cut-off, severe compression of the synchronizing pulses will occur and may lead to synchronizing difficulties.

An alternative method of reducing the degeneration due to the cathode resistor is to employ a metal rectifier as the bias resistor. When operated at a suitable point on its forward characteristic the a.c. resistance is much less than the d.c. resistance, and the feedback is appreciably reduced. The a.c. impedance is almost independent of frequency down to zero and, therefore, enables the full d.c. component of the signal to be retained. However, the a.c. resistance of such a rectifier can never be zero and there must always be a residual loss of gain.

With both the preceding methods, the additional components are rather expensive and neither gives a really satisfactory result.

A more elegant method for restoring the gain when a cathode resistor is used for bias is to bypass the resistor for high frequencies only, when it is possible by a correct choice of bypass capacitor and anode load resistor to attain the same gain and bandwidth as that obtainable with zero cathode impedance. The loss of high-frequency response (normally entailed by the high value of anode load necessary to give the required gain), is offset by the reduced degeneration at high frequencies brought about by the cathode bypass capacitor. Although not of great importance in domestic television receivers, the cathode degeneration appreciably improves the linearity of the stage.

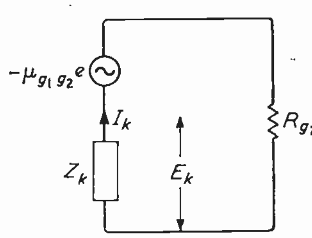


Fig. 2. *Equivalent cathode circuit for circuit of Fig. 1*

It has been shown[2,3] that if we make the anode circuit and cathode circuit time-constants ($C_L R_L$ and $C_k R_k$ respectively) equal, the performance of the stage is identical with that of the undegenerative stage, providing that the anode load consists of only parallel resistance and capacitance. Unfortunately, it is not possible to obtain an appreciable improvement in the transient response by using inductive components in the anode circuit. This is because the time-constant of the anode circuit is much longer than is the case when there is no cathode degeneration and, therefore, the usual design equations for inductive compensation are inapplicable. However, if we make $C_k R_k > C_L R_L$, the step response exhibits an overshoot and an improvement in the rise-time is obtained. The relationship between the overshoot and rise-time with increasing $C_k R_k$ is not a simple one and depends on the degree of feedback $(1 + g' R_k)$ present. General curves for the step response and steady-state response do not appear to have been published, and it is the purpose of this article to derive the necessary design equations and plot curves showing the relationship between the various aspects of the performance of the stage and, in particular, to show the improvement in rise-time and bandwidth possible if some overshoot in the step response is permitted.

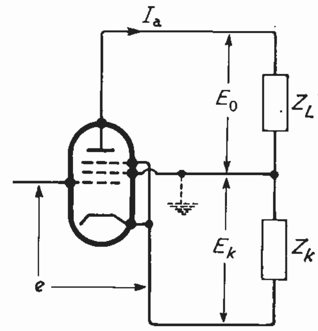


Fig. 3. *Equivalent anode circuit for circuit of Fig 1*

Analysis

The analysis of the circuit of Fig. 1 will be done by expressing the response function in terms of the general complex variable p. By this means, the step and steady-state responses become special cases of the general equation. The step response is obtained by the use of the Laplace transform when p is real, and the steady-state response by direct substitution for $p = j\omega$; i.e., p imaginary.

The gain of the stage of Fig. 1 is shown in Appendix 1 to be

$$\frac{E_0}{E_{in}} = \frac{g_m Z_L}{1 + g' Z_k} \quad \cdots \quad (1)$$

where Z_L and Z_k are the anode and cathode impedances respectively, g_m is the anode-to-grid mutual conductance and $g' = g_k (1 + 1/\mu_{g_1 g_2})$. In the case under consideration, the anode and cathode impedances each consist of parallel R and C.

$$\therefore Z_k = \frac{R_k}{pC_k R_k + 1}$$

$$\text{and } Z_L = \frac{R_L}{pC_L R_L + 1}$$

We then have

$$\frac{E_0}{E_{in}}(p) = \frac{g_m R_L}{B} \cdot \frac{pC_k R_k + 1}{(pC_L R_L + 1)\left(1 + \frac{pC_k R_k}{B}\right)} \quad (1a)$$

where $B = 1 + g' R_k$.

The equation will be normalized by the following substitutions:

$$a = C_k R_k / C_L R_L \; ; \quad \alpha = C_L R_L$$

$$\therefore \frac{E_0}{E_{in}}(p) = \frac{g_m R_L}{B} \cdot \frac{pa\alpha + 1}{(p\alpha + 1)\left(\frac{pa\alpha}{B} + 1\right)} \quad \cdots \quad (2)$$

This is the general-response function and we may now proceed to an examination of the special cases of step and steady-state responses.

Step Response

The input voltage E_{in} is assumed to be a unit step and will, therefore, have the Laplace transform $1/p$. Thus the transform of the output will be obtained from Equ. (2).

$$\mathcal{L}(E_0) = \frac{g_m R_L}{B} \cdot \frac{p\alpha + 1}{p(p\alpha+1)\left(\frac{p\alpha}{B}+1\right)} \quad .. \quad (3)$$

This may be expressed in partial fractions as

$$\mathcal{L}(E_0) = \frac{g_m R_L}{B}\left\{\frac{1}{p} + \frac{(a-1)\,B}{(B-a)\,(p+1/\alpha)} - \frac{a\,(B-1)}{(B-a)\left(p+\frac{B}{a\alpha}\right)}\right\}$$

By the use of the inverse transform[4] we find that the response to unit step is given as

$$E_0 = \frac{g_m R_L}{B}\left\{1 + \frac{(a-1)\,B}{(B-a)}\cdot e^{-t/\alpha} - \frac{a\,(B-1)}{(B-a)}\cdot e^{-Bt/a\alpha}\right\} \quad .. \quad (4)$$

and when $a = 1$, this reduces to

$$E_0 = \frac{g_m R_L}{B}(1 - e^{-Bt/\alpha}).$$

This expression shows that the effective time-constant of the anode circuit is α/B. Thus, the gain reduction due to cathode feedback is counterbalanced by the reduced effective time-constant of the anode circuit. When $a = 0$,

$$E_0 = \frac{g_m R_L}{B}(1 - e^{-t/\alpha})$$

and we can see that the gain is reduced by the factor B. The effective anode time constant, however, is that of the actual value present.

The solution given by Equ. (4) is valid for all values of a and B, providing $a \neq B$. The solution when $a = B$ may be found as follows:

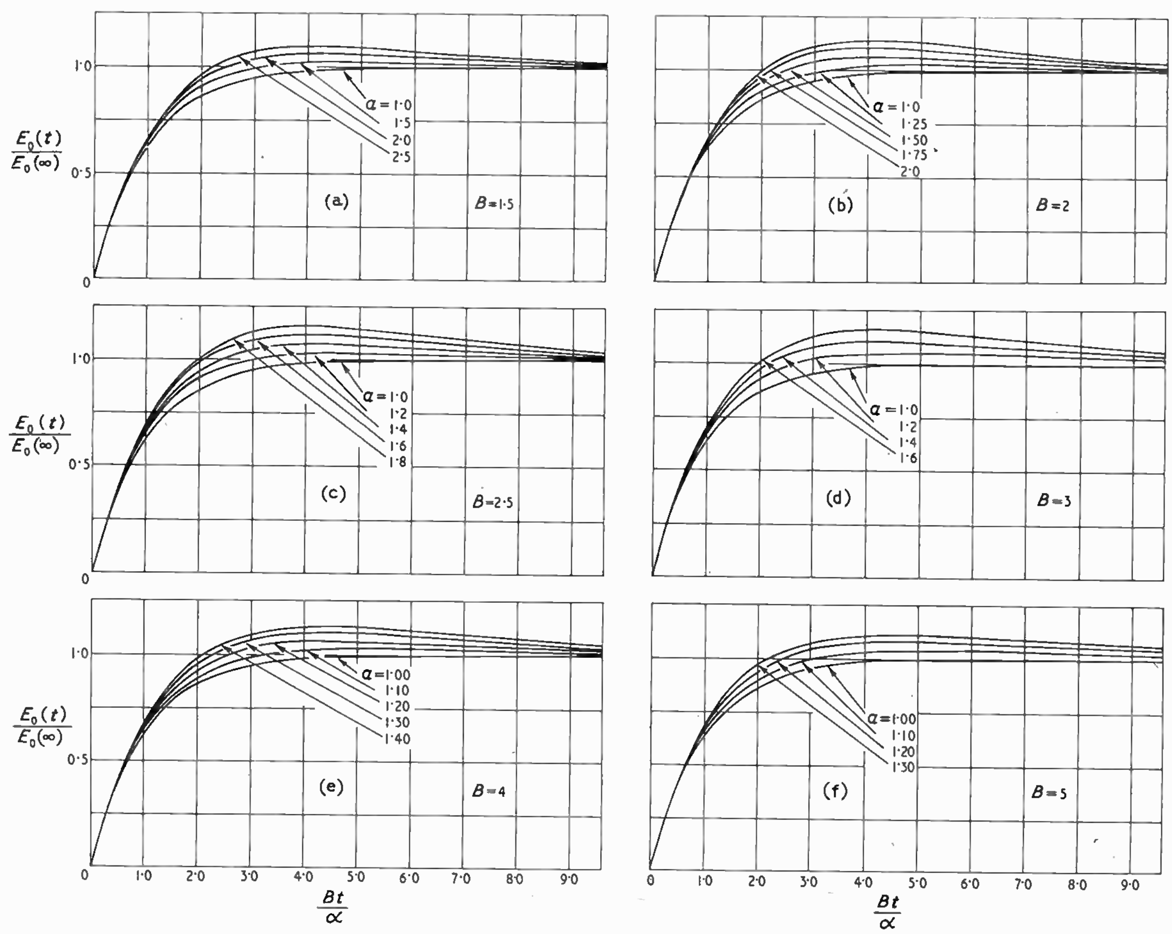
Referring to the fundamental transform of Equ. (3) and making the substitution $B = a$ we get

$$\mathcal{L}(E_0) = g_m R_L \cdot \frac{p\alpha + 1}{a(1+p\alpha)\,(1+p\alpha)} \cdot \frac{1}{p} \quad (5)$$

Putting into partial fractions,

$$\mathcal{L}(E_0) = \frac{g_m R_L}{a}\left\{\frac{1}{p} - \frac{p}{\left(\frac{1}{\alpha}+p\right)^2} + \frac{(a-2)}{\alpha\left(\frac{1}{\alpha}+p\right)^2}\right\}$$

Fig. 4. *Step response of cathode-compensated pentode video amplifier for various values of a and B;* $B = 1 + g' R_k$ *where* $g' = g_k\,(1 + 1/\mu_{g_1g_2})$; $a = C_k R_k/C_L R_L$; *and* $\alpha = C_L R_L$



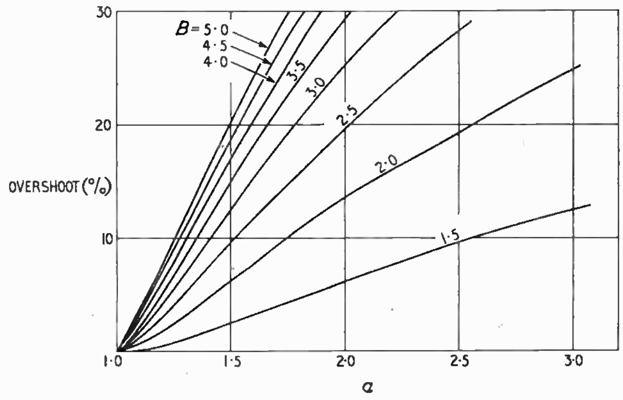


Fig. 5. *Percentage overshoot as a function of 'a' for cathode-compensated pentode video amplifier*

By the inverse transform, the step response (when $a = B$) is

$$E_0 = \frac{g_m R_L}{a}\left\{1 - e^{-t/\alpha}\,[1 - t/\alpha.\,(a-1)]\right\} \quad .. \quad (6)$$

The form of the output response, according to Equs (4) and (6) has been plotted in Figs 4(a) to (f) for various values of a and B. The values of a have been limited to those giving between zero and 10% of overshoot, as this is the range usually of interest. The two most useful aspects of the step response are the overshoot and the rise-time and we shall now proceed to evaluate these directly.

Overshoot

If we let the maximum value of Equ. (4) occur at a time t_m, then at this point the first differential with respect to t will be zero. Thus

$$\frac{d}{dt}\cdot f(p) = 0 - \frac{(a-1)B}{(B-a)}\cdot\frac{1}{\alpha}\cdot e^{-t_m/\alpha} + \left(\frac{B-1}{B-a}\right)\cdot\frac{B}{\alpha}\cdot e^{-Bt_m/a\alpha} = 0$$

whence $-(a-1)\,e^{-t_m/\alpha} + (B-1)\,e^{-Bt_m/\alpha a} = 0$

$$(a-1)\,e^{-t_m/\alpha} = (B-1)\,e^{-Bt_m/a\alpha}$$

$$\left(\frac{B-1}{a-1}\right) = e^{+\frac{t_m}{\alpha}\left(\frac{B}{a}-1\right)}$$

whence $\log_e\left(\frac{B-1}{a-1}\right) = \frac{t_m}{\alpha}\left(\frac{B}{a}-1\right).$

Therefore $$t_m = \frac{\alpha}{\left(\frac{B}{a}-1\right)}\cdot\log_e\left(\frac{B-1}{a-1}\right) \quad .. \quad (7)$$

Substituting this value for E_{in} in Equ. (4) we obtain

$$E_0 = \frac{g_m R_L}{B}\left\{1 + (a-1)\cdot\exp. - \frac{\log_e\frac{B-1}{a-1}}{B/a - 1}\right\} \quad (8)$$

The overshoot is, therefore,

$$(a-1)\cdot\exp. - \frac{\log_e\frac{B-1}{a-1}}{B/a-1} \quad .. \quad .. \quad .. \quad (9)$$

This is valid for all values of B and a other than when $B = a$. The overshoot when $a = B$ is obtained by the same process of differentiating and equating to zero starting from Equ. (6). This yields for t_m

$$\frac{t_m}{\alpha} = \frac{a}{a-1} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (10)$$

and for the overshoot

$$(a-1)\ .\ \exp. - \frac{a}{a-1} \quad .. \quad .. \quad .. \quad (11)$$

Curves showing the overshoot as a function of a for various values of B have been plotted in Fig. 5 using Equs (9) and (11).

Rise-Time

The generally accepted definition of the rise-time in terms of the time of transition from 10% to 90% of the ultimate steady value will be taken as the basis for calculation.

Evaluation of the rise-time requires the solution for t of the equation

$$m = 1 + \frac{(a-1)B}{B-a}\cdot e^{-t/\alpha} - \frac{a(B-1)}{B-a}\cdot e^{-Bt/a\alpha} \quad (12)$$

where m has the values of 0·10 and 0·90.

If we choose Bt/α as the variable, its value at the 10% point varies less than 0·6% from a straight line of unity slope, for all values of a greater than unity. We may, therefore, take $Bt/\alpha = 0{\cdot}1$ at the 10% point for all values of B and a.

The evaluation of the time at which the 90% point occurs requires the solution of the equation

$$\frac{a(B-1)}{B-a}\,e^{-Bt/a\alpha} - \frac{B(a-1)}{B-a}\cdot e^{-t/\alpha} = 0{\cdot}1$$

Putting $Bt/\alpha = x$, this becomes

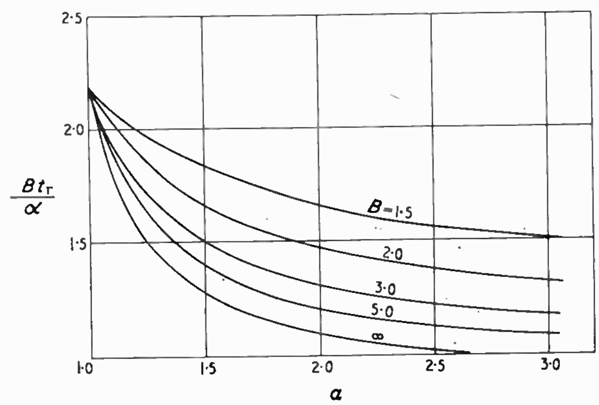
$$\frac{a(B-1)}{B-a}\cdot e^{-x/a} - \frac{B(a-1)}{a-B}\cdot e^{-x/B} = 0{\cdot}1$$

The simplest method of obtaining the solution of this equation is by successive approximation for given numerical values. This has been done and the curves of Fig. 6 drawn from the results.

Steady-State Response

The steady-state response is obtained by substituting

Fig. 6. *Rise-time t_r, between 10% and 90% points, as a function of 'a' for cathode-compensated pentode video amplifier*



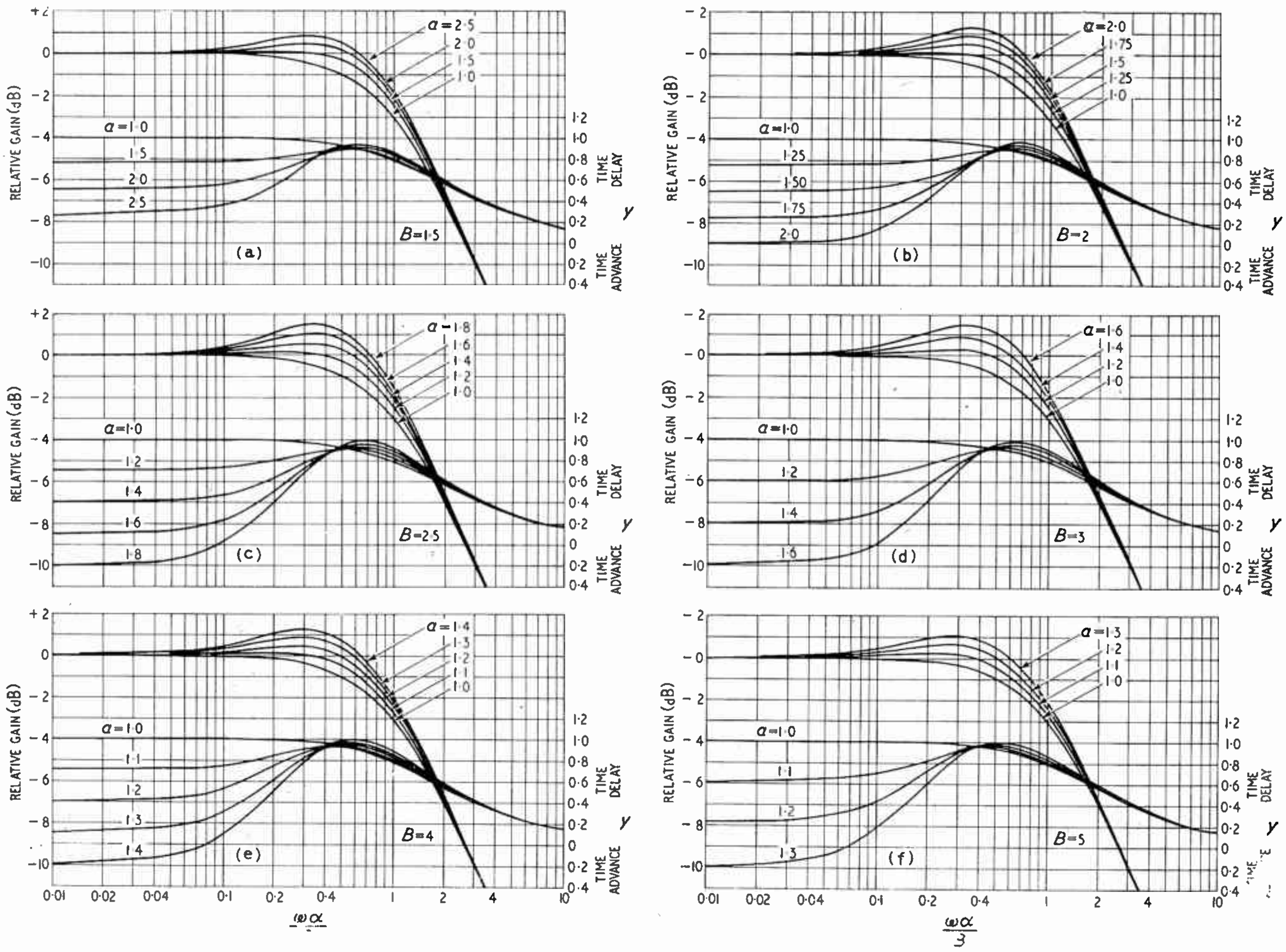


Fig. 7. *Amplitude response and time delay as a function of the frequency variable* $(\omega\alpha/B)$ *for a cathode-compensated video stage;* $B = 1 + g' R_k$ *where* $g' = g_k (1 + 1/\mu_{g_1g_2})$*;* $a = C_k R_k/C_L R_L$*;* $\alpha = C_L R_L$*;* $y = \phi/(\omega\alpha/B)$

$p = j\omega$ in Equ. (1a) when we obtain

$$\frac{E_0}{E_{in}} = \frac{g_m R_L}{B} \cdot \frac{j\omega C_k R_k + 1}{(j\omega C_L R_L + 1)\left(\frac{j\omega C_k R_k}{B} + 1\right)}$$

Making the following substitutions, as in the previous section,

$$\alpha = C_L R_L \quad a = C_k R_k / C_L R_L$$

we have

$$\frac{E_0}{E_{in}} = \frac{g_m R_L}{B} \cdot \frac{j\omega a\alpha + 1}{(j\omega\alpha + 1)\left(\frac{j\omega a\alpha}{B} + 1\right)} \quad \text{.. (13)}$$

The amplitude response is given by the modulus of this expression and will be

$$\left|\frac{E_0}{E_{in}}\right| = \frac{g_m R_L}{B} \sqrt{\frac{1 + a^2 (\omega\alpha)^2}{[1 + (\omega\alpha)^2]\left[1 + \frac{a^2}{B^2}(\omega\alpha)^2\right]}} \quad (14)$$

The phase angle ϕ is given by

$$\tan\phi = \frac{-(\omega\alpha)[1 + a(1/B - 1) + (a^2/B)\cdot(\omega\alpha)^2]}{1 + a\left(1 + \frac{a-1}{B}\right)(\omega\alpha)^2} \quad (15)$$

The two aspects of amplitude and phase response will now be examined in greater detail.

Amplitude Response

In order that the curves for different values of B may be directly comparable, it is necessary to choose $\omega\alpha/B$ as the frequency variable. The amplitude-response function of Equ. (14) thus becomes

$$\left|\frac{E_0}{E_{in}}\right| = \frac{g_m R_L}{B} \sqrt{\frac{1 + a^2 B^2 \cdot (\omega\alpha/B)^2}{[1 + B^2(\omega\alpha/B)^2]\,[1 + a^2(\omega\alpha/B)^2]}} \quad \text{.. (16)}$$

This expression clearly consists of the gain-multiplier factor $g_m R_L/B$, and the frequency-dependent function under the root sign. For the purpose of evaluating the variation of gain with frequency we may ignore the gain multiplier and express the amplitude function A as

$$A(\text{dB}) = 10\log_{10}\frac{1 + a^2B^2\,(\omega\alpha/B)^2}{[1 + B^2\,(\omega\alpha/B)^2]\,[1 + a^2\,(\omega\alpha/B)^2]} \quad \text{.. (17)}$$

Curves of this function, for the values of a and B used previously in evaluating the step response, are given in Figs 7(a) to (f).

The maximally-flat response is that giving the maximum bandwidth without the amplitude response rising above the low-frequency value; this is an important factor in the design of amplifiers handling sinusoidal waveforms. The condition for the maximally-flat response is obtained by equating to zero the first differential of the amplitude-response function with respect to the frequency variable. This process yields the frequency of the maximum as

$$(\omega\alpha)^2 = \frac{1}{a^2}\left[\pm\sqrt{1 - B^2\left(1 - \frac{a^2}{B^2} - a^2\right)} - 1\right] \quad (18)$$

The negative root is obviously inadmissible so that the condition for the existence of a maximum at a real frequency will be

$1 - B^2\,[1 + (a^2/B^2) - a^2] > 1$

i.e., $[1 + (a^2/B^2) - a^2] < 0$

or $a^2 > 1/(1 - 1/B^2)$ .. .. (19)

The condition for no maximum will be given by

$a^2 < 1/(1 - 1/B^2)$ .. .. (20)

and the condition for the maximally-flat response as

$a^2 = 1/(1 - 1/B^2)$ .. .. (21)

The value of a obtained from Equ. (21), is tabulated in Table 1 together with the approximate degree of overshoot present in each case.

TABLE 1

B	1·25	1·5	2·0	2·5	3·0	3·5	4·0	4·5	5·0	∞
a	1·67	1·34	1·154	1·092	1·06	1·043	1·032	1·026	1·020	1
Percentage of overshoot	1·38	1·30	1·2	1·1	1·0	0·9	0·8	0·75	0·7	0

From Table 1 we see that the condition for a maximally-flat response gives only a slight overshoot for the usual range of values of B.

Phase Response

From Equ. (15) the phase response is given by

$$\phi = \tan^{-1}\frac{-\,\omega\alpha\,[1 + a\,(1/B - 1) + (a^2/B)\,(\omega\alpha)^2]}{1 + a\,[1 + (a-1)/B]\,(\omega\alpha)^2} \quad .. \ (22)$$

The phase characteristic will be expressed in terms of the normalized time delay y which is given by

$y = \phi\,/\,(\omega\alpha/B)$ .. .. (23)

Curves of y against $\omega\alpha/B$, for the values of a and B previously employed in evaluating the step response, are shown in Figs 7(a) to (f).

Design Aspects

Examination of the curves of Figs 4 to 7 reveals some factors of considerable interest in the design of a video stage of the type being studied. One of the most striking factors is that a large variation of overshoot and rise-time is accompanied by only a slight change in the amplitude response. For example, when $B = 1{\cdot}5$ a change of a from 1 to 2·5 gives an increase in overshoot from 0% to 10% and a reduction of rise-time (Bt_r/α) from 2·2 to 1·55; i.e., a ratio of 0·7. The corresponding change in amplitude response at $\omega\alpha/B = 1$ is only 1 dB increasing to 1·5 dB at $\omega\alpha/B = 0{\cdot}5$. The practical significance of this is that if it is desired to deduce the step response by a measurement of the amplitude response then very great accuracy is required; an error of $\pm$ 0·5 dB, in the present case, would lead to a possible error of 7 or 8% in the overshoot and 30 or 40% in the rise-time.

It is apparent from the time-delay curves that it is variation in the time delay which is largely responsible for the change in step response and, in fact, these curves give a good indication of the circuit performance. Unfortunately, the time-delay curve is not a convenient quantity to measure practically, consequently it is of little use as a performance criterion.

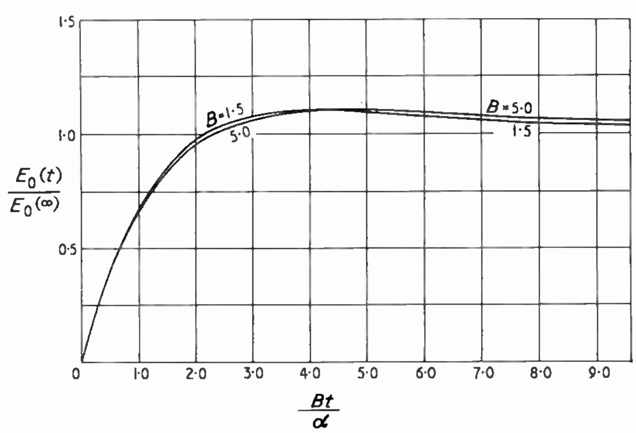
The influence of the value of the feedback factor B is quite marked, and the curves of Fig. 5 show how the rate of rise of overshoot with a is dependent on this factor. The value of a required to give a specified overshoot increases with decreasing B and, as the rate of change of overshoot with a decreases with decreasing B, the influence of a given tolerance on A (owing to component and circuit capacitance tolerances) decreases with B.

For example, with a 5% overshoot, $a = 1{\cdot}85$ when $B = 1{\cdot}5$, and 1·13 when $B = 5$

A 10% tolerance on a will give rise to variation in the overshoot from 3·7 to 6·5% for $B = 1{\cdot}5$, and from 0·8 to 9·5% for $B = 5$. The less critical nature of the compensation for low values of B is obvious.

A further point of interest in the step response concerns the general shape of the output waveform as a function of B. Direct comparison of the curves having an equal degree of overshoot for different values of B shows that the rise time and the duration of the overshoot are slightly shorter for the lower value of B. An example showing this is given in Fig. 8 where the step

Fig. 8. *Comparison of step-response functions for feedback factors of $B = 1{\cdot}5$ ($a = 2{\cdot}563$) and $B = 5{\cdot}0$ ($a = 1{\cdot}25$), when the overshoot is 10%.*



response for a 10% overshoot is given for both $B = 1{\cdot}5$ and 5, the shorter rise-time and shorter overshoot when $B = 1{\cdot}5$ being just perceptible.

The foregoing treatment has shown that the best performance from the cathode-compensated stage is obtained when a low value of the feedback factor is used, and it is essential, if this is to be realized in practice, that the cathode resistor is not fixed by bias considerations. Although this must be the case if the current through the bias resistor is solely due to the valve cathode current, we may overcome this limitation by employing an auxiliary 'bleed' resistor from (say) the h.t. line as is shown by the resistor R in Fig. 9. By this means, we may select the value of the cathode bias resistor as low as we please, the required value of bias being obtained by the appropriate value of 'bleed' resistor.

A consequence of this procedure is an increase in the total h.t. current drain and this will impose a limit on how far one may reduce B.

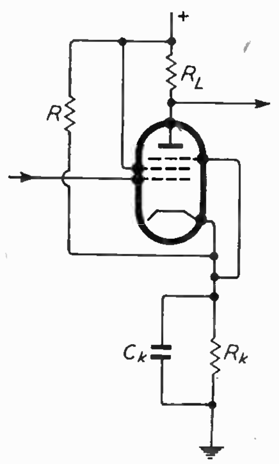


Fig. 9. *Circuit of a pentode video stage with cathode compensation showing the use of an auxiliary 'bleed' resistor R to provide correct bias voltage when R_k is fixed by feedback considerations*

Further factors influencing the value of B are:

(1) For a given performance, reduction of B requires a reduction in the value of the anode load R_L and this may limit the output voltage swing available.

(2) The cathode time-constant controls the amplitude and rate of fall of the input waveform that can be handled without cut-off occurring, and the situation is less favourable at high values of B.

(3) The degree of improvement in input/output linearity is proportional to B and there will be, therefore, little improvement with low values of B.

Conclusion

It has been shown that the cathode-compensated pentode video stage can give a considerable improvement in rise-time over that obtained with critical compensation ($a = 1$) if some overshoot is permitted. It has also been shown that the value of the feedback factor B has an influence on the step response and steady-state response when overshoot is present. This is in contrast to the case of critical compensation in which the step and steady-state performances are independent of B. Examination of the theoretical results shows that low values of B give an improved step-response shape and are less critical with regard to component and circuit tolerances.

Acknowledgement

The author wishes to acknowledge the management of Messrs. Mains Radio Gramophones Limited for permission to publish this article.

APPENDIX 1

The cathode equivalent circuit of Fig. 1 is shown in Fig. 2 from which we have

$$\mu_{g_1g_2}\cdot e = I_k\,(R_{g_2} + Z_k) \qquad (24)$$

and

$$E_k = I_k Z_k \qquad (25)$$

where R_{g_2} is the screen-grid plus anode-slope resistance; i.e., when the pentode is strapped as a triode.

Whence from Equs (24) and (25)

$$\mu_{g_1g_2}\cdot e = E_k(R_{g_2}/Z_k + 1) \qquad (26)$$

The equivalent anode circuit of Fig. 1 is shown in Fig. 3.

The anode current I_a is a function of both the control-grid and screen-grid voltages and is given by

$$I_a = g_m\,(e_{g_1} + e_{g_2}/\mu_{g_1g_2}) \qquad (27)$$

Now $e_{g_2} = -E_k$

therefore

$$I_a = g_m\,(e - E_k/\mu_{g_1g_2}) \qquad (28)$$

Substituting for E_k from Equ. (26) in Equ. (28) and multiplying by Z_L we obtain

$$Z_L\,I_a = E_0 = g_m\,Z_L\left(e - \frac{e}{R_{g_2}/Z_k + 1}\right) \qquad (29)$$

Now from Fig. 1, $E_{in} = e + E_k$ (30)

so that by substituting for E_k from Equ. (26) in (30) we have

$$E_{in} = e + \frac{\mu_{g_1g_2}\cdot e}{R_{g_2}/Z_k + 1}$$

whence

$$e = \frac{E_{in}}{1 + \dfrac{\mu_{g_1g_2}}{R_{g_2}/Z_k + 1}} \qquad (31)$$

Substituting this in the expression for E_0, Equ. (29), we obtain

$$\frac{E_0}{E_{in}} = g_m\,Z_L\cdot\frac{R_{g_2}/Z_k}{R_{g_2}/Z_k + 1 + \mu_{g_1g_2}} \qquad (32)$$

If we denote the slope when strapped as a triode by g_k (this will also be the cathode-current—grid-voltage slope) then

$$R_{g_2} = \frac{\mu_{g_1g_2}}{g_k} \qquad (33)$$

Thus Equ. (32) becomes

$$\frac{E_0}{E_{in}} = g_m\,Z_L\cdot\frac{1}{1 + g_k\,Z_k + g_k\,Z_k/\mu_{g_1g_2}}$$

$$= \frac{g_m\,Z_L}{1 + g'\cdot Z_k} \qquad (34)$$

where

$$g' = g_k\left(1 + \frac{1}{\mu_{g_1g_2}}\right) \qquad (35)$$

the term $1/\mu_{g_1g_2}$ is often $\ll 1$, and the expression then approximates to the well-known form

$$\frac{E_0}{E_{in}} = \frac{g_m\,Z_L}{1 + g_m\,Z_k} \qquad (36)$$

This expression is accurate if g_2 is decoupled to the cathode, but this practice is very rare in television video stages.

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WHAT GLASSTONE SAID IN '58

The first edition of Samuel Glasstone's "Sourcebook on Atomic Energy" was published in the United States in 1950, and in Britain, by Macmillan's, in 1952. For some six years it has been my regular stand-by as a reference book. It was, I remember, also recommended as suitable reading for Technical Reconnaissance Officers in the Civil Defence service; but I have never regretted the investment in a copy for my own private use. I think that the T.R.O. and the 1952 edition rather go together. You probably realize that the new status and duties of the scientist in Civil Defence, the Scientific Intelligence Officer, are rather different from those of the T.R.O. It's not that he needs to know less, but that he has to do such a great deal more. Real speed and accuracy in computation and decision are more important nowadays than being able to distinguish between the rontgen, the rep, and the rem. There is, then, a sort of parting of the ways. The second (1958) edition of Glasstone will put you among the aristocracy that can tell the difference between the rontgen, the rep, and the rem, *and also the rad*, which is one up on the Lord President of the Council. (The need for this discrimination was, indeed, mentioned by G. D. Smith in the September 1958, issue of this journal; a point on which, relying on the earlier edition of the "Sourcebook", I was simply not up to date.) But I somehow feel that even S.I.O.s would be well enough served by the earlier edition. The new work has a more theoretical or philosophical outlook, simply because that is the way in which physics has been developing during the past decade. And, although I shall not stick exclusively to what it says, and can only deal with one major section this time, I think it would be profitable to see what Glasstone said in 1958.

This 1958 edition appears to be a much more valuable compilation than its predecessor, which was one of a number of good reference books; for it is, as far as I know, the *only* book of this standard which takes the story of atomic physics right up to the beginning of 1958. If you ask what exactly "this standard" is, I can only answer the question obliquely. I should say that the "Sourcebook" is readable and intelligible and a storehouse of accurate information, and that in everything but syllabus slant and mathematical treatment it is somewhere about graduate level; but no great familiarity with modern physics is really needed in order to be able to tackle it. If you really understand what a volt is, that will see you a long way through it. The present edition gives full prominence to the new lines of work which can be grouped under the following headings: (1) Fundamental particles; (2) Fundamental theory, including such news-worthy topics as parity, and the less popularized ones of strangeness and isotopic spin; (3) New transuranic elements; (4) Progress in thermonuclear work; (5) Nuclear structure; (6) New techniques, such as the Čerenkov counter, the liquid scintillator, and the bubble chamber. One might, in fact, take these topics as the framework for a series of articles; and I have indeed got one on parity and one on Čerenkov radiation simmering on the hob. The present tale of fundamental particles, and the almost sudden change of ownership of some of them to the high energy workers from the cosmic-ray field does, however, command more immediate attention.

The New Fundamental Particles

Not all the particles named since 1950 or so are new discoveries. The earlier picture was of a system containing electron and positron, proton and neutron, and a number of mesons of mass intermediate between those of electron (m_e) and proton ($1836m_e$) which were first observed in cosmic radiation. The μ-mesons (mass about $206m_e$), the π-mesons (mass about $270m_e$), and what might have been *either* a rather large number of heavier mesons *or else* a few of them which enjoyed several alternative methods of decaying (the V-particles of Rochester and Butler), were well established by that time also. Hyperons of mass round about $2{,}000m_e$ had also been observed. What is new is the classification and naming of the particles, which brings out an orderly pattern not unlike that of the periodic table. In the first place, the discovery of the anti-proton and the anti-neutron (1955) confirmed the prediction of the newer theory of particle-pair creation as applied to particles other than Dirac electrons and, therefore, of its generality; and some of the mesons, as the μ^-- and μ^+-mesons, and the π^-- and π^+-mesons (Fig. 1) are simply the 'pro-' and 'anti-' members of a particle-pair. Secondly, with the development of bevatrons and proton-synchrotrons and things of that kind it has been possible to produce various massive particles artificially, and to study in the laboratory what had previously only been available as cosmic-ray 'events'; so that much more is known about the masses, life-times, and modes of decay of the various-particles. There now appears to be a total of 16 particle–anti-particle pairs, with two oddfellows or freemartins, the π^0-meson and the photon, making a total of 34 fundamental particles. A new discovery in the list is the neutrino—anti-neutrino pair; which is which in this case, and how they differ from one another, is a rather long story.

There seem to be two classifications in use. The first, concerned only with mesons, sorts them out as L-mesons (up to $283m_e$), K-mesons ($283m_e$ to $1836m_e$) and Y-mesons which are hyperons with mass between that of proton and deuteron ($1836m_e$ to $3670m_e$). This is the

CLASSIFICATION	SPIN, IN UNITS OF $h/2\pi$	TYPE OF PARTICLE	MASS, IN UNITS OF m_e	PARTICLES	ANTI-PARTICLES	LIFETIME (sec)	DECAY PROCESS OR PRODUCTS
HYPERONS	$\frac{1}{2}$	XI-PARTICLE	2,570	Ξ^0 Ξ^-	$\bar{\Xi}^+$ $\bar{\Xi}^0$	ORDER OF 10^{-10} sec OR LESS	"CASCADE"; $\pi + \Lambda^0$
		SIGMA-PARTICLE	2,330	Σ^- Σ^0 Σ^+	$\bar{\Sigma}^+$ $\bar{\Sigma}^0$ $\bar{\Sigma}^-$		p^+, n^0, π ; Λ^0
		LAMBDA-PARTICLE	2,182	Λ^0	$\bar{\Lambda}^0$		$p^+ + \pi^-$ & $n^0 + \pi^0$
NUCLEONS	$\frac{1}{2}$	NEUTRON	1,838	n^0	$\bar{n}^0$	10^3	$n^0 \rightarrow p^+ + e^-$ (BETA-DECAY)
		PROTON	1,836	p^+	p^-	STABLE	
MESONS	0	K-MESON	966	K^+ K^0	$\bar{K}^0$ K^-	10^{-2}	VARIOUS MODES YIELDING π, μ, e & ν^0
			965	K_1^0 K_2^0		$0{\cdot}95 \times 10^{-10}$; 9×10^{-8}	
		π-MESON	273	π^-	π^+	$2{\cdot}56 \times 10^{-8}$	$\mu^- + \nu^0$
			263	π^0		$< 4 \times 10^{-16}$	2γ; OR, $e^+ + e^- + \gamma$
LEPTONS	$\frac{1}{2}$	μ-MESON	206	μ^-	μ^+	$2{\cdot}2 \times 10^{-6}$	$e^- + 2\nu^0$
		ELECTRON	1	e^-	e^+	STABLE	
		NEUTRINO	0	ν^0	$\bar{\nu}^0$	STABLE	
PHOTON	1		0	γ			

THE INDICES +, −, 0 FOR PARTICLES & ANTI-PARTICLES, INDICATE CHARGE. THE BAR MEANS "ANTI"
CORRESPONDING MEMBERS OF PARTICLE PAIRS ARE AT EQUAL DISTANCES FROM THE BROKEN LINE, AS IF THEY WERE OBJECT & IMAGE FOR REFLECTION IN A MIRROR LYING ALONG THAT LINE

Fig. 1. *Fundamental particles and their properties. The last column refers to the decay of the 'particles' only; that of the corresponding anti-particles is, in most cases, of a similar type*

system used by Glasstone. On the other hand, that of Fig. 1, in which the four categories hyperon, nucleon, meson, and lepton (which means small or weak particle) embrace the lot, has a certain symmetry. Spin considerations, as well as the behaviour of the μ^--meson in substituting for an electron in the outer orbital parts of the atom, justify the inclusion of the μ-mesons as leptons. Beside the π^0-meson and the photon, there seems only one other departure from complete symmetry about the dividing line of Fig. 1; for while the K^0 meson and its anti-particle the $\bar{K}^0$ meson are distinct particles, they appear to decay instantaneously, each particle giving the same two products, K_1^0 and K_2^0 mesons.

In present-day terminology, the symbol 'V' is not for a kind of particle but for a class of event; that is, a 'V-event' is one that discloses a Λ-particle. And the lower-case Greek letters θ (or χ), τ and κ, formerly used as names for what were thought to be different particles, are now used to denote different modes of decay for the appropriate K meson. Thus, θ-decay yields a pair of π-mesons; τ-decay gives three π-mesons; and κ-decay a μ-meson, a π-meson, and a neutrino. Any of these fates may befall a single kind of K-meson.

All of this looks rather baffling, and perhaps the first way of bringing a little system into it is to consider what a particle and anti-particle pair really means. Such a pair, for example the electron–positron, is created together from radiation; and it is annihilated to give radiation. Both particles come from the vacuum, which has no charge or spin or angular momentum; and if charge and angular momentum are conserved n the act of creation, then the charges and angular momenta of the two particles must be equal and opposite. It follows that any particles possessing properties which are conserved during interactions can only be created from radiation in pairs; and that particle and anti-particle differ only in the sign or direction attached to these properties. If you wonder how an anti-neutron differs from a neutron, since neither is charged, the answer is in the sign (though not the size) of the ratio of the magnetic moment to the angular momentum. The odd men out of the scheme, the π^0-meson and the K_1^0 and K_2^0 mesons, are decay products only. As to the photon, I believe we could quite happily speak of photons and anti-photons if we wanted to, but any differences between the photons resulting from a pair annihilation are covered by ordinary optics; they can differ only in the vector aspects of direction of propagation and polarization.

The classification goes beyond a mere sorting out by mass, for the hyperons and K-mesons participate in what are called 'strong interactions', and the leptons in 'weak interactions'. The words 'strong' and 'weak' refer to the forces between particles as compared with the Coulomb electrostatic force to be expected between charged particles at the appropriate distance. The proton-neutron forces within the nucleus are 'strong', about 137 times greater than the corresponding Coulomb forces. An example of a 'weak' force is that binding the proton and electron together within a neutron; this is many orders of magnitude smaller than a Coulomb force would be; and the β-decay of the neutron, when it gives up an electron and becomes a proton, is a typical 'weak interaction'.

Further, the hyperons seem to bear the same sort of relation to nucleons as the π-meson and the μ-meson do to electrons. The hyperons are excited nucleons, which decay to a proton or a neutron; and during their brief lifetime they can take a nucleon's part in a nucleus, forming a partnership called a 'hyperfragment'. The μ-mesons decay to the electron and neutrino of the appropriate 'sign', as $\mu^- \rightarrow e^- + \nu^0$, and $\mu^+ \rightarrow e^+ + \bar{\nu}^0$. And, as already mentioned, the μ^--meson can replace an electron in the outer orbital electron shells—an interesting point here being that, since the mass is so much greater than m_e, the radius of the Bohr orbit is correspondingly less than that of the replaced electron.

If the hyperons are excited nucleons, and the μ-mesons excited electrons, where do the genuine mesons come in? We seem to be left only with the K-particles and the π-mesons; and these are the ones which are readily produced by nucleon-nucleon interactions, or are readily absorbed by nuclei.

The arrangement of Fig. 1 follows that given in an article by F. J. Dyson in *Scientific American*, September 1958. The inclusion of anti-particles for all the hyperons really derives from the same source. Glasstone says that, "for Λ^0, p^+, n^0, K^+, K^0, Ξ^-, Ξ^0, Σ^+, Σ^0, and Σ^-, anti-particles are expected to exist", and that "some are known, while others remain to be discovered". Confidence in the theory is nowadays great enough for people to be pretty certain that expectations will be fulfilled. And, just as the proofs for this article were being corrected, it was reported in the press that anti-Ξ particles had been definitely observed. There is one rather awkward question; if e^-, e^+ and p^+, p^-, and μ^-, μ^+ are particle-pairs, why are not Σ^-, Σ^+, or at least Σ^-, $\bar{\Sigma}^+$? The answer would simply appear to be the experimental fact that "some are, and some aren't", where charge sign-reversal is concerned.

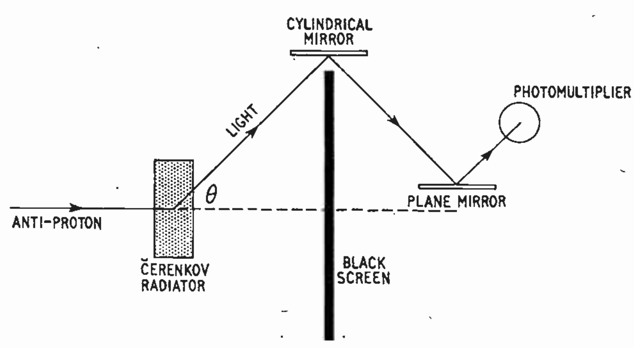
Anti-Proton p^- and Anti-Neutron $\bar{n}^0$

According to the early Dirac theory, a γ-ray photon entering the field in the neighbourhood of a nucleus can generate a pair of particles each of mass m from the surrounding vacuum if (a) the necessary energy $2mc^2$ is available from the photon, and (b) everything conservable is conserved; that is, the two taken together have a total charge, spin, momentum, etc. adding up to zero. It is not so much a matter of 'creation' as of supplying the necessary energy. In the vacuum, electrons, for example, are already there in plenty and occupying negative-energy states which are all filled. To produce an electron, it has to be extracted from its negative-energy world; and it leaves behind it a positive-hole which is its anti-particle, the positron. The energy need not necessarily come from a γ-ray, but this seems by far the most efficient process.

For an electron-pair, $2mc^2$ works out to about $2 \times 9 \times 10^{-28} \times (3 \times 10^{10})^2$ ergs, which is some 1·2 MeV, corresponding to a γ-ray photon of wavelength 10^{-10} cm. For a proton–anti-proton pair, as m is about $1836m_e$, the energy required is about 2,000 MeV. This is very much a lower limit; if particle bombardment is used to supply the energy, something like three times as much is needed; but it was realised that this should be within the range of the Berkeley proton-synchrotron. The idea of the experiment which resulted in the detection of the anti-proton seems to have been to turn on very much more than 2,000 MeV, and to see if proton-pair production could be detected in the resulting general upheaval.

In the first anti-proton experiments, a 6,200 MeV beam of protons from the Berkeley proton-synchrotron was directed on to a copper target. All sorts of things came out, including K-mesons and hyperons, positive, negative, and neutral. First, a mass-spectrometer-like magnetic field, using "$mv^2/r = Bev$" was used to separate out negatively-charged particles from the rest, and also to concentrate the particles with the same mv in the same place. The momentum of the suspected anti-proton was thus easily found. But, in order to identify it, m itself had to be determined, which involved measuring v. In the first experiments, this was done by straightforward timing of the particles over a path of about 40 feet. Later, the velocity was found using the Čerenkov radiation which came (at two stages removed really) from the particles when they were stopped in a thin glass plate. The Čerenkov detector shown in Fig. 2 is essentially a trap set to catch anti-protons travelling at a definite speed, or rather to detect only such particles. The photomultiplier can only receive light via the two reflectors if it strikes the first at (or close to) a certain angle. The anti-protons are annihilated with protons in the glass, giving rise to γ-ray photons, which eject high-energy electrons from the atoms they traverse. These electrons in their turn cause the glass to emit Čerenkov radiation as pulses of a bluish light, the angle θ in the diagram, between the direction of motion of the original particle and that of the light, depending on the energy of the original particle. This story sounds rather like the House that Jack Built and, before I had looked very closely into it, I used to think that the chain of operations was a good deal less complicated. But this is the way in which the device is described both by Glasstone, and also by J. V. Jelley in his book on "Čerenkov radiation". Further evidence for the new particle was found in the annihilation 'star' shown where the capture of an anti-proton in a photographic emulsion gave some dozen or so tracks of energetic particles whose total energy added up to the 2,000 MeV expected from the annihilation. The

Fig. 2. The Čerenkov velocity-selector of Chamberlain and Wiegand. Only light emitted at a certain angle θ. corresponding to a definite anti-proton velocity, can reach the photomultiplier via the cylindrical mirror and plane mirror; so the photomultiplier responds only to one anti-proton velocity



discovery of the anti-proton in 1955, reported by O. Chamberlain, E. Segre, C. E. Wiegand, and T. Ypsilantis, was a sign that the high-energy accelerating machines were by then powerful enough to be set to attack theoretical problems. And it was most encouraging to find that the theoretical predictions had been confirmed.

The anti-neutron, reported in September 1956 by P. Cork, Y. Lamberton, O. Piccioni, and W. Wenzel, came almost as an anti-climax. It had been so confidently awaited. It was produced by bombarding protons with anti-protons, and a strong source of the latter had to be developed before the job could be done. Fired at the protons in a liquid-hydrogen target, the majority of the anti-protons were simply annihilated. But some 0·3 per cent stopped short of this, and underwent only *charge* neutralization, as : *proton + anti-proton → neutron + anti-neutron*. The anti-neutrons were identified by the energy released on annihilation by collision with a neutron.

Since both the neutron and the anti-neutron are uncharged, what have they got to differ about, and how can we tell one from the other? The neutron has both angular-spin momentum and magnetic moment; so has the anti-neutron, but with the direction of the magnetic moment reversed with respect to that of the angular momentum. We are getting perilously close to another topic now, the business of Parity, and of Left-handedness and Right-handedness; this has been sorted out in a number of popular articles during the past two years, all quite excellently done, so perhaps we can leave it for the moment and come back to it in an unpopular way later. For, if we get too deeply involved on this point with the neutron, we shall certainly get bogged down for good with the neutrino.

The Neutrino, ν^0, and Anti-Neutrino, $\bar{\nu}^0$

Just to start off straight with this pair, we must first try to visualize how a particle with zero rest-mass and no charge can differ from its anti-particle at all; and the answer as for the neutron is in the sense of its spin. But there was at first some uncertainty as to whether there are two particles at all, and if so which was being observed. Certainly, the particle first detected by E. Reines and C. L. Cowan at Los Alamos was really an anti-neutrino. Before we confuse ourselves to the extent of calling a plain neutrino an anti-anti-neutrino, and possibly building up a long rally in this way, it might be well to look back in energy at the history and pedigree of this elusive little particle. It was first suggested by Fermi as a sort of face-saver. For there were a number of radioactive changes, of the type called β-emission or β-decay, in which the law of conservation of energy was apparently contravened; at the end, the *observed* particles always had less total energy than those at the start. Loyalty to the law of conservation of energy suggested that an *unobserved* (and, it might well have turned out, unobservable) particle was generated and stole away with the missing energy. The simplest example of such a change is the β-decay (so called because a β-particle, which is an electron, is emitted) of the neutron, as $n^0 \rightarrow p^+ + e^-$. Here, about one-third of the energy released is carried by the proton and electron. A neutrino, which was supposed to have very small or zero rest-mass, and no charge, but which travelled at a speed very close to that of light, was supposed to have made away with the other two-thirds. In fact, the above equation might have been written $n^0 \rightarrow p^+ + e^- + \nu^0$.

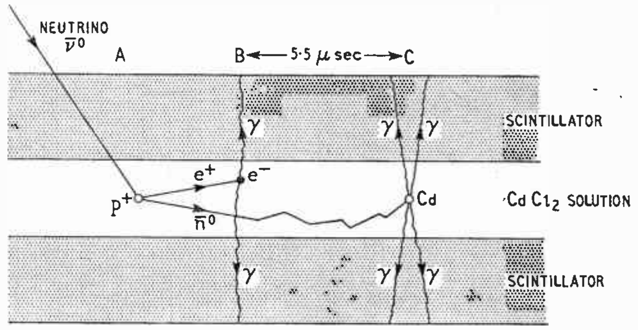


Fig. 3. *Scheme of the neutrino detector of Reines and Cowan. At A, the neutrino (or anti-neutrino) reacts with a proton:* $p^+ + \bar{\nu}^0 \rightarrow n^0 + e^+$. *At B, the positron is annihilated with an electron* e^- *giving* γ*-ray photons. The neutron travels for* 5.5 *microseconds, losing energy by collisions, until it is slowed up sufficiently to be captured by a cadmium nucleus, with emission of* γ*-ray photons. The recognition signal of the neutrino is thus two sets of coincidences in the scintillator, separated by* 5.5 *microseconds*

The law of conservation of momentum was the next principle to clamour for the neutrino. Nuclear changes can occur by the capture of electrons from the K-shell, as ${}_4Be^7 + e^- \rightarrow {}_3Li^7$. Cloud-chamber observations of the recoil of the ${}_3Li^7$ nucleus, from which its momentum-change could be calculated, showed that a neutrino must have been shot off in the opposite direction with that momentum, besides taking off some 50 eV or so of energy. The K-capture equation for beryllium-7 should, in fact, be written ${}_4Be^7 + e^- \rightarrow {}_3Li^7 + \nu^9$. Sometimes the neutrino appeared written in equations, sometimes not; but it was by no means hypothetical. It *had* to be there if the conservation laws were to hold universally. But it was a most unpromising character for detection. And a search for it could hardly have been organized before the establishment of large reactors which, generating large quantities of β-emitting isotopes, could be expected to give a large flux of neutrinos.

The effect looked for was a rare and improbable one, a reversed β-decay process, in which a neutrino is captured by a proton to give a neutron and a positron. In 1956, Reines and Cowan exposed a layer of cadmium chloride solution which was sandwiched between two large liquid scintillator vessels (Fig. 3), to the neutrino flux from a large reactor. The water of the solution provided target protons; it also slowed down the neutrons emitted by the reversed β-decay until they reached a speed at which they could be captured by cadmium nuclei. The expected reaction, then, was $p^+ + \nu^0 \rightarrow n^0 + e^+$. (Really, as it turned out, the ν^0 should have been $\bar{\nu}^0$.)

The simultaneous production of a neutron and a positron showed that this had happened. The effect sought, as evidence of this, was the emission of γ-ray photons of energy corresponding to the annihilation of the positron by an electron, and *at a calculated interval later* (which, with the arrangement of Fig. 3, was 5·5 microseconds) the emission of further γ-ray photons denoting the capture of the neutron by a cadmium nucleus.

Do not let Fig. 3 mislead you as to the size and complexity of the apparatus. Some idea of its scale and dimensions is given by Reines in a contribution to the Pergamon Press volume "Liquid Scintillation Counting". The liquid scintillators had a volume of 470 gallons; photomultiplier tubes were used by the hundred; and a giant computer-like sorting system was used for recording only those pairs of signals with the right 5·5–μsec delay between them. If the original neutrino source, a multi-megawatt pile, is included in the apparatus, Reines and Cowan must have achieved a record in the ratio of equipment size to objective size. The original scintillator outfit was called "El Monstro", and looked it; and there is a picture in the article of a vehicle like a rather large petrol lorry, which one would like to think was used for 'topping up'.

Since the laws of conservation of momentum and of conservation of energy have been completely vindicated, and since the neutrino had been invoked to satisfy them, one might be tempted to think that the detection of the neutrino had nicely closed a chapter. Not so. There is the question whether the neutrinos of β-decay and of K-capture are the same, and whether, if they are different, this would be detectable or would matter at all. In the first place, the two particles are different. The neutrino from β-decay and, of course, from the Reines-Cowan reactor source, is really the anti-neutrino with (viewed from behind as it travels away from you, so to speak) an anti-clockwise spin. That from K-electron capture is a plain neutrino, with clockwise spin. Many of the nuclear reactions occurring in the sun, involving positron emission (which is β-decay with the 'anti' sign) do indeed generate neutrinos; and since these are almost unstoppable, we must be receiving a strong neutrino flux from the sun.

Next, as to whether the difference between ν^0 and $\bar{\nu}^0$ matters very much anyhow. The theoretical side of the parity question, which shows that it does matter, must be left for now. But there is a practical case also, that of a radioactive nucleus undergoing *double* β-decay, and thus shooting off two neutrinos in succession. It has been shown that, if the two neutrinos are different, then the parent element should have a much longer half-life than if they are the same; and, so far as measurements have been made, they confirm the longer period, and so show that the two neutrinos are unlike.

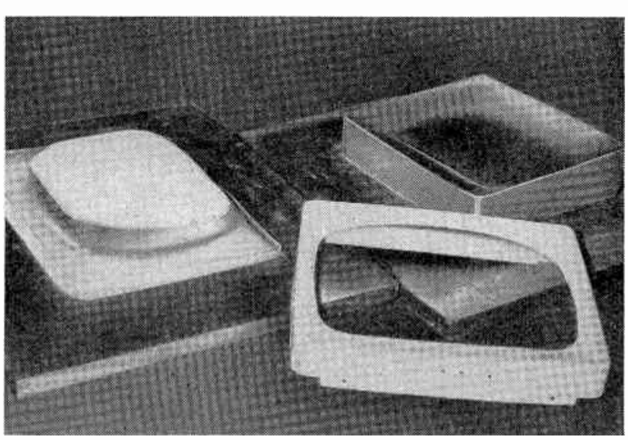
Of course, there may be another line of approach. You hear all this jargon about particles 'seeing' fields and motions and other particles, just as if they were almost human. This anthropomorphic tendency is certainly implied in the popular term for spinning electrons, or indeed any particles possessing spin; they are (I understand) referred to, not as little spinning tops, but as 'Top People'. It would be going too far to assume from this phrase that neutrinos are urban, squat, and full of guile; but it may well be that they have the all too human propensity (exacerbated whenever electromagnetism creeps in) of being unable to discern between their right hand and their left hand. Why should physics accept them at their own valuation when they themselves may simply be looking at us the wrong way round?

But there is a clear objective means of distinguishing between the two. In their action on nucleons we have for the anti-neutrino from a reactor, $\bar{\nu}^0 + p^+ \rightarrow n^0 + e^+$, and for a neutrino from the sun, $\nu^0 + n^0 \rightarrow p^+ + e^-$. So, in their actions on the nuclei of atoms, the anti-neutrino, converting a proton to a neutron, should cause a decrease in atomic number; while the neutrino, converting a neutron to a proton, should cause the atomic number to increase, as ${}_{17}Cl^{37} + \nu^0 \rightarrow {}_{18}A^{37} + e^-$. It has been suggested that this conversion from chlorine-37 to argon-37 in sea water by the action of solar neutrinos might be sought for and detected, for the argon-37 is radioactive. This would be a large-scale experiment to undertake. It has, however, been established that the reactor-produced anti-neutrino does *not* transmute chlorine-37 to argon-37.

Glasstone goes no further into neutrino physics, which already seems to have developed into a vast discipline of its own; and he does not mention the more recent 'two-component theory' of the neutrino, which seems to be a very abstruse business anyhow. One experimental point should be mentioned in conclusion. The statement above, about the senses of the spins of neutrino and anti-neutrino, is not generally true. This depends on the kind of action that they arise from. The neutrino emitted in π^+-meson decay *has* a clockwise spin; but the decay of the μ^+-meson yields an anti-neutrino with a clockwise spin, and a neutrino with an anti-clockwise spin. I have been trying to look into this, in some articles by C. S. Wu and L. Lederman, and also by a team from the Argonne laboratory, which describe some 'time-reversal' experiments; it is reasonably easy to follow the experiments, at least. How they tie up with the theory is indeed rather harder.

PLASTIC TOOLING

Plastic press tools, made from ready-mixed Double Bond Toolform compounds, for forming a television mask in aluminium. Produced by the Kenilworth Manufacturing Co. Ltd., West Drayton, Middlesex, these componnds can be used for making jigs, moulds, fixtures, etc., with considerable reduction in tooling time



Response of Cascaded Double-Tuned Circuits

By Yona Peless, M.S.E.E.*

SUMMARY. *The transient and steady-state responses of cascaded identical double-tuned circuits are developed in terms of the locations of the poles of the transfer function. Results are obtained for two arbitrarily placed complex conjugate pole pairs so that the work applies to networks with an amplitude response which is not necessarily symmetrical about the band centre; however, the narrow-band restriction is imposed. Detailed results are given for the bandwidth variation with the number of stages and circuit parameters and these are compared with the impulse response obtained for corresponding conditions. It is found that the peak of the impulse response is very nearly linearly related to the 6-dB bandwidth, provided that no appreciable asymmetry in the amplitude response exists.*

The work reported here was undertaken as part of a study of the response of radio interference and field intensity meters to a number of commonly encountered input waveforms. The results are general, however, and may be applied wherever the response characteristics of double-tuned circuits are required.

The radio interference and field intensity meter may be viewed as a specially-designed narrow-band receiver, with the band limiting usually provided by the i.f. amplifier, followed by suitable detectors and a d.c. indicator. Impulse measurements are made by reading the peak and/or average amplitude of the envelope of the i.f. amplifier output, and from these the impulse strength at the input is inferred. However, this inference must take into account the bandpass characteristics since the peak reading is affected when the bandwidth is less than the inverse of the pulse duration, and the average reading is affected in cases of large 'undershoot' at the trailing edge of the impulse-response envelope. For those cases where the i.f. amplifier comprises identical single-tuned circuits, the results of previous studies[1,2,3,4] appear adequate.[†] Here we shall further develop the impulse response of cascaded identical double-tuned circuits, with the following objectives in mind:

(a) To determine the relationship between the system bandwidth and the peak value of the impulse response and, in particular, to find the region in which their ratio is substantially constant.

(b) To determine the extent of undershoot (or, more appropriately, the amplitude of the second maximum) of the impulse-response envelope and, in particular, to note the conditions under which the undershoot is small.

Problem Formulation

The circuit under discussion throughout this article is a cascade of identical interstage networks of the type shown in Fig. 1, separated by thermionic valves.

* *Work done at the Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia., Pa. Now with the Israel Institute of Technology, Haifa, Israel.*

† Earlier work in the area of transformer coupling was done by C. C. Eaglesfield[5].

The transfer function of a double-tuned circuit may be written as:

$$Z(p) = \frac{e_0}{i_p} = \frac{Kp}{(p-p_1)\,(p-p_2)\,(p-p_1^*)\,(p-p_2^*)} \quad (1)$$

where p_j^* denotes the conjugate of p_j.

The gain function (which is the ratio of the voltages at two successive grids) is this expression multiplied by the mutual conductance of the valve. Here we have assumed that the driving-point impedance of the interstage network is small as compared to the anode resistance of the valve. It is seen from the above equation that the

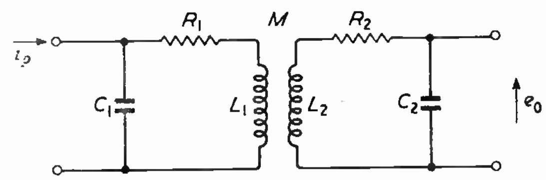


Fig. 1. *Series-loaded mutual-inductance coupled circuit*

transfer function has a fourth-order polynomial in the denominator. This transfer function has a zero of first order at the origin, and two pairs of poles in the left-half of the p-plane. The location of the pole pairs is determined by the circuit parameters and may be found by the method outlined in Appendix 1.

In the case of the bandpass amplifiers with small fractional bandwidth, the poles of the transfer function appear in two conjugate pairs with the real parts α_1 and α_2 small as compared with the imaginary parts β_1 and β_2. $p_0 = \alpha_0 + j\omega_0$ is the arithmetic mean of p_1 and p_2 and the angle ϕ is measured as indicated in Fig. 2. The case usually dealt with in the literature is the one in which $\phi = 0$ (i.e., p_1 and p_2 have the same real part), when the amplitude response is symmetrical about the centre frequency in the narrow-band case. To study the effect of asymmetry and to enable the use of the results for cascades of narrow-band single-tuned pairs we let ϕ take values different from zero.

The impulse response of n cascaded stages is solved in Appendix 2. The resulting envelope is given by

$$E(\tau) = \frac{\sqrt{2\pi}\, A^n}{2^{n-1}\,(n-1)!\,\rho^{2n-1}} \exp\left(\frac{\alpha_0}{\rho}\tau\right) \tau^{n-\frac{1}{2}} \left| J_{n-\frac{1}{2}}\left(\tau\, e^{j\phi}\right) \right|$$

where $A = \dfrac{K}{4\omega_0}$, $\tau = \rho t$ (normalized time), and ρ is defined in Fig. 15.

The Symmetrical Amplitude Response Case $\phi = 0°$

Location of poles.

The symmetrical response in narrow-band double-tuned circuits is associated with a pole configuration in which the line joining the poles is parallel to the $j\omega$ axis

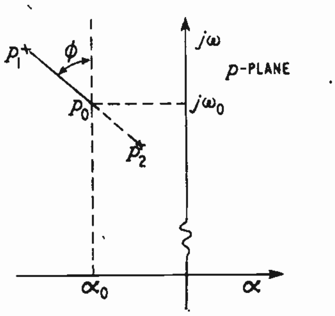


Fig. 2. *Detail of the pair of poles in the upper half of the p-plane*

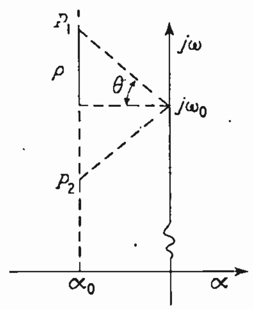


Fig. 3. *Pole configuration for the symmetrical amplitude response*

(Fig. 3). All circuits having the same $q = -\alpha_0/\rho$ ratio will have similar amplitude, phase, and transient responses. It is therefore convenient to classify the different cases according to $q = -\alpha_0/\rho = \cot\theta$ (α_0 being always negative). The case when $\theta = 45°$ and $q = 1$ is the one for which the amplitude response of one stage has a flat top (maximally flat or Butterworth response). If $\theta > 45°$ two peaks are noted. For $\theta = 30°$ ($q = 1{\cdot}732$) the phase response is approximately linear (the envelope delay is maximally flat).

The angular-frequency variable is normalized to $\Omega/\sqrt{\alpha_0^2 + \rho^2}$ where $\Omega = \omega - \omega_0$, ω_0 being the centre angular frequency.

Amplitude Characteristics

Although amplitude characteristics were computed in previous publications on the subject of double-tuned circuits they were always given as a function of circuit parameters such as Q and coefficient of coupling. For the sake of completeness similar curves are included here, but as a function of the location of the poles of the transfer function. (This way they may also be used for the analysis of cascaded single-tuned pairs.)

Fig. 4 shows the amplitude response of double-tuned circuits. The response is given in decibels per stage. To get the response of n cascaded stages the attenuation scale should be multiplied by n.

From the amplitude response the normalized 3-dB and 6-dB bandwidths $\Delta f_{3dB}/\sqrt{\alpha_0^2 + \rho^2}$ and $\Delta f_{6dB}/\sqrt{\alpha_0^2 + \rho^2}$ were computed for cascades of 1, 2, 3, 4, 5, 6, and 10 stages and are shown in Figs. 5 and 6.

Impulse Response

The impulse response was computed for the cases $q = 0{\cdot}577$, $0{\cdot}649$, $0{\cdot}726$, $0{\cdot}810$, $0{\cdot}900$, $1{\cdot}000$, $1{\cdot}111$, $1{\cdot}235$, $1{\cdot}376$, $1{\cdot}540$, and $1{\cdot}732$ (corresponding to q between 60° and 30° in 3° steps, respectively). Two to six stages in cascade were considered. Figs. 7 and 8 show the envelopes of these responses for three and five stages respectively, q being the variable parameter. The amplitude was normalized so that the peak of the response is equal to the normalized impulse bandwidth defined below. From the computed data the normalized impulse bandwidth $\Delta f_{imp}/\sqrt{\alpha_0^2 + \rho^2}$ was computed and is shown in Fig. 9. The definition used for the impulse bandwidth is

$$\Delta f_{imp} = \frac{E_{max}}{2G(f_0)}$$

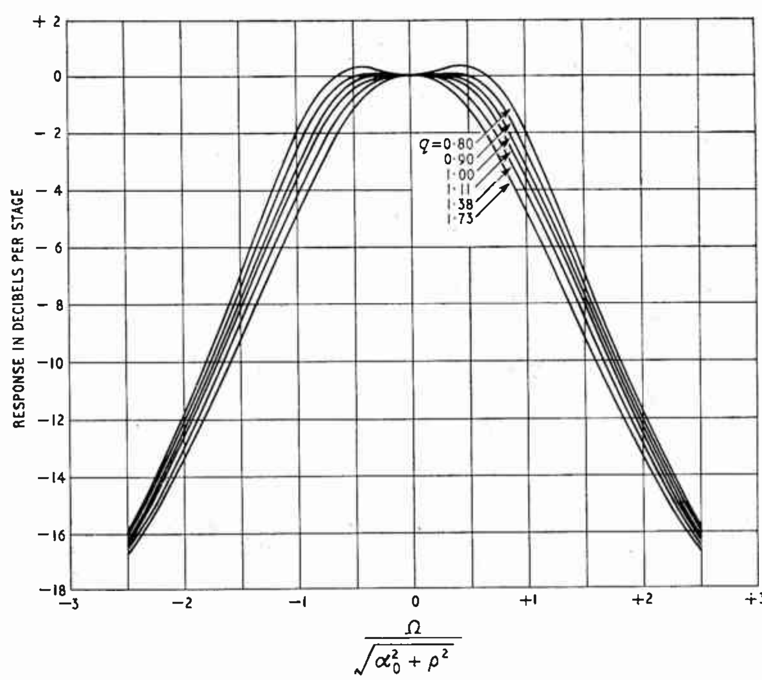
where E_{max} is the peak voltage at the output of a cascade in response to a unit impulse, and $G(f_0)$ is the centre-frequency gain of the cascade.

Comparison of the Impulse Bandwidth and the 6-dB Bandwidth

The ratio $\Delta f_{imp}/\Delta f_{6dB}$ was computed and is given in Table 1 as a function of q and n.

From this data it is possible to find the impulse bandwidth from steady-state measurements or the 6-dB

Fig. 4. *Amplitude response*



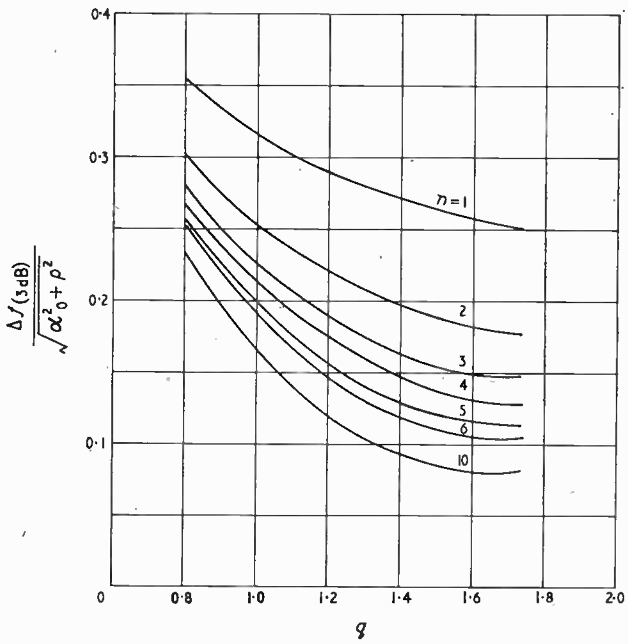


Fig. 5. *Variation of 3-dB bandwidth*

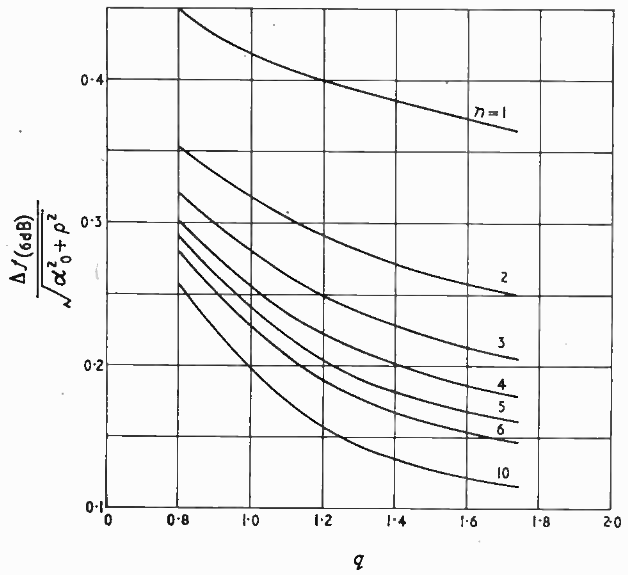


Fig. 6. *Variation of 6-dB bandwidth*

bandwidth from impulse-response measurements.

Overshoots

As a measure of the amount of overshoot γ, we use the ratio of the second peak of the impulse-response envelope to the first one. This ratio is a function of the number of stages of the cascade and the parameter $q = \alpha_0/\rho$. For a given number of stages it is a function of q only. This enables one to find q by measuring the overshoot of a cascade, if it can be assumed that all stages are identical and have a symmetrical response about the centre frequency of the amplifier. With the knowledge of q and n the steady-state and transient-response curve shapes are defined. Fig. 10 shows the variation of γ with q for $n = 2, 3, 4, 5$ and 6 stages.

The Non-Symmetrical Response, $\phi \neq 0°$

Location of Poles and Steady-State Response

The pole configuration for the non-symmetrical response is shown in Fig. 11. Fig. 12 shows the amplitude

TABLE 1

The Ratio of Impulse Bandwidth to the 6-dB Bandwidth

q / n	0·810	0·900	1·000	1·111	1·376	1·732
1			1·09			
2	1·08	1·06	1·05	1·04	1·05	1·06
3	1·07	1·04	1·03	1·03	1·04	1·05
4	1·06	1·04	1·02	1·02	1·03	1·05
5	1·05	1·03	1·02	1·02	1·03	1·05
6	1·05	1·02	1·01	1·01	1·03	1·05

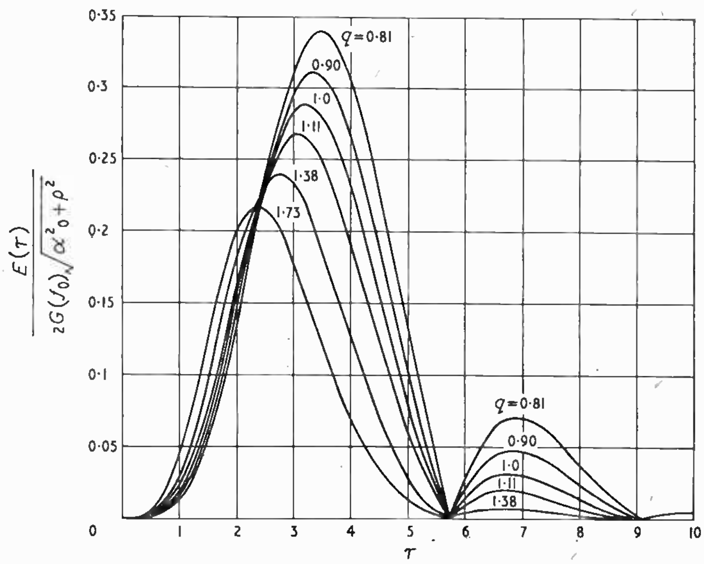
TABLE 2

The Normalized 6-dB Bandwidth $\Delta f_{6\,dB}/\sqrt{\alpha_0^2+\rho^2}$ for the case $\phi = 5°$

n / k	1	2	3	4	5	6
0·60	0·474	0·389	0·366	0·352	0·340	0·335
1·00	0·413	0·312	0·276	0·253	0·236	0·221
1·20	0·393	0·289	0·247	0·219	0·202	0·188

response resulting from such a configuration of poles for the case $\phi = 5°$, and different values of $q = -\alpha_0/\rho$. The curves are given in decibels per stage as a function of the normalized deviation frequency Ω/ρ. Here, as in the symmetrical case, we distinguish two peaks in the response in cases for which $q < 1$ and a single peak for $q > 1$.

Fig. 7. *Impulse response of three stages*



The normalized 6-dB bandwidth was computed for cascades of $n = 1$ to $n = 6$ stages and are given in Table 2.

The Impulse Responses

The expression for the envelope of the impulse response has been derived, see reference 2 and Equ. (16). From the cases computed, two representative figures were drawn. Fig. 13 shows the effect of changing the angle ϕ of the pole location in a single stage. It is seen that

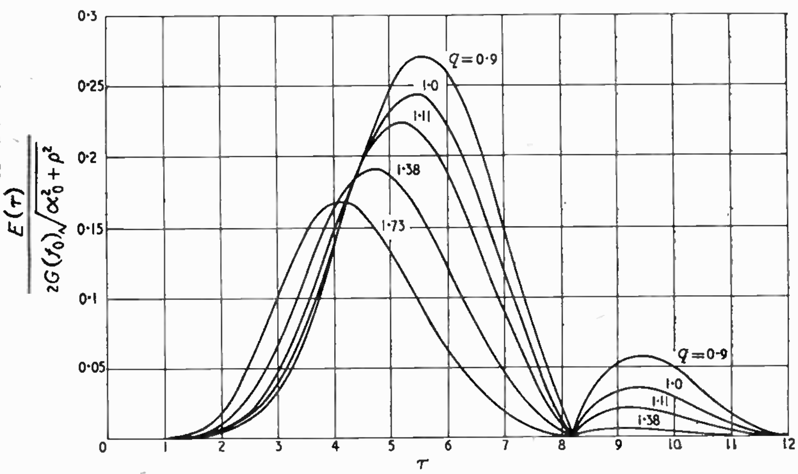


Fig. 8. *Impulse response of five stages*

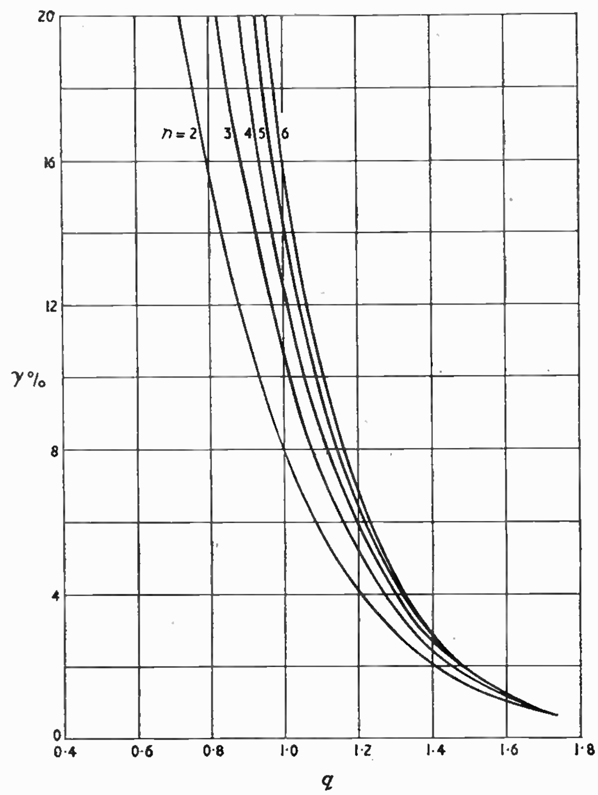


Fig. 10. *Percentage of overshoot*

as ϕ becomes larger the transient dies out smoother but slower. A response similar to the one denoted by $\phi = 15°$ may be desirable when a smooth transient response is important. The particular value of $q = 0{\cdot}6$ was chosen in this figure to provide a relatively large amount of overshoot, thereby allowing the effect of varying ϕ to be seen more clearly. The same effect, on a smaller scale, results with larger values of q.

Fig. 14 shows the impulse response of a cascade of five stages with $\phi = 2°$ and q as the parameter. The amplitude is normalized to give $(\sqrt{\pi/2})\ \Delta f_{imp}/\sqrt{\alpha_0^2+\rho^2}$ at the peak of each curve.

Table 3 gives the normalized impulse bandwidth for a cascade of $n = 1$ to $n = 6$ stages and $\phi = 5°$.

Fig. 9. *Variation of impulse bandwidth*

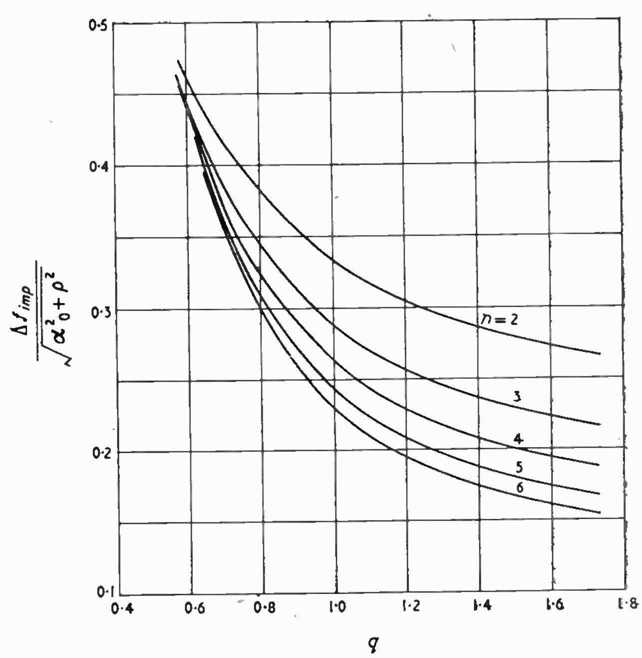


TABLE 3

The Normalized Impulse Bandwidth $\Delta f_{imp}/\sqrt{\alpha_0^2+\rho^2}$ *for the case* $\phi = 5°$

n / k	1	2	3	4	5	6
0·60	0·544	0·468	0·456	0·460	0·470	0·486
1·00	0·456	0·336	0·294	0·271	0·253	0·240
1·20	0·432	0·306	0·261	0·230	0·214	0·202

TABLE 4

The Ratio $\Delta f_{imp}/\Delta_{6\,\mathrm{dB}}$ *for the case* $\phi = 5°$

n / k	1	2	3	4	5	6
0·60	1·15	1·20	1·24	1·31	1·38	1·45
1·00	1·10	1·08	1·06	1·07	1·07	1·08
1·20	1·10	1·06	1·05	1·05	1·06	1·07

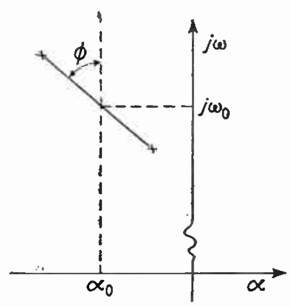
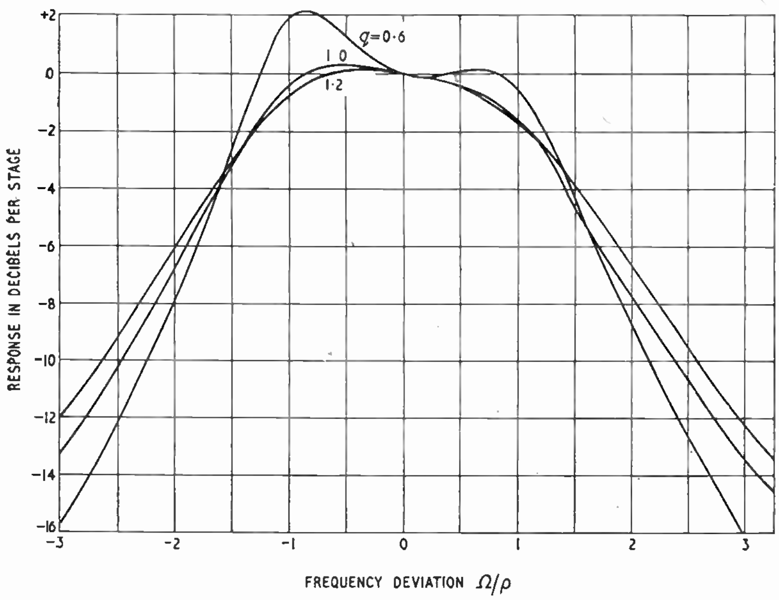


Fig. 11. *Pole configuration for non-symmetrical response*

Fig. 12. *Amplitude response;* $\phi = 5°$



From the tabulated values in Tables 2 and 3 the ratio $\Delta f_{imp}/\Delta f_{6\,\mathrm{dB}}$ was computed and is given in Table 4.

Comparing Table 4 with Table 1 it is seen that the ratio $\Delta f_{imp}/\Delta f_{6\,\mathrm{dB}}$ becomes larger when the asymmetry is greater. However, if we disregard the cases such as $q = 0{\cdot}60$ and $\phi = 5°$, the amplitude response of which is shown in Fig. 12 to be far from what we find in available i.f. amplifiers, then the ratio of impulse bandwidth to 6-dB bandwidth varies from 1·01 to 1·08 in most practical cases. A more accurate value for this ratio may be obtained from the given tables for each particular case.

Summary

The data presented here enables one to analyse a cascade of double-tuned stages or single-tuned pairs with narrow relative bandwidth. From the location of the poles of the transfer function of such a cascade the steady state and the envelope of the impulse response may be easily found. In the symmetrical-passband case, curves for the 3-dB, 6-dB and impulse bandwidth are shown. Also the overshoot of the transient response is given. For both the symmetrical and some non-symmetrical cases the ratio of the impulse bandwidth to the 6-dB bandwidth was computed and given in the form of tables. It is shown that this ratio is between 1·01 and 1·08 in a large range of practical cases. For most cases likely to be experienced in amplifier design (3 to 6 transformers, transitional coupling or somewhat undercoupled) the range 1·01 to 1·03 can be expected. However, for large asymmetries in the amplitude characteristic, the ratio of impulse to 6-dB bandwidth can be expected to differ from values in the range given above by substantial amounts.

Acknowledgement

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APPENDIX 1

Location of Poles

In the process of computing the impulse response of a double-tuned circuit or single-tuned pair it is necessary to find the pole location from a known transfer function. This amounts to finding the roots of a fourth-order polynomial. Although methods of solution are well known they are sometimes long and tedious. In our case, however, a short method may be used since we know that the real parts of the roots are small compared with their imaginary parts.

If we call r_j the distance of a pole p_j from the origin and α_j its real part, then the fourth-order polynomial may be written in terms of its roots as shown by Equ. (1).

$$\left.\begin{aligned} & p^4 + b_3p^3 + b_2p^2 + b_1p + b_0 \\ & = (p-p_1)\,(p-p_2)\,(p-p_1^*)\,(p-p_2^*) \\ & = p^4 - 2\,(\alpha_1 + \alpha_2)\,p^3 + (r_1^2 + r_2^2 + 4\alpha_1\alpha_2)p^2 \\ & \quad - 2\,(\alpha_1 r_2^2 + \alpha_2 r_1^2)p + r_1^2 r_2^2 = 0 \end{aligned}\right\} \quad \ldots \quad (1)$$

For circuits with Q higher than 10

$$r_1^2 + r_2^2 \gg 4\alpha_1\alpha_2$$

The coefficient b_2 of p^2 may therefore be approximated by

$$b_2 \approx r_1^2 + r_2^2$$

But we also have

$$b_0 = r_1^2 r_2^2$$

Solving these two equations we find r_1^2 and r_2^2

$$r_1^2 = \frac{b_2 + (b_2^2 - 4b_0)^{1/2}}{2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

and $$r_2^2 = \frac{b_2 - (b_2^2 - 4b_0)^{1/2}}{2} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

We then have the equations for the other two coefficients

$$b_3 = -2(\alpha_1 + \alpha_2)$$

and $$b_1 = -2(\alpha_1 r_2^2 + \alpha_2 r_1^2)$$

the solutions of which are given by

$$\alpha_1 = \frac{b_1 - b_3 r_1^2}{2(r_1^2 - r_2^2)} \quad \ldots \quad (4)$$

and $$\alpha_2 = \frac{b_3 r_2^2 - b_1}{2(r_1^2 - r_2^2)} \quad \ldots \quad (5)$$

After finding r_j and the real part α_j the imaginary part of the pole is given by

$$\beta_j = (r_j^2 - \alpha_j^2)^{1/2} \quad \ldots \quad (6)$$

The procedure gives accurate enough results for high-Q circuits. For moderate and low Q it may be repeated correcting the value of $r_1^2 + r_2^2$ by the use of the results obtained for α_1 and α_2.

APPENDIX 2

The Impulse Response

The impulse response of $Z(p)$ is given by

$$f(t) = \frac{1}{2\pi j} \cdot \int_{c-j\infty}^{c+j\infty} Z(p)\, e^{pt} dp \quad \ldots \quad (7)$$

since all the poles of our transfer function are in the left-half plane we may write ($c = 0$; $p = j\omega$)

$$f(t) = \frac{1}{2\pi} \cdot \int_{\omega=-\infty}^{\infty} Z(j\omega) \cdot e^{j\omega t}\, d\omega = \frac{2}{2\pi} \cdot \int_0^{\infty} \mathrm{Re}\,[Z(j\omega)\, e^{j\omega t}]\, d\omega \quad (8)$$

The transfer function of n identical inductively-coupled double-tuned stages is given by

$$Z(p) = \frac{K^n p^n}{(p-p_1)^n (p-p_2)^n (p-p_1^*)^n (p-p_2^*)^n}$$

or $$Z(j\omega) = \frac{K^n (j\omega)^n}{(j\omega-p_1)^n (j\omega-p_2)^n (j\omega-p_1^*)^n (j\omega-p_2^*)^n} \quad \ldots \quad (9)$$

The exact evaluation of the impulse response is very complicated even for low values of n. Looking at Equs. (7) and (8) we notice that the integrand is of considerable amplitude only in the neighbourhood of the poles (in the pass band) and tends to zero rapidly even for moderate Q if the number of stages is two or more. We may therefore introduce the approximation

$$\frac{(j\omega)^n}{(j\omega-p_1^*)^n (j\omega-p_2^*)^n} \approx \frac{1}{(4j\omega_0)^n}$$

The approximate form of the transfer function is now

$$Z(j\omega) \approx \frac{(A/j)^n}{(j\omega-p_1)^n (j\omega-p_2)^n} \quad \ldots \quad (10)$$

where $A = \dfrac{K}{4\omega_0}$.

We now substitute this transfer function into the expression for the inverse transform. The amplitude of the integrand for negative frequencies is very small and dies out as fast as $1/\omega^{2n}$ as ω approaches $-\infty$. We may therefore integrate along the whole $j\omega$-axis and, as a result, the impulse response is given by

$$f(t) \approx 2\,\mathrm{Re.}\left\{\frac{1}{2\pi} \cdot \int_0^{\infty} \frac{(A/j)^n}{(j\omega-p_1)^n (j\omega-p_2)^n} d\omega\right\}$$

$$\approx 2\,\mathrm{Re.}\left\{\int_{-\infty}^{\infty} \frac{(A/j)^n e^{j\omega t}}{(j\omega-p_1)^n (j\omega-p_2)^n} d\omega\right\}$$

$$= 2\,\mathrm{Re.}\left[\frac{1}{2\pi j} \oint \frac{(A/j)^n e^{pt}}{(p-p_1)^n (p-p_2)^n} dp\right] \quad \ldots \quad (11)$$

where Re. stands for "the real part of".

To solve Equ. (11) we shift the origin of the p-plane to the point p_0 using the following transformation and notations:

$$s = p - p_0 \qquad p_0 = (p_1 + p_2)/2$$

$$s_1 = p_1 - p_0$$

$$s_2 = p_2 - p_0 = -s_1$$

$$s_1 = \rho e^{j(\phi + \pi/2)}$$

Fig. 15 shows the location of the poles in the s-plane.

Using the above transformation Equ. (11) takes the form

$$f(t) = 2\,\mathrm{Re.}\left[\left(\frac{A}{j}\right)^n \exp(\rho t)\, \frac{1}{2\pi j} \oint \frac{e^{st}\, ds}{(s-s_1)^n (s+s_1)^n}\right] \quad \ldots \quad (12)$$

The inverse transform may now be found in tables; e.g., reference (6).

$$\frac{1}{2\pi j} \oint \frac{e^{st}\, ds}{(s-s_1)^n (s+s_1)^n} = \frac{\sqrt{\pi}}{(n-1)!} \left(\frac{t}{2s_1}\right)^{n-1/2} \cdot \mathrm{I}_{n-1/2}(s_1 t)$$

$$= j^{-n+1/2} \frac{\sqrt{\pi}}{(n-1)!} \left(\frac{t}{2s_1}\right)^{n-1/2} \cdot \mathrm{J}_{n-1/2}(j s_1 t) \quad \ldots \quad (13)$$

since $\mathrm{I}_k(x) = j^{-k}\, \mathrm{J}_k(jx)$.

$\mathrm{J}_k(x)$ and $\mathrm{I}_k(x)$ are the Bessel function and the modified Bessel function of first kind and kth-order respectively.

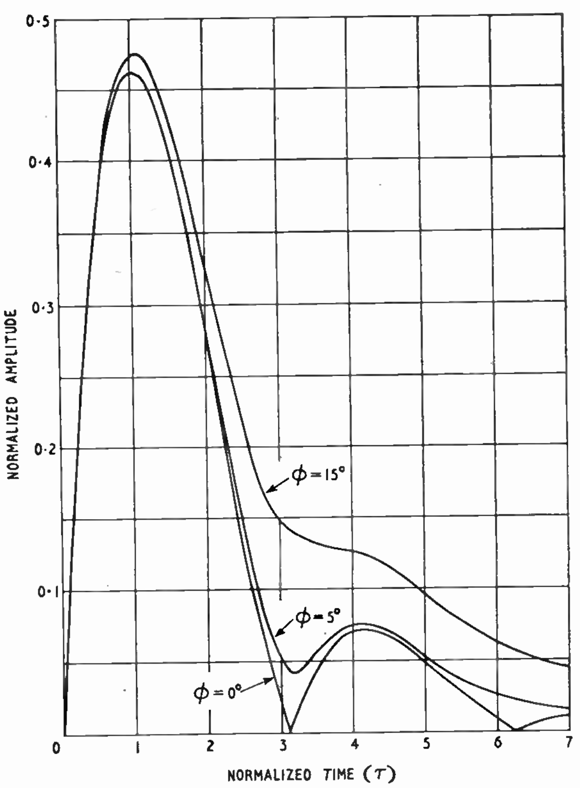
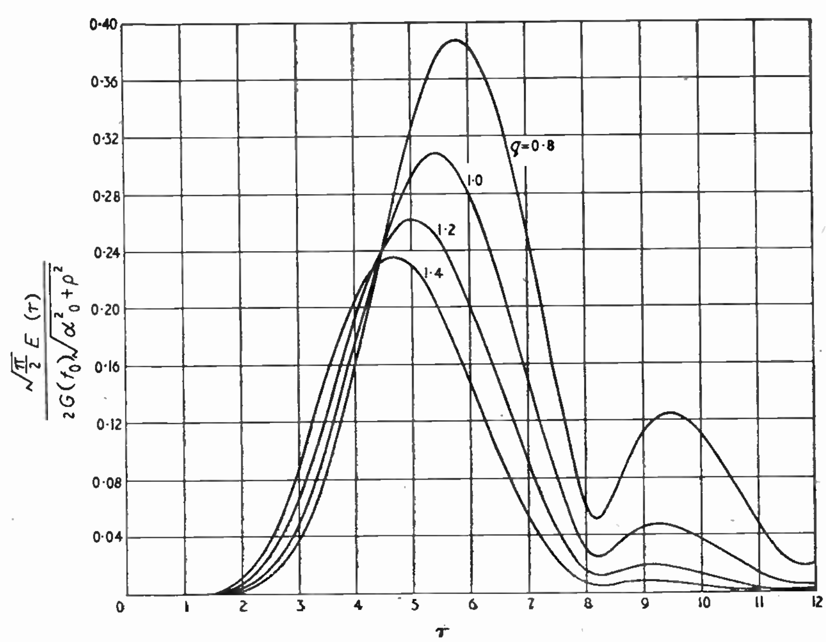


Fig. 13. *Impulse response of a single stage;* $-\alpha_0/\rho = q = 0{\cdot}6$

Fig. 14. *Impulse response of five stages;* $\phi = 2°$



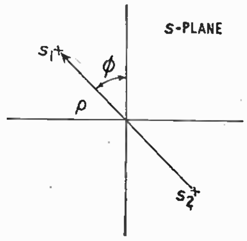


Fig. 15. *Location of poles in the s-plane*

The impulse response may now be written as

$$f(t) = \frac{2\sqrt{\pi}}{(n-1)!} A^n \exp(\alpha_0 t)$$

$$\text{Re.}\left\{\exp(j\omega_0 t)\left[j^{-2n+1/2}\left(\frac{t}{2s_1}\right)^{n-1/2} \mathrm{J}_{n-1/2}(js_1 t)\right]\right\} \quad (14)$$

where p_0 was replaced by its real and imaginary parts $p_0 = \alpha_0 + j\omega_0$.

To find the envelope E of the impulse response, we make use of the following equation

$$\text{Envelope of } [\text{Re. }\{[\exp(j\omega_0 t)]\ (\alpha_0 + j\beta_0)\}] = \text{Envelope of } (\alpha \cos \omega_0 t - \beta \sin \omega_0 t) = (\alpha^2 + \beta^2)^{1/2} = |\alpha + j\beta| \quad .. \quad (15)$$

Using Equ. (15) and normalizing the time scale to $\tau = \rho \cdot t$ (ρ is defined in Fig. 15) the envelope of the impulse response of n cascaded identical double-tuned stages is given by

$$E(\tau) = \frac{\sqrt{2\pi}\, A^n \exp(\alpha_0 \tau/\rho)}{2^{n-1}(n-1)!\,\rho^{2n-1}} \tau^{n-1/2} \cdot |\mathrm{J}_{n-\frac{1}{2}}(\tau e^{j\phi})| \quad .. \quad (16)$$

The impulse-response envelope given by Equ. (16) contains the Bessel function of half-integer order of the complex argument ($\tau e^{j\phi}$). These have been tabulated[7] only for the special case $\phi = 0°$. For $\phi \neq 0°$, no tables are available.

MATHEMATICAL TOOLS

By Computer

The Binomial Theorem

Last month we considered the evaluation of a polynomial $f_1(x)$ in the independent variable x by methods other than direct substitution, and we found that if several values of x in the neighbourhood of say 2 were required, it was advantageous to express $f_1(x)$ in the equivalent form $f_2(y)$, where y stands for $(x-2)$. We obtained $f_2(y)$ from $f_1(x)$ by a process involving repeated division by $(x - 2)$, but it was noted that an alternative approach was possible if we turned round the relation

$$y = x - 2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

and put it in the equivalent form

$$x = y + 2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

If this is done, it is formally possible to write down $f_2(y)$ immediately. As in last month's article, we shall illustrate the procedure in terms of the example

$$f_1(x) \equiv x^6 - 3x^5 + 9x^4 - 27x^3 + 40x^2 - 60x + 100 \quad (3)$$

Substituting from Equ. (2) into Identity (3), we have

$$f_1(x) \equiv f_2(y) \equiv (y+2)^6 - 3(y+2)^5 + 9(y+2)^4 - 27(y+2)^3 + 40(y+2)^2 - 60(y+2) + 100 \quad .. \quad .. \quad (4)$$

but Identity (4) is of little use unless we can write down an explicit formula for $(y + 2)^n$. From the repeated division process used last month we know that Identity (4) ought to be the same as

$$f_1(x) \equiv f_2(y) \equiv y^6 + 9y^5 + 39y^4 + 85y^3 + 94y^2 + 16y + 36 \quad (5)$$

which was Identity (8) of last month. We now seek to show that Identity (4) is not merely formal, but genuinely gives us an alternative way of deriving Identity (5); it is a most useful check on numerical and algebraic working if we can derive an expression by two different methods which involve different steps for, if we arrive at the same answer by both methods, we can be almost certain that that answer is correct. If we obtain different answers, there must be an error somewhere and, when an error is known to exist in a calculation of limited extent, it can usually be traced quite easily. If necessary the two calculations can be carried out independently. A different person is very unlikely to make exactly the same mistake as the original calculator.

Our main task now, therefore, is to find a formula for $(a + x)^n$ where in the first instance n is a positive integer and a and x are unrestricted. Later, we shall show how this formula can be extended to values of n other than positive integers, but this extension involves important restrictions on the permissible values of a and x.

First, we can derive $(a + x)^2$ and $(a + x)^3$ and $(a + x)^4$ by direct multiplication; the working, though straightforward, is included because it makes clear the way in which the coefficients of the various terms of $(a + x)^n$ are built up.

$$\begin{array}{rr} & a + x \\ & a + x \\ \hline & a^2 + ax \\ & ax + x^2 \\ \hline (a+x)^2 = & a^2 + 2ax + x^2 \\ \hline \end{array}$$

$$
\begin{array}{rl}
(a+x)^2 = & a^2+2ax+x^2 \\
 & a+x \\
\hline
 & a^3+2a^2x+ax^2 \\
 & \qquad a^2x+2ax^2+x^3 \\
\hline
(a+x)^3 = & a^3+3a^2x+3ax^2+x^3 \\
\hline
\end{array}
$$

$$
\begin{array}{rl}
(a+x)^3 = & a^3+3a^2x+3ax^2+x^3 \\
 & a+x \\
\hline
 & a^4+3a^3x+3a^2x^2+ax^3 \\
 & \qquad a^3x+3a^2x^2+3ax^3+x^4 \\
\hline
(a+x)^4 = & a^4+4a^3x+6a^2x^2+4ax^3+x^4 \\
\end{array}
$$

Now let us arrange the various coefficients that have so far arisen in an orderly manner (known as Pascal's Triangle) and see how we can extend the triangle without actually having to do the work of multiplication, rather as a cricketer who hits a boundary scores his runs without actually having to run them.

TOP CORNER					1				
FIRST ROW $(a+x)$				1		1			
SECOND ROW $(a+x)^2$			1		2		1		
THIRD ROW $(a+x)^3$		1		3		3		1	
FOURTH ROW $(a+x)^4$	1		4		6		4		1

We now notice that any number in Pascal's Triangle is the sum of the numbers in the row above which are immediately to the left and right of it, as indicated by the pairs of arrows and by the working of the multiplications. This law of formation persists indefinitely, and enables us to form successive rows. Thus, in the fifth row, we begin with a 1 half a place to the left of the initial 1 at the left-hand end of the fourth row, the next number, to the right of the 1 in the fourth row and to the left of the 4 is 5, the result of adding 1 and 4, then, below and to the right of the 4 and below and to the left of the 6 comes $4 + 6 = 10$, and the remaining entries are similarly $6 + 4 = 10$, $4 + 1 = 5$, and finally a 1 half a place to the right of the 1 in the fourth row. Similarly the sixth row starts with a 1 on the left, followed by $1 + 5 = 6$, $5 + 10 = 15$, $10 + 10 = 20$, $10 + 5 = 15$, $5 + 1 = 6$ and the usual 1 on the extreme right. Continuing this process as far as the eighth row, the arrangement becomes

$$
\begin{array}{c}
1 \\
1 \quad 1 \\
1 \quad 2 \quad 1 \\
1 \quad 3 \quad 3 \quad 1 \\
1 \quad 4 \quad 6 \quad 4 \quad 1 \\
1 \quad 5 \quad 10 \quad 10 \quad 5 \quad 1 \\
1 \quad 6 \quad 15 \quad 20 \quad 15 \quad 6 \quad 1 \\
1 \quad 7 \quad 21 \quad 35 \quad 35 \quad 21 \quad 7 \quad 1 \\
1 \quad 8 \quad 28 \quad 56 \quad 70 \quad 56 \quad 28 \quad 8 \quad 1
\end{array}
$$

and the significance of this is that the numbers in any row, say the nth row, give us the coefficients of the various powers in $(a + x)^n$, thus

$$
\begin{aligned}
(a+x)^5 &\equiv a^5+5a^4x+10a^3x^2+10a^2x^3+5ax^4+x^5 \\
(a+x)^6 &\equiv a^6+6a^5x+15a^4x^2+20a^3x^3+15a^2x^4 \\
&\quad +6ax^5+x^6 \\
(a+x)^7 &\equiv a^7+7a^6x+21a^5x^2+35a^4x^3+35a^3x^4 \\
&\quad +21a^2x^5+7ax^6+x^7 \\
(a+x)^8 &\equiv a^8+8a^7x+28a^6x^2+56a^5x^3+70a^4x^4 \\
&\quad +56a^3x^5+28a^2x^6+8ax^7+x^8 \quad .. \quad (6)
\end{aligned}
$$

We are now in a position to appreciate that it may be possible to find a general formula for $(a + x)^n$ which will include all the identities (6), and the lower powers of $(a + x)$ which we multiplied out, as special cases. This formula is well known as the 'Binomial Theorem', and is

$$
(a+x)^n \equiv a^n + \frac{n}{1}\cdot a^{n-1}x + \frac{n}{1}\cdot\frac{(n-1)}{2}\, a^{n-2}x^2 + \frac{n}{1}\cdot\frac{(n-1)}{2}\cdot\frac{(n-2)}{3}\, a^{n-3}x^3 + \ldots \quad (7)
$$

The coefficient of the general term $a^{n-r}\, x^r$ in the identity (7) is

$$
{}^nC_r = \frac{n}{1}\cdot\frac{(n-1)}{2}\cdot\frac{(n-2)}{3}\cdot\ldots\frac{(n-r+1)}{r} \quad (8)
$$

that is to say, it has r factors in numerator and denominator, and the sum of each numerator and its associated denominator is $(n + 1)$. The peculiar symbol ${}^nC_r$ is used for this coefficient because it happens to be the number of possible choices, or combinations, of r objects out of n when the objects are all different and we are not interested in the order in which the objects are chosen. The law of formation associated with Pascal's Triangle can be expressed in the form

$$
{}^nC_r + {}^nC_{r-1} = {}^{n+1}C_r \quad .. \quad .. \quad .. \quad (9)
$$

Hitherto we have assumed that n is a positive integer and that there is no restriction whatever on a and x. When n is a positive integer, we see from Equ. (8) that putting $r = n$ gives us ${}^nC_n = 1$ as we should expect from Pascal's Triangle, and that if r has any value greater than n, one of the factors in the numerator of the right-hand side of Equ. (8) will be zero and, therefore, the identity (7) gives us $(n + 1)$ terms on the right-hand side and no more. This again confirms what we have already found out for values of n up to 8 by means of Pascal's Triangle.

If in Identity (7) we now put 2 for a and y for x, and n has the successive values 6, 5, 4, 3 and 2, we obtain

$$
\begin{aligned}
(2+y)^6 &\equiv y^6+12y^5+60y^4+160y^3+240y^2+192y+64 \\
(2+y)^5 &\equiv y^5+10y^4+40y^3+80y^2+80y+32 \\
(2+y)^4 &\equiv y^4+8y^3+24y^2+32y+16 \\
(2+y)^3 &\equiv y^3+6y^2+12y+8 \\
(2+y)^2 &\equiv y^2+4y+4 \quad (10)
\end{aligned}
$$

and if we now substitute from the Identities (10) into Identity (4), and simplify, we confirm the correctness of Identity (5), derived originally by the division process of last month.

We now have to consider what use (if any) can be made of the Identity (7) when n is no longer restricted to being a positive integer. If we are thus to remove the restriction on n, the price we have to pay for it is that the expression on the right-hand side of Identity (7) becomes an infinite series instead of a finite one and, therefore, in order to obtain useful information, we must manipulate a and x in such a way that the series is convergent. The simplest way to ensure this is to say that a must be 1, and x if real must be between -1 and $+1$ (*not* equal to -1 or $+1$), or if complex must have its modulus less than (*not* equal to) 1. We shall show that certain preliminary manipulations are possible which make this restriction on a and x much less than it appears at first sight to be.

Applying Identity (7) with $a = 1$, x restricted as just indicated, and various values of n, we obtain such series as

$$\left.\begin{aligned}(1+x)^{-1} &\equiv 1-x+x^2-x^3+\ldots\\(1+x)^{-2} &\equiv 1-2x+3x^2-4x^3+\ldots\\(1+x)^{1/2} &\equiv 1+\tfrac{1}{2}x-\tfrac{1}{8}x^2+\tfrac{1}{16}x^3-\ldots\\(1+x)^{0.7} &\equiv 1+0{\cdot}7x-0{\cdot}105x^2+0{\cdot}0455x^3\ldots\end{aligned}\right\} \qquad (11)$$

All the series (11) are rapidly convergent if $|x|$, the modulus of x, is sufficiently small, and converge more slowly as $|x|$ approaches 1. If therefore we require an expansion for $(a+x)^n$, it is important to carry out preliminary manipulations as far as possible to make the series actually used rapidly convergent. Thus, if we are concerned with $(a+x)^n$ when $|a|$ is numerically greater than $|x|$, we take a outside the bracket and write

$$(a+x)^n = a^n\left(1+\frac{x}{a}\right)^n \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

The expansion actually used will then be

$$(a+x)^n = a^n\left[1+\frac{n}{1}\cdot\frac{x}{a}+\frac{n}{1}\cdot\frac{n-1}{2}\cdot\left(\frac{x}{a}\right)^2+\ldots\right] \quad (13)$$

If however the modulus of x is greater than that of a, we have to write

$$(a+x)^n = x^n\left(1+\frac{a}{x}\right)^n \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

and the expansion actually used, instead of the series (13) is

$$(a+x)^n = x^n\left[1+\frac{n}{1}\cdot\frac{a}{x}+\frac{n}{1}\cdot\frac{n-1}{2}\cdot\left(\frac{a}{x}\right)^2+\ldots\right] \quad (15)$$

but the series (15) differs from the series (13) in that descending powers of x are involved instead of ascending powers. Whether the appropriate series is given by Equ. (13) or Equ. (15), it will converge slowly if $|a|$ and $|x|$ are nearly equal. In such cases, convergence can often be made more rapid by writing

$$b = a + y;\ \xi = x - y \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

so that

$$a + x = b + \xi \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

and choosing y in such a way that $|b|$ and $|\xi|$ are not nearly equal. Then we can write

$$(a+x)^n = (b+\xi)^n$$

$$= b^n\left[1+\frac{n}{1}\cdot\frac{\xi}{b}+\frac{n(n-1)}{1\,.\,2}\left(\frac{\xi}{b}\right)^2+\ldots\right] \quad (18a)$$

$$\text{or} = \xi^n\left[1+\frac{n}{1}\cdot\frac{b}{\xi}+\frac{n}{1}\cdot\frac{n-1}{2}\left(\frac{b}{\xi}\right)^2+\ldots\right] \quad (18b)$$

the first being used if $|\xi|<|b|$ and the second if $|\xi|>|b|$. If $x \approx ae^{j\theta}$, y can conveniently be taken as $ae^{j\theta}$; in this case

$$b = a(1+e^{j\theta}) = 2a\cos\tfrac{1}{2}\theta\,.\ e^{\frac{1}{2}j\theta}$$

and ξ will be small compared to b unless θ is nearly π, in which case the procedure of Equ. (16) may have to be repeated with another value of y.

If we wish to use the binomial theorem to calculate $(2{\cdot}1)^{1/3}$, then since

$$1{\cdot}3^3 = 2{\cdot}197, \quad 1{\cdot}28^3 = 2{\cdot}097152$$

we can write $(2{\cdot}1)^{1/3}$ in any of the following three forms:

$$1{\cdot}1^{1/3}\left(1+\frac{1}{1{\cdot}1}\right)^{1/3}$$

$$1{\cdot}3\left(1-\frac{0{\cdot}097}{2{\cdot}197}\right)^{1/3}$$

$$1{\cdot}28\left(1+\frac{0{\cdot}002848}{2{\cdot}097152}\right)^{1/3}$$

before applying the series (13). The last form would be much the most effective, as the series is rapidly convergent, and the factor 1·28 outside is exact whereas, in the first form, the series for $\{1+(1/1{\cdot}1)\}^{1/3}$ would be slowly convergent, and it would be necessary to determine $1{\cdot}1^{1/3}$ separately.

As long as care is taken, by manipulations of the kind indicated in Equs. (13) and (16) where necessary, to make the binomial series actually used rapidly convergent, the binomial theorem can be satisfactorily used for any value of n, positive, negative, integral or fractional. When n is a positive integer, and not too large, Pascal's triangle is the easiest way to obtain the coefficients, and the number of terms has the finite value $(n+1)$.

PREMIUMS FOR TECHNICAL WRITING

The Radio Industry Council has announced awards of 25 guineas to the authors of the following articles:

"New Types of D.C. Amplifier," D. J. R. Martin, B.Sc., A.Inst.P., *Electronic & Radio Engineer*, January and February 1958.

"A New High-Efficiency High-Power Amplifier," V. J. Tyler, B.Sc., A.M.I.E.E., *The Marconi Review*, 3rd Quarter, 1958.

"A Survey of Microwave Radio Communication," W. J. Bray, M.Sc.(Eng.), M.I.E.E., *Electronic Engineering*, May 1958.

"All Travelling-Wave Tube Systems," S. Fedida, B.Sc.(Eng.) (Hons), A.C.G.I., A.M.I.E.E., *Electronic Engineering*, May 1958.

"Electronic Developments at Very Low Temperatures," E. Mendoza, M.A., Ph.D., *British Communications and Electronics*, April 1958.

"Analogue Computation," E. Lloyd Thomas, B.Sc., A.C.G.I., M.I.E.E., A.F.R.Ae.S., *British Communications and Electronics*, May 1958.

The awards will be presented at a lunch on 15th April.

MANUFACTURERS' LITERATURE

G.E.C. Valves and Cathode-Ray Tubes for Industry. Pp. 40. The 1959 catalogue of industrial valves and c.r.ts manufactured by The M-O Valve Co. Ltd.
The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2.

Some Properties of Wiggin Electrical Resistance Materials. A 35-page publication describing the properties and uses of electrical resistance materials (Brightray series, Ferry, Monel and nickel), manufactured by Henry Wiggin & Co. Ltd.
Henry Wiggin & Co. Ltd., Publications Department, 20 Albert Embankment, London, S.E.11.

Kingsnorth Filling and Sealing Compounds for Electrical Components Manufacture. A 16-page booklet giving details of the applications, specifications, and tests for purity and performance of the Kingsnorth bituminous filling and sealing compounds.
Berry Wiggins & Co. Ltd., Field House, Breams Buildings, Fetter Lane, London, E.C.4.

Transistor Junction Temperature

CONTINUOUS MEASUREMENT IN CLASS C CIRCUITS

By H. Sutcliffe, M.A., A.M.I.E.E. and D. J. Matthews, B.Sc., Graduate I.E.E.*

SUMMARY. *A circuit is described which can be interposed between the driving source and the output stage in class C transistor circuits in order to measure the temperature-dependent base leakage current. The circuit permits a continuous measurement of junction temperature without disturbing the normal operation of the class C circuit.*

A limitation to the power-handling capacity of transistor circuits arises because the temperature at the transistor junction must not be allowed to exceed a certain maximum value. Junction temperature is dependent on the ambient temperature, the thermal arrangements for removing heat from the interior of the transistor, and the electrical power dissipated in the transistor. Transistor manufacturers supply adequate information concerning the thermal properties of their products, and for class A or B circuits (where the electrical power dissipation is known) it is a straightforward matter to arrive at a design in which the rise of junction temperature will not be excessive. This situation does not occur with the class C circuits which are used, for instance, in d.c. converters[1,2]. It is difficult, with such circuits, to make an accurate estimate or measurement of power dissipation, since the transistor is either passing a large current at low voltage (the 'on' state) or a very small current at high voltage (the 'off' state). Charge storage effects during change-over from 'on' to 'off' states also contribute to the difficulty of finding the power dissipation[3].

In class C circuits then, at the design and development stage, a direct measurement of junction temperature is particularly desirable. The method described in this article is an extension of the well-known technique of using the temperature-dependent reverse current of p-n junctions as a measure of junction temperature[4]. The measuring circuit was devised in order to perform measurements on transistors in their normal working conditions, without any interruption or major change to the driving circuit or load connections.

Circuit for the Measurement of Temperature-Dependent Current

The circuit is shown in Fig. 1 and typical waveforms are shown in Fig. 2. The diodes ensure that during the 'on' period, when the base driving current i_{in} is large and negative, this current flows entirely through D_1. No current flows through R and the voltage input to the differential amplifier is zero. During the 'off' period, after an initial transient caused by hole storage[5], the transistor is in a condition with its base at a positive

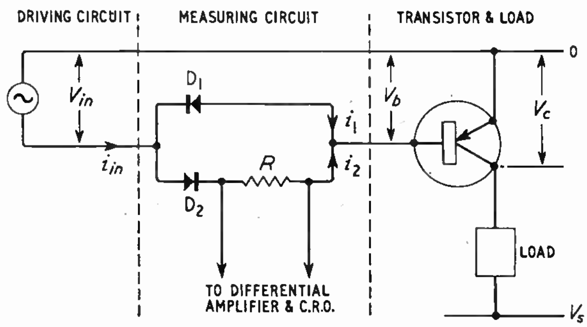
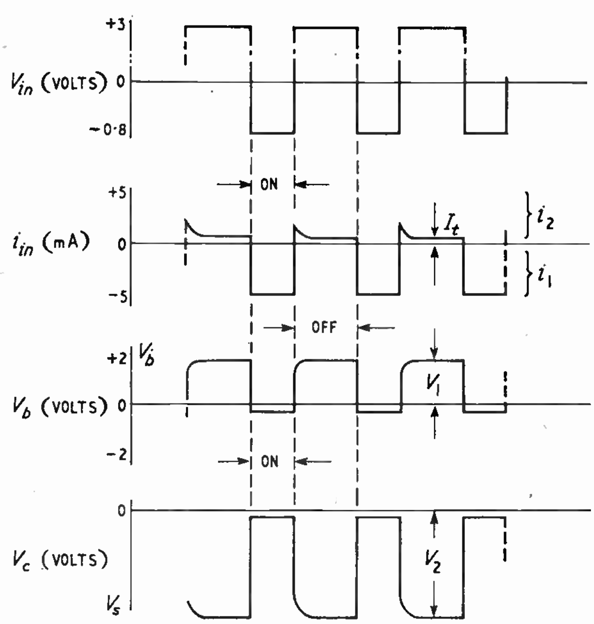


Fig. 1. *A circuit for the measurement of junction temperature interposed between the driving circuit and a transistor operating in class C*

Fig. 2. *Waveforms in Fig. 1. Current I_t depends on junction temperature*



* *Department of Electrical Engineering, University of Bristol.*

potential with respect to both emitter and collector. The base leakage current I_t, which is a temperature-dependent quantity, flows entirely through D_2 and R. The voltage input to the differential amplifier is RI_t and the value of I_t may be deduced from the c.r.o. display and the sensitivity of the amplifier.

The differential amplifier is not required to be of the directly-connected type, since it need only measure the amplitude of the step RI_t which results when i_2 falls from I_t to zero. The choice of diodes D_1 and D_2 depends on the power level of the transistor-circuit and on the requirement that the reverse current in the diodes must be small compared with I_t. Silicon junction diodes (e.g. type ZS 20A) permit driving currents of up to 100 mA with only about 1-V loss of p.d. in the forward direction, and have reverse currents which at room temperature are very much less than 1 μA. Thermionic diodes (CV.140) have also been used in low-power circuits, but require a slight modification to the design. Sufficient resistance, or a battery, must be placed in series with each thermionic diode to ensure that the p.d. across the conducting branch of the measuring circuit is adequate to produce current cut-off in the branch which is required to be non-conducting.

The choice of the resistance R represents a compromise between conflicting requirements. Too small a value will require an elaborate differential amplifier with a good rejection ratio[6], since the differential voltage RI_t will be much less than the variation in v_b. Too large a value of R will result in an excessive duration of the time interval during which the stored charge in the transistor base is being extracted. A satisfactory compromise is to select a value of R such that RI_t is in the region of one-fifth of the positive excursion (V_1) of v_b.

The Temperature-Sensitive Current I_t

It has been shown in the previous section that the circuit measures the current I_t which flows into the transistor base during a period when the emitter-to-base voltage V_e is $-V_1$ volts and the collector-to-base voltage V_c is $(-V_1 - V_2)$ volts, as shown in Fig. 2. I_t is a function of temperature and of V_e and V_c. It is necessary, because of the spread in transistor characteristics, to prepare a calibration curve for each transistor by measuring I_t as a function of temperature with values of V_e and V_c appropriate to the circuit under investigation. It will be found that, since I_t is the reverse current through a pair of p-n junctions and so varies approximately exponentially with temperature, a plot of I_t against temperature on log-linear graph paper will be almost a straight line. Three constant-temperature enclosures at say 30, 50 and 70 °C are thus adequate to establish the calibration and a suitable 'Three Holer' fluid bed has been developed for this purpose. A typical calibration curve for an OC 72 transistor was a straight line passing through the points 30 μA at 30 °C and 770 μA at 70 °C, with $V_e = -2$ V and $V_c = -12$ V. Thus a 1 °C change of junction temperature changes I_t by approximately 8 %.

It has so far been assumed that the voltages V_1 and V_2 are known constants, but during any experimental investigation it will be necessary to make modifications and adjustments to the circuit which may alter the values of these quantities considerably. It thus becomes necessary to assess the extent to which I_t is affected by changes of V_e and V_c from the nominal values used in preparing the calibration curve. A particular illustration of this assessment is shown in Fig. 3, which is typical of results obtained with OC 72 transistors. The nominal calibration point is at P, but within the hatched region the value of I_t differs from that at P by less than 5 %. It follows that, for pairs of values of V_e and V_c within this extensive region, the error in the assessment of junction temperature will be less than 0·7 °C. It should be mentioned that not all types of

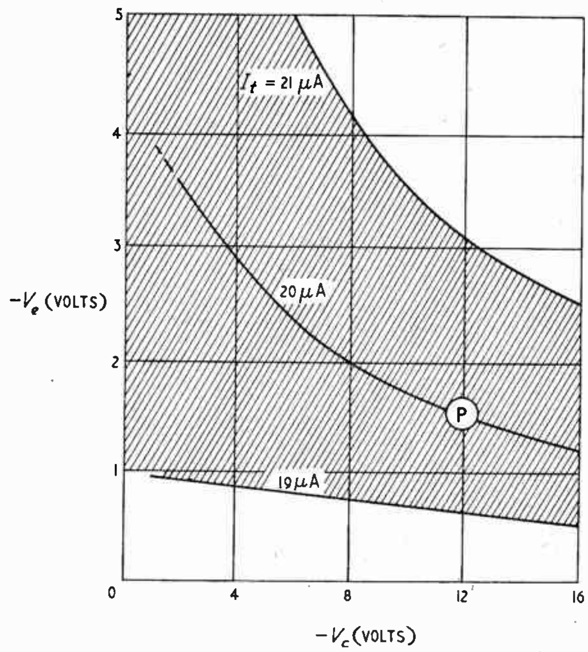


Fig. 3. *Lines of constant I_t with V_e, V_c as ordinates, for an OC 72 at 45 °C*

transistor show this favourable independence of leakage current with reverse voltage, and in such cases the convenience of the circuit is restricted to situations where the values of V_1 and V_2 have little variation.

Conclusions

The circuit provides an instantaneous and continuous measurement of junction temperature in class C transistor circuits and has proved extremely useful in an investigation into switching circuits. It is of particular value as a warning device when a transistor circuit is being adjusted near its maximum loading.

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Group Delay and Group Velocity

CONCEPT IN TERMS OF THE TRANSFER FUNCTION OF A NETWORK

By W. Proctor Wilson, C.B.E., B.Sc. (Eng.), M.I.E.E.*

We consider two unit cisoidal† electromotive forces $\exp i\omega_1 t$ and $\exp i\omega_2 t$ continuously applied to one pair of terminals of a linear network. The resultant e.m.f. E at another pair of terminals will be

$$E = |g(i\omega_1)| \exp i\{\omega_1 t - \phi(i\omega_1)\} + |g(i\omega_2)| \exp i\{\omega_2 t - \phi(i\omega_2)\} \quad .. \quad (1)$$

where $g(i\omega) = |g(i\omega)| \exp[-i\phi(i\omega)]$;

$$\phi(i\omega) = \tan^{-1}\left\{\frac{1}{i}\cdot\frac{g(i\omega) - g(-i\omega)}{g(i\omega) + g(-i\omega)}\right\}$$

is the transfer function represented by the network structure between the input and output terminal pairs.

Writing g_1, ϕ_1, g_2, ϕ_2 as abbreviations for $g(i\omega_1)$, $\phi(i\omega_1)$ etc., we may write Equ. (1) as

$$E = \exp\{i[\tfrac{1}{2}(\omega_1 + \omega_2)t - \tfrac{1}{2}(\phi_1 + \phi_2)]\} [|g_1| \exp \tfrac{1}{2} i\{(\omega_1 - \omega_2)t - (\phi_1 - \phi_2)\} + |g_2| \exp -\tfrac{1}{2} i\{(\omega_1 - \omega_2)t - (\phi_1 - \phi_2)\}] \quad (2)$$

We now let
$$\left.\begin{aligned}\omega_1 + \omega_2 &= 2\omega\\ \omega_1 - \omega_2 &= 2\delta\omega\end{aligned}\right\} \quad .. \quad (3)$$

and correspondingly
$$\left.\begin{aligned}\phi_1 + \phi_2 &= 2\phi\\ \phi_1 - \phi_2 &= 2\delta\phi\end{aligned}\right\} \quad .. \quad (4)$$

We shall subsequently let $\delta\omega$ and $\delta\phi$ tend to zero.

Making use of Equs. (3) and (4), we now express Equ. (2) in terms of

$$E = \exp\{i(\omega t - \phi)\}[|g\{i(\omega + \delta\omega)\}| \exp i(\delta\omega.t - \delta\phi) + |g\{i(\omega - \delta\omega)\}| \exp -i(\delta\omega.t - \delta\phi)] \quad .. \quad (5)$$

$$= \exp i\omega\left(t - \frac{\phi}{\omega}\right)\left[|g\{i(\omega + \delta\omega)\}| \exp i\delta\omega\left(t - \frac{\delta\phi}{\delta\omega}\right) + |g\{i(\omega - \delta\omega)\}| \exp -i\delta\omega\left(t - \frac{\delta\phi}{\delta\omega}\right)\right] \quad (6)$$

As $\delta\omega$ approaches zero, we may express (6) by the first order approximation

$$E \approx \exp i\omega\left(t - \frac{\phi}{\omega}\right)\left[\left\{|g| + \delta\omega\frac{d|g|}{d\omega}\right\}\exp i\delta\omega\left(t - \frac{d\phi}{d\omega}\right) + \left\{|g| - \delta\omega\frac{d|g|}{d\omega}\right\}\exp -i\delta\omega\left(t - \frac{d\phi}{d\omega}\right)\right] \quad .. \quad (7)$$

$$\approx 2\exp i\omega\left(t - \frac{\phi}{\omega}\right)\left[|g|\cos\delta\omega\left(t - \frac{d\phi}{d\omega}\right) + i\delta\omega\frac{d|g|}{d\omega}\sin\delta\omega\left(t - \frac{d\phi}{d\omega}\right)\right] \quad .. \quad .. \quad (8)$$

where $|g|$ is an abbreviation for $|g(i\omega)|$.

* *Research Department, British Broadcasting Corporation.*

† The term 'cisoidal', an abbreviation for '$\cos + i\sin$', is due to G. A. Campbell, who used it to denote the class of simple harmonic oscillations typified by $\exp(i\omega t)$.

The input e.m.f. E_0 corresponding to the output e.m.f. E in (8) is derived by making $g(i\omega)$ equal to unity, and is simply

$$2\exp i\omega t\cos\delta\omega.t \quad .. \quad .. \quad .. \quad .. \quad (9)$$

From a comparison of Equ. (8) with Equ. (9) it is seen that the cisoidal component, $\exp(i\omega t)$, in Equ. (9) has been retarded in time by an amount

$$\tau_0 = \frac{\phi}{\omega}$$

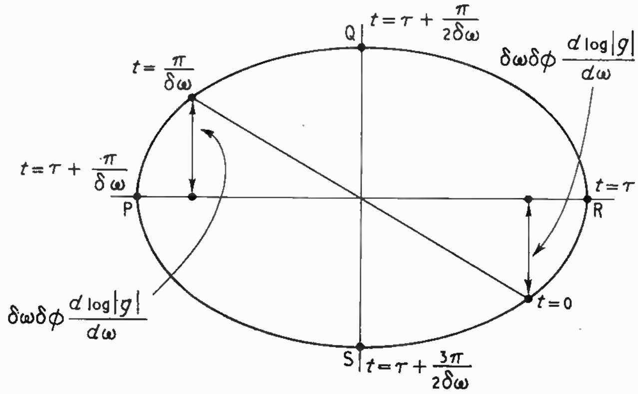
The 'low-frequency' cosine component, $\cos\delta\omega.t$, which may be regarded as a modulating signal, has been converted by the network into a complex signal. The in-phase component of this signal is the predominant one, the quadrature signal vanishing with $\delta\omega$, (provided that $d|g|/d\omega$ be finite). Both components of this complex signal are equally retarded in time by an amount

$$\tau = \frac{d\phi}{d\omega}$$

The retardation or delay time $\tau = (d\phi/d\omega)$ is often called 'group delay'; it is closely related to the classical concept in physics of 'group velocity', as will be shown later. Another name, used by American writers extensively, is 'envelope delay'. This is also a logical definition; the delay is just that imposed by the network on the transmission of the envelope of the beat pattern resulting from the superposition of the two simple harmonic motions differing in frequency by $\delta\omega/2\pi$.

The transformation of the signal $2\exp i\omega t\cos\delta\omega.t$ in (9) by the network of transfer function $g(i\omega)$ may be

Fig. 1. $\tau = d\phi/d\omega$, $g(i\omega) = |g|\exp(-i\phi)$. Loci: ellipse PQRS is locus of $Z = \cos\delta\omega(t - \tau) + i\delta\omega\{d\log|g|/d\omega\}\sin\delta\omega(t - \tau)$. Straight line PR is locus of Z when $g(i\omega)$ is of the form $\exp - i\omega\tau_0$, where τ_0 is a constant



interpreted conveniently by rewriting Equ. (8) in the form

$$E = 2|g|\exp i\omega\left(t - \frac{\phi}{\omega}\right)\left[\cos\delta\omega\left(t - \frac{d\phi}{d\omega}\right)\right.$$
$$\left. + i\delta\omega\frac{d\log|g|}{d\omega}\sin\delta\omega\left(t - \frac{d\phi}{d\omega}\right)\right] \quad .. \quad .. \quad (10)$$

since $\frac{d|g|}{d\omega}\Big/|g| = \frac{d\log|g|}{d\omega}$

We also note that the amplitude of the input signal in Equ. (9) is modified by a 'scale factor' $|g|$, the modulus of the transfer function. The locus of the modulating function $\cos\delta\omega.t$, which is that part of the real axis between -1 and $+1$, is transformed by the effect of the network into an ellipse with semi-major axis of unit length, and semi-minor axis the length of which is $\delta\omega.(d\log|g|/d\omega)$. This locus is shown in Fig. 1.

We now consider the network with transfer function $g(i\omega)$ as representing the effect of dispersion in a line or cable with a propagation path of, say, length x. The phase retardation of the network at frequency $\omega/2\pi$ will therefore be $\phi = (\omega x/v)$, where v is the phase velocity of the network. The group delay of the modulating signal will now correspond to a *group velocity* $u = (x/\tau)$, say.

We can therefore write

$$u = \frac{x}{\dfrac{d\phi}{d\omega}} = \frac{1}{\dfrac{d}{d\omega}\left(\dfrac{\omega}{v}\right)} = \frac{v}{1 - \dfrac{\omega}{v}\cdot\dfrac{dv}{d\omega}} \quad .. \quad (11)$$

The last expression in (11) is the classic equation relating the group velocity to the phase velocity as a function of frequency.

An expression for $\tau(i\omega)$ in terms of the function $g(i\omega)$ and of its conjugate may be obtained by differentiating $\phi(i\omega)$, as defined in (1), with regard to ω. Alternatively, use may be made of one of an interesting pair of relations derived by Gouriet in a recent I.E.E. Monograph††. These, following the notation in this paper, are

$$\left.\begin{aligned}\tau(p) &= \mathrm{Re}\{g'(p)/g(p)\}\\ \frac{d\log|g(p)|}{dp} &= \mathrm{Im}\{g'(p)/g(p)\}\end{aligned}\right\} \quad .. \quad .. \quad (12)$$

where $p \equiv i\,\omega$.

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

Transmission of Power in Radio Propagation

By James R. Wait, M.A.Sc., Ph.D.*

SUMMARY. *Some remarks concerning the theoretical foundations of the transmission-loss concept are made. It is emphasized that the influence of the ground on the input resistances of the transmitting and receiving aerials must be accounted for.*

Norton introduced the concept of transmission loss in radio propagation almost a decade ago[1]. It is a natural and logical method to describe the characteristics of a radio circuit. Some difficulty has been encountered, however, when the aerials are in close proximity to the ground. If the intended meaning of transmission loss is to be preserved, the losses associated with the imperfect ground in the immediate vicinity of the aerial must be included. Furthermore, the patterns of the aerials as well as their power gains must be properly accounted for.

It is the purpose of this article to formulate the problem and provide an expression for the transmission loss L. In order to facilitate the presentation of propagation data, this loss L will be split into two parts, one of which L_p can be described logically as a '*propagation transmission loss*' and the remaining part L_g accounts for the *proximity* of the aerials to the ground.

The writer admits at the outset that the division of the transmission loss into parts will always be somewhat arbitrary, but prefers the present scheme. His reasoning is given in what follows.

Statement of the Problem

To simplify the discussion, the transmitting and receiving aerials are both considered to be vertical electric dipoles or small linear wire elements. The dipoles designated as A and B are located at heights h_a and h_b over a smooth spherical earth of finite conductivity. The separation between A and B is the great circle distance d as indicated in Fig. 1. The time factor is taken to be $\exp(i\omega t)$ where ω is the angular frequency.

An expression for the mutual impedance Z between

* *National Bureau of Standards, Boulder Laboratories, Colorado, U.S.A.*

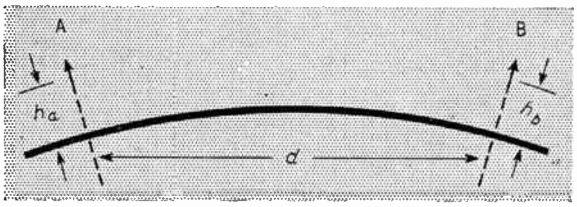


Fig. 1. *Vertical dipoles A and B located at heights h_a and h_b above a smooth spherical earth of finite conductivity. Their separation is denoted by the great-circle distance d*

these dipoles can be written in the form

$$Z_m = 2\ Z_0\ W \quad \ldots \quad (1)$$

where Z_0 is the mutual impedance between the same dipoles if both are located in free space and separated by a linear distance d and oriented for maximum coupling and W is a certain 'attenuation function'. It is well known that[2]

$$Z_0 = \frac{i\mu\omega l_a l_b}{4\pi d}\left[1 + \frac{1}{ikd} - \frac{1}{k^2d^2}\right] e^{-ikd} \quad \ldots \quad (2)$$

where l_a and l_b are the effective lengths of the dipoles, $k = 2\pi/\text{wavelength}$, and $\mu = 4\pi \times 10^{-7}$. The function W depends on the nature of the propagation medium. It has been discussed frequently in the literature for various types of propagation paths.†

In the case of ground-wave propagation over a smooth earth, it becomes Norton's attenuation function $F(p)$ which is a function of the numerical distance p. When the ground is flat and perfectly conducting and the aerial heights approach zero, $W = 1$ or $Z = 2Z_0$. In the case of a flat earth of finite conductivity σ_1 and dielectric constant ϵ_1, W can be written

$$W \simeq F(p)\, f(h_a)\, f(h_b)$$

where[3,4] $F(p) \simeq [1 - i\sqrt{\pi p}\ e^{-p} \text{ erfc } (ip^{\frac{1}{2}})] \quad \ldots \quad (3)$

with $p = -\dfrac{ikd}{2}\Delta^2$ and $\Delta^2 = \dfrac{k^2}{k_1{}^2}\left(1 - \dfrac{k^2}{k_1{}^2}\right)$

and $k_1{}^2 = \epsilon_1\mu\omega^2 - i\sigma_1\mu\omega$.

For $p \ll 1$, $F(p) \simeq 1$ and for $p \gg 1$, $F(p) \simeq -1/(2p)$ $f(h_a)$ and $f(h_b)$ are height–gain functions which approach

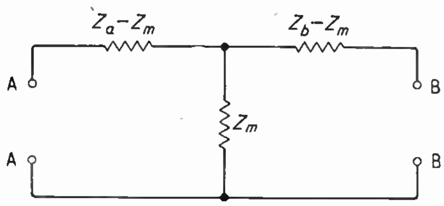


Fig. 2. *Equivalent T-network of the aerial system*

unity for zero heights. For small heights they have the approximate form

$$\left.\begin{aligned} f(h_a) &\simeq 1 + i\Delta kh_a \\ f(h_b) &\simeq 1 + i\Delta kh_b \end{aligned}\right\} \quad \ldots \quad (4)$$

subject to $(kh_a)^2$ and $(kh_b)^2 \ll kd$. Since Δ is usually small compared to unity, $f(h_a)$ and $f(h_b)$ vary relatively slowly with heights. For arbitrary heights and distances, other forms of W are available[5].

It now proves convenient to introduce the 'T' representation of a four-terminal network, illustrated in Fig. 2.

The accessible terminals of the two aerials are represented by A and B. By definition, the central member of the T-network is equal to the mutual impedance Z_m. It is noted that the input impedance at A, with B open circuited, is Z_a. Conversely, the input impedance at B, with A open circuited, is Z_b.

Unlike Z_m, the quantities Z_a and Z_b are very much dependent on the immediate environment of the aerials. The real parts of Z_a and Z_b play an important role and are described as the input resistances R_a and R_b, respectively. It is convenient to write

$$\left.\begin{aligned} R_a &= R_{0a}\,(1 + \Delta_a) \\ \text{and } R_b &= R_{0b}\,(1 + \Delta_b) \end{aligned}\right\} \quad \ldots \quad (5)$$

where R_{0a} and R_{0b} are the input resistances of the dipoles if they were located in free space. In terms of the effective length l_a

$$R_{0a} = 80\pi^2\,(l_a/\lambda)^2 \text{ ohms} \quad \ldots \quad (6)$$

and similarly for R_{0b}. If the dipoles were located on a

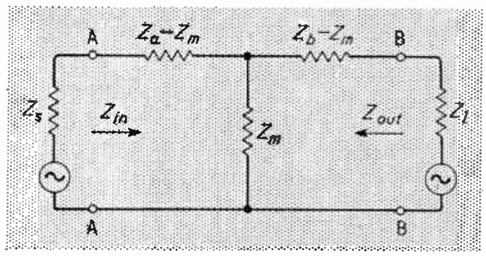


Fig. 3. *Complete equivalent circuit for calculating the power transfer between dipoles A and B*

flat perfectly-conducting ground, it follows from elementary reasoning that

$$R_a = 2R_{0a} \text{ or } \Delta_a = 1$$

and similarly for R_b. For intermediate heights over a perfectly conducting ground $\Delta_a R_{0a}$ can be regarded as the mutual resistance between the source dipole at height h_a and its perfect image at a distance h_a below the ground plane. Δ_a and Δ_b may also include losses associated with non-ideal ground planes beneath the aerials and can be greater than unity.

The Power Calculation

The relevant quantities needed to calculate the power transfer between the dipoles are

$$\begin{aligned} R_a &= \text{Real part of } Z_a \\ &= 80\pi^2\,(l_a/\lambda)^2\,(1 + \Delta_a), \quad \ldots \quad (7) \\ R_b &= \text{Real part of } Z_b \\ &= 80\pi^2\,(l_b/\lambda)^2\,(1 + \Delta_b), \quad \ldots \quad (8) \end{aligned}$$

$$\text{and } Z_m = i\,\frac{60\pi}{\lambda}\,\frac{l_a l_b}{d}\left(1 + \frac{1}{ikd} - \frac{1}{k^2d^2}\right)(2W)\,e^{-ikd} \quad (9)$$

When considering power transfer, one must specify the impedance Z_l of the load and the impedance of the

† It is often described, rather loosely, as 'attenuation relative to inverse distance'.

source Z_s. The complete equivalent circuit now has the form shown in Fig. 3.

The input impedance Z_{in} at the terminals A is

$$Z_{in} = Z_a - Z_m + \frac{Z_m\,(Z_b - Z_m + Z_l)}{Z_b + Z_l} \quad \text{.. (10)}$$

and the output impedance at the terminals B is

$$Z_{out} = Z_b - Z_m + \frac{Z_m\,(Z_a - Z_m + Z_s)}{Z_a + Z_s} \quad \text{.. (11)}$$

The ratio of the power p_t supplied at the terminals A to the power p_r delivered to the load Z_l at B, is easily shown to be

$$\frac{p_t}{p_r} = \frac{\text{Re. } Z_{in}}{\text{Re. } Z_l}\left|\frac{Z_b + Z_l}{Z_m}\right|^2 \quad \text{.. .. .. (12)}$$

The maximum power delivered to the load is when conjugate matching is used. In this case, Z_l is chosen to be equal to the complex conjugate of the output impedance, namely Z_{out}^*. The transmission loss L is then defined by‡

$$L = 10 \log\,(p_t/p_r). \quad \text{.. .. .. (13)}$$

Formally, this is the complete solution of the problem. It should be noted that a knowledge of both the real and imaginary parts of Z_a and Z_b is required. These can be either measured or computed from the form of the transmitting and receiving aerials. Furthermore, the impedance Z_s of the source must also be known. Fortunately, in an actual circuit the mutual impedance Z_m is very much less than the aerial impedances Z_a and Z_b. In this case, $Z_{in} \simeq Z_a$ and $Z_{out} \simeq Z_b$, and therefore for conjugate matching, $Z_l \simeq Z_b^*$. In other words, the interaction between the aerials is neglected. The approximation is valid when

$$\left|\frac{Z_m^2}{Z_a Z_b}\right| \ll 1.$$

In the case of two dipoles this restriction can be replaced by

$$\left(\frac{\lambda}{d}\right)^2 \ll 1 \quad \text{.. .. .. (14)}$$

which is demonstrated in Appendix 2.

Assuming negligible interaction, the power ratio becomes

$$\frac{p_t}{p_r} = \left(\frac{2}{3}\right)^2 \left(\frac{4\pi d}{\lambda}\right)^2 \frac{(1 + \Delta_a)\;(1 + \Delta_b)}{(2|W|)^2} \quad \text{.. (15)}$$

The transmission loss L in decibels can then be written as the sum of two parts as follows

$$L = L_p + L_g \quad \text{.. .. .. (16)}$$

where

$$L_p = 20 \log\left(\frac{4\pi d}{\lambda}\right) - 3{\cdot}52 - 20 \log\,(2|W|) \quad (17)$$

$$\text{and } L_g = 10 \log\,[(1 + \Delta_a)\;(1 + \Delta_b)] \quad \text{.. .. (18)}$$

The two components can be interpreted as follows: L_p is a transmission loss resulting from propagation[6], L_g is a loss resulting from loss due to the proximity of the aerials to the ground plane. It is to be noted that if the aerials are raised sufficiently above the ground (i.e., h_a and $h_b > \lambda$), L_g can be replaced by zero. Then L, the total transmission loss, is equal to L_p the propagation loss.

The discussion up to this point has been restricted to vertical electric dipoles. Equ. (17) can be generalized to other aerials by writing it in the form

$$L_p = 20 \log\left(\frac{4\pi d}{\lambda}\right) - G_a - G_b - 20 \log |2W| \quad (19)$$

where G_a and G_b are the effective gains of the aerials relative to an isotropic aerial. Note that for dipoles, G_a and G_b are each 1·76, and for half-wave aerials, they are each 2·15. Equ. (19) for arbitrary aerial heights is valid only if the distance d is large compared to the aerial heights. If this condition is violated, G_a and G_b can be reduced somewhat as they may in other special situations.

Relation to Earlier Notation

In much of the past literature on radio wave propagation, the field E in microvolts per metre is presented for a vertical dipole whose moment is fixed so that it would radiate 1 kilowatt if placed on a flat perfectly-conducting ground. In terms of W, the radiation field (r.m.s.) can be written

$$E = \frac{(3 \times 10^5)\,|W|}{d_{km}} \quad \text{.. .. .. .. (20)}$$

where d_{km} is the distance in kilometres. (This representation is employed by Bremmer[7] and others.) In other words, the current on the dipole aerial is to be fixed as the aerial height varies and consequently, the input power would vary. For example, in free space the radiated power would only be ½ kilowatt.

To obtain the relation between E and the propagation transmission loss, L_p, it is noted that

$$20 \log E = 20 \log \frac{300}{d_{km}} + 60 + 20 \log |W|$$

$$\text{but } L_p = 12{\cdot}44 - 20 \log |W| + 20 \log \frac{d_{km}}{\lambda_{km}} \quad \text{.. (21)}$$

and therefore $20 \log E = 20 \log f_{kc} + 72{\cdot}44 - L_p$ where f_{kc} is the frequency in kilocycles per second. $20 \log E$ is often described as the field in decibels above 1 volt per metre for a radiated power (in the sense described above) of one kilowatt.

In recent publications[8] of the C.C.I.R. the radiation field E' is expressed in terms of a transmitter whose moment is fixed such that it would radiate 1 kilowatt if located in free space. With the same definition of W, the field (r.m.s.) is now obtained from

$$E' = \frac{2{\cdot}12 \times 10^5}{d_{km}}\,|2W| = 2E \times \frac{1}{\sqrt{2}} \quad \text{.. (22)}$$

and consequently

$$\begin{aligned} 20 \log E' &= 20 \log E + 3{\cdot}01 \\ &= 20 \log f_{kc} + 75{\cdot}45 - L_p. \end{aligned}$$

Conclusion

It has been indicated that the transmission of power between the terminals of the transmitting aerial and the terminals of the receiving aerial can be influenced appreciably by the proximity of the aerials to the ground. In view of this, it is suggested that the total transmission L be broken into two parts, the first of which is a propagation loss L_p, and the second is a proximity factor L_g. L_p is closely related to the field strength E as defined by the C.C.I.R. and is not critically

‡ The logarithms are all to the base 10.

dependent on the immediate environment of the aerial[9]. The factor L_g includes the losses associated with the finite earth conductivity and the presence of a ground screen[10-12]. For high aerials L_g vanishes.

Acknowledgment

The author has had the benefit of several discussions concerning this subject with K. A. Norton, W. Q. Crichlow, P. L. Rice and A. D. Watt. They should, however, be absolved from responsibility for any shortcomings of this article.

APPENDIX 1

It is the purpose of this appendix to provide the derivation for the transmission loss between vertical dipoles located at heights h_a and h_b over a flat perfectly-conducting earth. Certain features of these results shed some light on the concept of transmission loss. The situation is illustrated in Fig. 4.

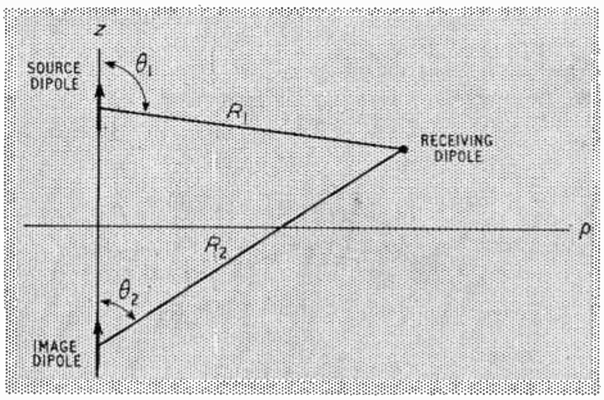


Fig. 4. *Illustrating the geometry for deriving the transmission loss between transmitted and receiving vertical dipoles located at heights h_a and h_b over a flat perfectly-conducting earth*

The transmitter is represented by a vertical electric dipole or current element with a current I_a supplied to its terminals. Its effective length is to be l_a. Choosing a cylindrical co-ordinate system (ρ, ϕ, z) the source is located at $(0, 0, h_a)$ and the ground plane is $z = 0$. The boundary conditions at the ground plane are satisfied by locating a virtual image dipole at $(0, 0 - h_a)$ with equal strength to the source dipole. The fields for $z > 0$ are then derivable from a Hertz vector which has only a z component Π which is given by

$$\Pi = C\left[\frac{e^{-ikR_1}}{R_1} + \frac{e^{-ikR_2}}{R_2}\right] \quad \ldots \quad (23)$$

where $C = \dfrac{I_a\, l_a}{4\pi i \epsilon_0 \omega}$, $\epsilon_0 = 8{\cdot}854 \times 10^{-12}$

and $R_1 = [(z - h_a)^2 + \rho^2]^{\frac{1}{2}}$

and $R_2 = [(z + h_a)^2 + \rho^2]^{\frac{1}{2}}$.

The vertical component of the electric field is then obtained from

$$E_z(\rho, z) = \left(k^2 + \frac{\partial^2}{\partial z^2}\right)\Pi. \quad \ldots \quad (24)$$

which of course is independent of ϕ. Carrying out the differentiations and grouping the terms in a certain fashion, leads to

$$E_z(\rho, z) = 2C\frac{e^{-ik\rho}}{\rho}\left[1 + \frac{1}{ik\rho} - \frac{1}{k^2\rho^2}\right] W(\rho, h_a, z) \quad \ldots \quad (25)$$

where $2\,W(\rho, h_a, z)\left[1 + \dfrac{1}{ik\rho} - \dfrac{1}{k^2\rho^2}\right]$

$$= \sin^3\theta_1 \,.\, e^{-ik(R_1-\rho)}\, N(kR_1, \theta_1) + \sin^3\theta_2 \,.\, e^{-ik(R_2-\rho)}\, N(kR_2, \theta_2)$$

where

$$N(kR, \theta) = \left(1 + \frac{1}{ikR} - \frac{1}{k^2R^2}\right) - 2\left(\frac{1}{ikR} - \frac{1}{k^2R^2}\right)\cot^2\theta \quad (26)$$

and $\sin\theta_1 = \rho/R_1$, $\sin\theta_2 = \rho/R_2$.

In free space (i.e., $h_a = z = \infty$), the terms containing the subscript 2 vanish.

The receiving dipole is now located at $\rho = d$ and $z = h_b$ and the mutual impedance Z_m between the two dipoles is obtained from

$$Z_m = -\frac{E_z(d, h_a)\, l_b}{I_a} \quad \ldots \quad (27)$$

which leads to $Z_m = 2\,Z_0\,W(\rho, h_a, h_b)$ .. .. .. (28)

where Z_0 is the mutual impedance between the two parallel dipoles located in free space and is given by

$$Z_0 = \frac{i\mu\,\omega l_a\, l_b}{4\pi d}\left(1 + \frac{1}{ikd} - \frac{1}{k^2d^2}\right)e^{-ikd} \quad \ldots \quad (29)$$

where $\mu = 4\pi \times 10^{-7}$.

The function $W(d, h_a, h_b)$, which is simply denoted W in the main text, is the ratio of the mutual impedance between the aerials located over the ground plane and twice the mutual impedance of the two aerials when located in free space. If $h_a = h_b$, W approaches 1/2 in free space and approaches unity when the aerials are located on the ground plane.

The above result for E_z can also be employed to calculate the input resistance of the dipoles but not their reactance. It can be seen that

$$R_a = -\lim_{\rho \to 0} \mathrm{Re.}\frac{E_z(\rho, h_a)}{I_a}\, l_a \quad \ldots \quad (30)$$

which readily leads to

$$R_a = R_{0a}(1 + \Delta_a) \quad \ldots \quad (31)$$

where $R_{0a} = 20\,(kl_a)^2 = 80\pi^2\,(l_a/\lambda)^2$ .. .. .. (32)

$$\text{and} \quad \Delta_a = \frac{3}{4k^2\,h_a^2}\left(\frac{\sin 2\,kh_a}{2\,kh_a} - \cos 2\,kh_a\right) \quad \ldots \quad (33)$$

in agreement with Schelkunoff[2].

It should be noted that $\Delta_a \simeq 0$ for $kh_a \gg 1$ and $\Delta_a \simeq 1$ for $kh_a \ll 1$. The expression for R_b is of the same form. If the ground plane is not a perfect conductor, R_a and R_b are much more complicated and are also functions of the size of the ground screen as well as the number of radial wires[13].

As a further check on the algebra, the dipoles are now both considered to be located on the flat perfectly-conducting plane. The vertical electric field at dipole B for p_t watts radiated at A is

$$|E_z| = \frac{(3 \times 10^5)\sqrt{p_t \times 10^{-3}}}{d}\ \text{volts/metre} \quad \ldots \quad (34)$$

in accordance with Equ. (20) when $|W| = 1$. But p_r, the power available at B, is given by

$$p_r = \frac{|E_z^2|\, l_b^2}{4R_b} \quad \ldots \quad (35)$$

where $R_b = 160\,\pi^2\,(l_b/\lambda)^2$. Therefore

$$\left(\frac{p_t}{p_r}\right)^{\frac{1}{2}} = \frac{8\pi}{3}\,\frac{d}{\lambda} \quad \ldots \quad (36)$$

and the transmission loss L is

$$L = 18{\cdot}46 + 20\log\frac{d}{\lambda} \quad \ldots \quad (37)$$

which can be rewritten as

$$L = L_p + L_g$$

$$\text{where } L_p = 12{\cdot}44 + 20\log\frac{d}{\lambda} \quad \ldots \quad (38)$$

and $L_g = 6{\cdot}02$. The equation for L_p is in agreement with Equ. (21) when $|W| = 1$ corresponding to a perfect ground plane.

APPENDIX 2

Equ. (15) for the power ratio was derived on the premise that interaction between the aerials was negligible. In other words, it was assumed that $\mathrm{Re.}Z_{in} \simeq R_a$. To indicate the significance of this assumption and to extend the range of validity of Equ. (15) it is desirable to employ a more accurate value of Re.Z_{in}. For $Z_l = Z_b^*$ it readily follows, from Equ. (10), that

$$\text{Re. } Z_{in} = R_a \left[1 - \frac{\text{Re. } Z_m^2}{2R_a R_b}\right] \quad .. \quad .. \quad .. \quad .. \quad (39)$$

Using Equs (7), (8) and (9) this can be written Re.$Z_{in} = R_a[1 + \delta(x)]$ where

$$\delta(x) = \text{Re.} \frac{9}{8x^2} \frac{\left(1 + \frac{1}{ix} - \frac{1}{x^2}\right)^2 e^{-2ix} (2W)^2}{(1 + \Delta_a)(1 + \Delta_b)} \quad .. \quad .. \quad (40)$$

and where $x = kd$. Equ. (12) for the power ratio now becomes

$$\frac{p_t}{p_r} = \frac{16}{9} x^2 \frac{(1 + \Delta_a)(1 + \Delta_b)[1 + \delta(x)]}{|2W|^2 \left[1 - \frac{1}{x^2} + \frac{1}{x^4}\right]} \quad .. \quad .. \quad (41)$$

It is immediately evident that when $x \gg 1$ or $d \gg \lambda$, the above result reduces to Equ. (15) in the text. As an interesting check, it can be seen that as x tends to zero the ratio p_t/p_r tends to 2 as it should.

For the special case of two vertical dipoles on a flat perfectly-conducting ground plane, $\Delta_a = \Delta_b = 1$ and $W = 1$ so that

$$\delta(x) = \frac{9}{8x^2}\left[\left(1 - \frac{3}{x^2} + \frac{1}{x^4}\right)\cos 2x - \frac{2}{x}\left(1 - \frac{1}{x^2}\right)\sin 2x\right] \quad .. \quad (42)$$

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Phase-Angle Measurement

NULL METHOD USING HEPTODE MIXER

By P. Kundu, M.Sc.*

SUMMARY. *This article describes a 'product' method of measuring the phase angle between two pre-adjusted out-of-phase sinusoidal voltages. The signals are applied to a heptode mixer whose differential anode current with respect to the reference value for quadrature inputs is a measure of the angle.*

An accurate determination of phase angle between a pair of sinusoids is frequently required in the investigation of the transfer functions of various systems and in the measurement of some physical quantities which can be directly related to this parameter of an a.c. signal. Although several methods[1,2,3,4,5] have been devised in recent years to deal with this common yet difficult problem, few have proved to be satisfactory from the point of view of accuracy of measurement over a wide frequency range when simple techniques, few components, and inexpensive equipment are employed. Utilization of these methods in practical measurement often imposes further limitation if the transfer process introduces distortion. The method described in this article combines the above requirements fairly satisfactorily, and can be set up and calibrated quite conveniently to measure directly phase angles between 0 and ± 180 degrees on a calibrated meter with a positive indication of sense over a 100-kc/s range. The amplitude transmission is also determined by the process of phase measurement.

* *Jadavpur University, Calcutta*

Principle of Measurement

The principle is based on the fact that when two sinusoidal signals of constant amplitude and the same frequency are applied to the oscillator and signal grids of a heptode mixer, then the d.c. component flowing in the anode circuit is a function of phase-angle difference between the two signals. Assuming that the signal grid voltage $e_{g1} = E_s \sin \omega_s t$ is small and that the oscillator grid voltage $e_{g3} = E_0 \sin(\omega_s t + \theta)$ is adjusted so that it varies linearly the mutual conductance g_m, then the the anode current i_b is given by the equation[7]:

$$i_b = (a_0 + a_1 e_{g1} + E_{c1})(b_0 + b_1 e_{g3} + E_{c3}) \qquad (1)$$

where a_0, a_1, b_0, b_1 are valve constants and E_{c1}, E_{c3} are static bias voltages.

Substituting the values of e_{g1} and e_{g3}, the anode current is given by i_b = d.c. + fundamental component + product of e_{g1}, e_{g3}. If all the a.c. components are filtered out at the output load, then the current at the anode load is given by:

$$i_b = I_{b0} + I_b = \text{d.c.} + \tfrac{1}{2}\, a_1 b_1 E_s E_0 \cos\theta$$

or $\quad i_b = \text{d.c.} + g_c E_s \cos\theta$

where $g_c = \tfrac{1}{2}\, a_1 b_1 E_0$

Here g_c is the conversion conductance of the mixer valve and so long as the valve operating point, electrode voltages and the amplitude of the oscillator grid signal E_0 are kept constant, g_c remains constant. If E_s is also maintained constant then the variable component of the anode current becomes a linear function of the cosine of the phase-angle difference θ between the voltages E_s and E_0; i.e.,

$$I_b \propto \cos\theta \qquad \ldots \qquad (2)$$

The scale of an indicating meter may, therefore, be calibrated to read θ directly. The scale divisions, however, are compressed when θ is near 0°, but almost linear for θ nearing 90°. If an additional phase shift of $\pi/2$ is introduced in one of the signals, then

$$I_b \propto \cos(\pi/2 + \theta) \qquad \ldots \qquad (3)$$

This relation yields a scale which is quite linear for θ near 0° but compressed for θ near 90°. Excessive scale compression may then be avoided by using the $\pi/2$ network only for phase angles near 0°, so that both the high and low phase angles may be measured under a similar condition of almost linear scale divisions.

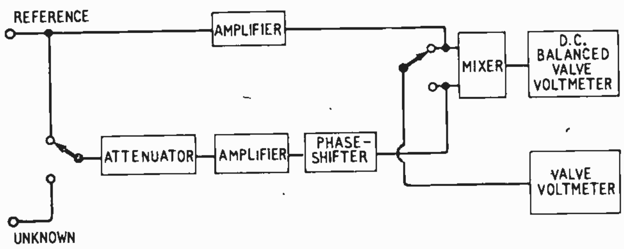


Fig. 1. *Block diagram illustrating the principle of phase-angle measurement*

It can be seen from the block diagram in Fig. 1 (illustrating the principle of measurement), that the reference signal and the unknown signal are taken through two separate channels to the oscillator and signal grids of the mixer valve. There is a provision for introducing a $\pi/2$ phase lag to the unknown signal. A high input impedance voltmeter is required to adjust the signal and oscillator grid inputs to predetermined values. The indicating meter (a centre-zero micro-ammeter) is connected in a balanced-bridge circuit. Meter deflection is then directly proportional to cos θ or sin θ, depending on whether the quadrature phase shifter is excluded from or included in the circuit.

Circuit Details

The essential features of the phase-measuring circuit are illustrated in the schematic diagram of Fig. 2. It has been used in the laboratory with a variable attenuator and an oscillator of 600-Ω impedance for measuring the phase and attenuation constant of unbalanced networks. The reference signal is taken from a constant gain amplifier V_1 to the oscillator grid of the mixer and adjusted to a predetermined value by means of the amplifier input potentiometer P_1. By selecting a suitable r.f. pentode it was found that a constant gain of 25 dB with negligible phase shift over a frequency range of 100 kc/s could be obtained with a small resistive anode load. At this upper frequency the phase lead is of the order of 40′ for maximum stray and output capacitances of about 50 pF across an anode load of 2·2 kΩ.

The unknown signal, whose phase is to be determined, is applied to the signal grid of the mixer via an attenuator, amplifier V_2 and phase shifter V_3. Identical signal-channel and reference-channel amplifiers are used so that any undesired phase shift produced will have negligible effect on the final measurement. The phase shifter consists of a resistance R and capacitance C connected in series across a balanced low-impedance source provided by a high-frequency triode phase inverter (of high g_m) with matched cathode and anode resistances.[8] A 90° phase shift is then introduced in the signal channel by properly adjusting the values of C and R. In a variable-frequency system, a desired frequency range is covered by varying C in steps and R continuously. The values of C in the different ranges were derived by considering two factors; the maximum value of frequency in a particular range, and the minimum value of resistance (see Appendix) required for the proper operation of the phase-shifter stage. Here a 1 : 10 frequency range is covered using a 50-kΩ potentiometer. The phase shifting may be excluded by taking the signal directly from the cathode of V_3 by switching S_3.

The mixer-stage electrode voltages and the magnitude of the respective inputs to the two grids are decided from the point of view of maintaining the linearity of operation. These values are not critical but, once decided upon, must be maintained; otherwise the calibration will be disturbed. The mixer load comprises a resistor bypassed by a capacitor so that the d.c. voltage is developed while the a.c. component is filtered out. The variation of this d.c. voltage (which is a function of the phase difference between the two mixer inputs), is then measured by the centre-zero micro-ammeter in the balanced-bridge circuit.

Procedure of Measurement

A preliminary zero adjustment of the meter (with the quadrature phase shifter included in the circuit) is necessary, ensuring that both channels are fed with the predetermined magnitude of the reference signal whose frequency is that at which measurements are to be made. At first, the oscillator grid of the mixer in Fig. 2 is fed with the reference-signal voltage adjusted to 4V by means of the input potentiometer P_1, while the grid of the stage V_2 is earthed by the push-button switch S_2. The indicating meter is then set to zero by varying the grid bias of the balancing valve V_5. The zero setting of the meter (with an input in the reference channel only) is required so that the effect of any undesired non-linearity of g_m — e_{g3} characteristic of the mixer is avoided. Next, the signal channel is also fed with the in-phase reference signal through the double-pole double-throw switch S_1 and the signal voltage of V_4 adjusted to 0·4 V with the input attenuator. A quadrature phase relationship between the two inputs of the mixer is then obtained by varying the resistance R of the phase shifter to get zero reading in the meter. At exactly 90° phase difference, a perfect null is indicated in the meter even if the grid of V_1 is earthed by the switch S_2; i.e., under this condition, the variation of the mixer anode current is zero and is independent of the signal-grid voltage. For variable-frequency work the phase-shifter resistance R should have a dial calibrated in frequency with a stepped decade switch for changing capacitors, so that a 90° phase shift at any frequency may be adjusted quickly over the desired frequency range. The instrument is now set up and needs no further adjustment unless the magnitude or the frequency of the voltages at the mixer inputs is changed. The signal-grid channel input is then switched over to the unknown signal by S_1, and the level is again adjusted to 0·4 V. The phase angle is then read from the meter deflection while the attenuation constant is determined from the required setting up of the signal to a fixed level with and without the network. For networks having an amplitude transmission of less than unity, the adjustment is made with a greater oscillator output and a larger attenuation of signal-grid input. The network is then switched in by S_1 and the signal-grid voltage is readjusted to 0·4 V. The attenuation and phase constant may then be read directly from the calibrated attenuator and meter respectively.

Calibration

Precision decade capacitance and resistance boxes are used as a series combination across a balanced low-impedance variable-frequency source to provide a pair of signals of known phase difference between 0 and ± 180° for the inputs of the instruments. The meter reading is then calibrated in degrees by calculating accurately the phase difference from the values of the components and the frequency used. A typical calibration curve showing the unbalanced current and the phase difference is shown in Fig. 3. The null indication

Table 1

Angle	Quadrant	Meter Deflection	
		With Network	Without Network
0° to +180°	First	Left	Left
	Second	Left	Right
0° to —180°	Third	Right	Right
	Fourth	Right	Left

corresponds to a phase difference of 0° or 90° depending on whether the $\pi/2$ network is included or not.

Since the centre-zero microammeter (shown in Fig. 2) is connected in such a way that an increase in the anode voltage is indicated by a deflection to the right, then an increase or decrease in the anode current is indicated by a deflection to the left or right respectively. When the quadrature network is included in the circuit and the phase angle between a pair of input signals lies between 0° and + 180° or 0° and — 180°, the meter deflects to the left (anticlockwise) for the leading angle and to the right (clockwise) for the lagging angle. The magnitude of the phase angle may be read directly over the 0° to ± 90° range, but for angles between 0° and ± 180° quadrant ambiguity has to be taken into account. This may be determined

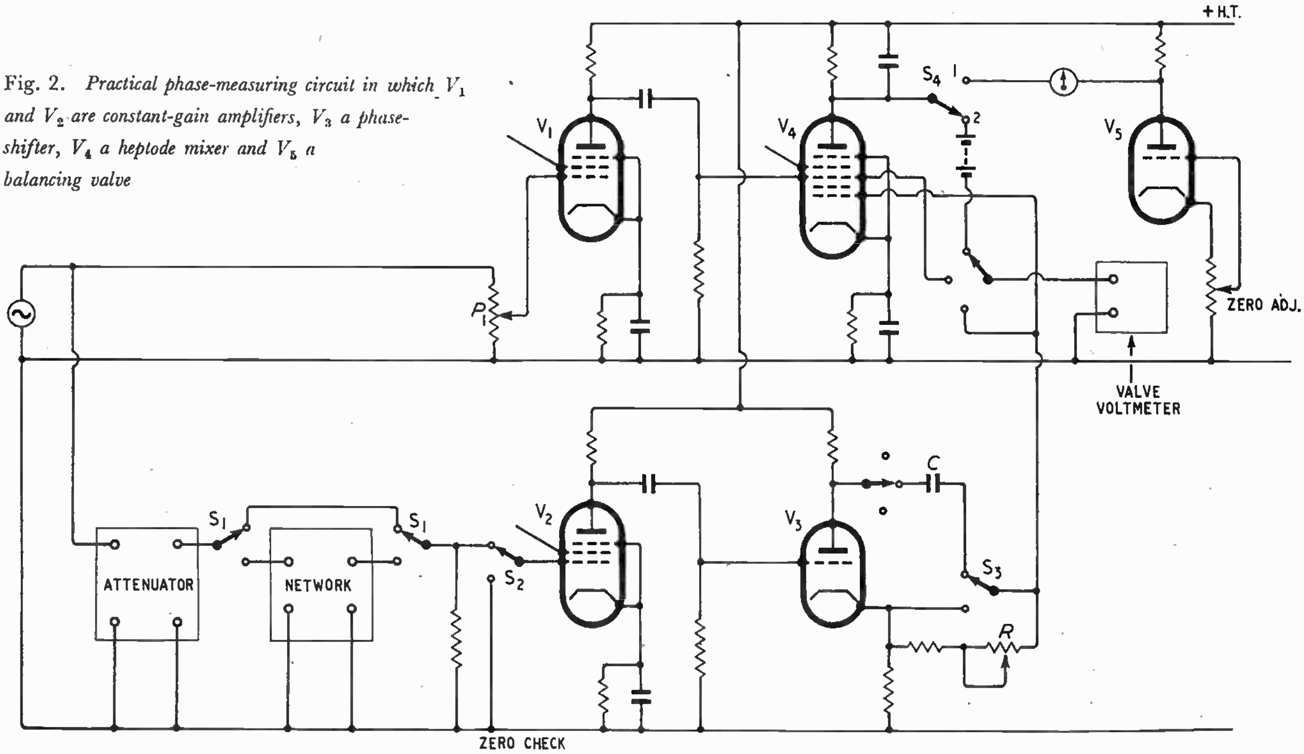


Fig. 2. *Practical phase-measuring circuit in which V_1 and V_2 are constant-gain amplifiers, V_3 a phase-shifter, V_4 a heptode mixer and V_5 a balancing valve*

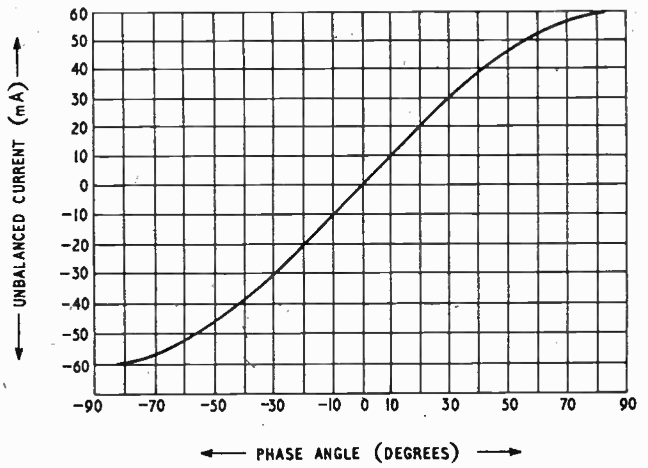


Fig. 3. *Calibration curve showing the variation in unbalanced current over a* ± 90° *range*

by noting the deflection of the meter with or without the network in circuit, as shown in Table 1.

It should be noted that for a phase angle $\theta = \pi/4$ between the pair of input signals, the magnitude of the angle indicated by the meter must be the same with or without the quadrature network included in the circuit. If not, it is the result of some unwanted phase shift in the circuit which should be avoided. The meter indication with the 90° phase shifter gives the correct reading since the effect of any undesired phase shift in the reference or in the signal channel prior to the mixer has been nullified by the preliminary zero setting and null adjustment of the meter.

Discussion

The ratio of the meter readings observed and the cosine of the corresponding phase angles between the pair of signals used during the calibration should always be constant. This has been checked, and the equivalent error angle calculated for the maximum departure from the mean value has been found to remain within ± 1% over the whole range.

It can be shown that if distortion (due to the transfer process) is introduced in the signal during the measurement of phase angle then the resulting output will consist of alternating components and will be integrated to zero over one cycle by the *RC* network in the output of the mixer. If the input frequency itself is distorted, then owing to the harmonic operation of the mixer stage the effect will be negligible so long as the distortion is small.

To obviate the need of meter calibration, a calibrated phase shifter could be incorporated in one channel and adjusted to a null to read the angle. Owing to the insensitivity of response at harmonic and unrelated frequencies, this arrangement produces a sharper null compared with the straight-line condition in an oscilloscope.[2,6]

An example of one application is the measurement of torque of a rotating shaft, in which an accurate determination of small phase angle[5] is required. Therefore the curve shown in Fig. 4 has been plotted to show how the mixer anode voltage varies over this range. For accuracy and consistency of results, a stabilized power supply of low-internal impedance, high-stability components in the quadrature network and a signal source of stable output (both in frequency and amplitude) should be used.

Acknowledgement

The author wishes to thank Dr. H. Rakshit, D.Sc. for his kind interest in the work and for some helpful discussions.

APPENDIX

Fig. 5 (b) is the equivalent circuit of the phase-shifter stage, shown in Fig. 5 (a), for frequencies at which the interelectrode capacitances have negligible effect. The loop equations are then given by

$$\mu (e_i - i_1 r_K) = i_2\left(R + \frac{1}{j\omega C}\right) + (i_1 + i_2)\, r_a \quad \ldots \quad (4)$$

$$i_1 (r_K + r_L) = i_2\left(R + \frac{1}{j\omega C}\right) \quad \ldots \quad (5)$$

Substituting i_1 from Equ. (5) in Equ. (4),

$$i_2 = \frac{\dfrac{\mu e_i (r_K + r_L)}{r_L + r_K (1 + \mu) + r_a}}{r_a \cdot \dfrac{r_L + r_K}{r_L + r_K (1 + \mu) + r_a} + \left(R + \dfrac{1}{j\omega C}\right)} \quad \ldots \quad (6)$$

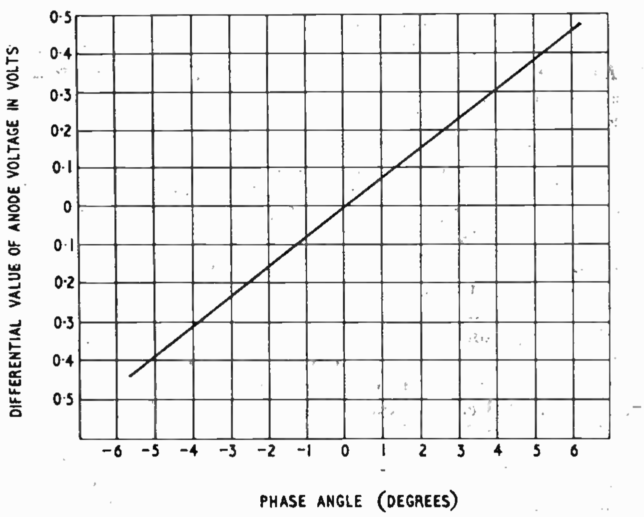
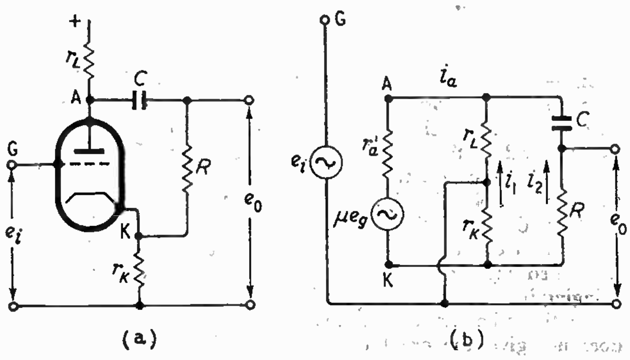


Fig. 4. *Showing how the mixer anode voltage varies over a* ± 6° *range*

Fig. 5. (*a*) *Phase-shifter stage;* (*b*) *equivalent circuit of* (*a*)



∴ The output voltage

$$e_0 = (V_R - V_{rK}) = i_2 \left[R - \frac{\left(R + \frac{1}{j\omega C}\right) r_K}{r_K + r_L} \right] \quad .. \quad .. \quad (7)$$

When $r_L = r_K = r$, then

$$e_0 = \tfrac{1}{2}\, g_m\, R_{eq}\, e_i \frac{R - \frac{1}{j\omega C}}{R_{eq} + R + \frac{1}{j\omega C}}$$

where $R_{eq} = \dfrac{2}{g_m + \frac{1}{r} + \frac{2}{r_a}}$

The phase of the output voltage, as seen from Equ. (8), varies as $\phi = 2 \tan^{-1} \dfrac{1}{\omega CR}$ and its magnitude remains constant so long as $R \gg R_{eq}$ and is equal to $e_i/2$ provided that $g_m R_{eq} = 1$.

REFERENCES

1. F. A. Benson and A. O. Carter, "Phase Angle Measurements", *Electronic Engng*, 1950, Vol. 22, p. 238.
2. J. R. Ragazzini and L. A. Zadeh, "A Wide Band Phase Meter", *Rev. Sci. Instrum.*, 1950, Vol. 21, p. 145.
3. L. H. O'Neil and J. L. West, "Direct Reading Phase Meter", *Rev. sci. Instrum.*, 1950, Vol. 21, p. 273.
4. G. N. Patchett, "A Versatile Phase-Angle Meter", *Electronic Engng*, 1952, Vol. 24, p. 224.
5. F. P. Moss, "An Audio Frequency Phase Meter", *Electronic Engng*, 1954, Vol. 24, p. 361.
6. J. F. Young, "Phase Measurement in Feedback Amplifier", *Electronic Engng*, 1955, Vol. 27, p. 311.
7. K. R. Sturley, "Radio Receiver Design", 2nd Edn, Pt 1, (Chapman & Hall Ltd., London, 1953).
8. P. Kundu, "R.C. Tuned Oscillators", *J. Brit. Instn Radio Engrs*, 1951, Vol. 11 p. 233.

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.

Currents on Strip Aerials

Sir,—In a paper of the above title (T. B. A. Senior, *Electronic & Radio Engineer*, February 1959, p. 60) a method of calculation of strip-aerial current distribution is suggested which raises hopes that a solution in closed form of Hallén's integral equation for a linear aerial may at last be forthcoming. Unfortunately, through an error in the physical reasoning which appears early in the paper, the current distribution found is not, as claimed, for the isolated thin strip aerial, and a comparison with the usual aerial formulae, though interesting, cannot indicate the form of distribution on thin linear aerials. Senior takes an infinite strip aerial and chops it cross-wise into thin rectangular segments and says that this process "cannot generate any transverse current nor effect the x-dependence of the longitudinal current". This statement is only true if the *totality of strips is retained.* The current distribution found is therefore that of an aerial in the presence of a large number of similar aerials, all fed equally and in-phase. These latter *cannot be removed without altering the distribution on the remaining strip.* Senior's implied (and unsupported) statement to the contrary is therefore invalid, as may also be seen by writing down the integral equations for the two cases. The two differ and cannot be transformed one into the other.

It is therefore to be expected that the comparison of the thin wire aerial and the "isolated" strip aerial will give different results, since the strip aerial is not, in fact, isolated; or alternatively, it is infinitely wide and would, therefore, be expected to differ appeciably.

Standard Telecommunication Laboratories Ltd., L. Lewin.
Enfield, Middlesex.
19th February 1959.

Sir,—Dr. Lewin's comments in the first part of his letter are, of course, quite correct. Though the 'chopping' of the infinite strip does not of itself perturb the longitudinal current distribution, the removal of the neighbouring portions does have some effect and, in consequence, the distribution is only an approximation to that for an isolated strip. Were this not so it would be a trivial matter to determine the current on any flat plate using, for example, the current distribution on an infinite sheet. Inasmuch as this is not stated explicitly in the paper, I am grateful to Dr. Lewin for bringing it to the readers' attention.

On the other hand, I cannot agree that because the solution does not give the exact current on an isolated strip it has no relevance to the problem. Bearing in mind that the method only purports to give the longitudinal variation of the longitudinal current (though the transverse variation can be estimated if the strip is narrow) and makes no statement about the transverse current, let us consider first a single strip of large but finite width. Along the central portion, at least, the longitudinal distribution will be very similar to that for an infinitely-wide strip and as such is characterized by an interaction between the two edges perpendicular to the current flow. When the length of the strip is only a fraction of a wavelength (ka not large compared with unity), the effect of this interaction dominates the distribution and, of particular interest in this case, is the fact that the variation $(x^2-a^2)^{\frac{1}{2}}$ demanded by the edge condition is the first term in the expansion in terms of the related functions $(x^2-a^2)^{r+\frac{1}{2}}$, $r = 0, 1, 2, \ldots\ldots$

As the width of the strip is decreased, the perturbation produced by the longitudinal edges will be felt even along the central portion of the strip, but an attempt to estimate the effect suggests it is of second order and that, by and large, the longitudinal distribution is still governed by the perpendicular edges. Unfortunately, it is almost impossible to give a precise statement of the approximation in the absence of an exact solution to the problem but experimental measurements do go some way towards justifying the approximation. A consequence of this is that the similar aerials referred to in Dr. Lewin's letter do not have an appreciable effect on the longitudinal current distribution.

In connection with the comparison between the distributions on thin wires and strips, I agree that the difference is not really surprising but would say that it is significant and is not merely a consequence of the approximation nature of my solution. Even without the analysis given in the paper an immediate distinction between the two can be indicated. Whereas the strip current must behave like $(x-a)^{\frac{1}{2}}$ as, for example, $x \to a$ in order to satisfy the edge condition, the thin-wire current vanishes as $(x-a)$ at this point.

Department of Electrical Engineering, T. B. A. Senior.
The University of Michigan,
Michigan, U.S.A.
10th March 1959.

Divide and Rule

Sir,—Computer's article "Divide and Rule" published in the February issue was read with interest. It does not appear to be very widely known, however, that a much faster way of evaluating this remainder is by means of a method due to Horner[1]. This method reduces this evaluation to a relatively simple multiply-and-add process, and is as follows:

The coefficients of the powers of x are detached and arranged in descending order preceded with their appropriate sign. It should be noted that any power of x that is missing should always be replaced by including a position for it and giving it a zero coefficient. Let us take the example given by Computer.

$$f(x) = x^6 - 3x^5 + 9x^4 - 27x^3 + 40x^2 - 60x + 100$$

The coefficients of x are detached as shown below,

$$1 \quad -3 \quad +9 \quad -27 \quad +40 \quad -60 \quad + \quad 100$$

If in the above example there was no term in, say, x^2, then the 40 would be replaced by 0.

The next step is to assign a value for x and multiply the first term on the left (coefficient of the highest power of x) by this assigned value and add the result to the succeeding coefficient, this result is once more multiplied by the assigned value and added to the succeeding term and so on until all terms are used. This is probably best demonstrated by example. We will give Computer's value for x (that is, 2·1) and evaluate the equation above.

$f(2{\cdot}1)$ 1	−3		+9	−27	+40	−60
	+2·1		−1·89	+14·931	−25·3449	+30·77571
	−0·9	×2·1	+7·11	−12·069	+14·6551	−29·22429

+100
−61·371009
38·628991

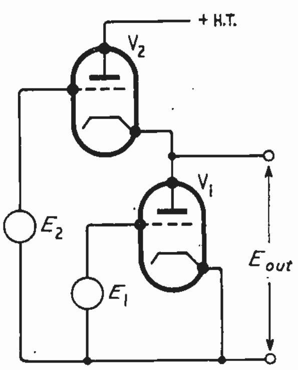
As can be seen, this is the same result achieved by Computer but done in one step. It is, of course, appreciated that the methods outlined by Computer are a means to an end. It is, however, as well to realize that, in general, it is more often required to determine where the roots of the equation lie (that is, when the remainder is zero) so that there is normally not very much justification for taking the evaluation of the remainder to too high a degree of accuracy. It is of interest to note that the method outlined in this letter can be extended to give roots of a polynomial, real roots that is, to a high degree of accuracy and a method was the subject of an article of mine published some years ago in *Machinery*[2].

[1] Prof. H. W. Turnbull, F.R.S., "Theory of Equations".
[2] "The Solution of Polynomial Equations", *Machinery*, 1st July 1955, Vol. 87.

Barkingside, R. A. LINDSEY.
Ilford, Essex.
5th March 1959.

New Anode-Follower

SIR,—The anode-follower circuit suggested by your correspondent J. Hajek in the February issue is an interesting idea. I have used an almost identical circuit, however, as a novel differential amplifier, as shown in the figure.



For two different valves, it is easy to show

$$E_{OUT} = \frac{\mu_2 R_{a1} E_2 - \mu_1 R_{a2} E_1}{R_{a1}(1 + \mu_2) + R_{a2}}$$

which for identical values reduces to

$$E_{OUT} = \frac{\mu}{\mu + 2}(E_2 - E_1)$$

or

$$E_{OUT} \approx (E_2 - E_1) \text{ if } \mu \gg 2$$

The circuit also has, of course, the low output impedance features of the anode-follower circuit.

Catford, EDWARD J. DYER.
London, S.E.6.
9th February 1959.

New Books

An Introduction to the Theory and Practice of Semiconductors

By A. A. SHEPHERD, M.Sc., Ph.D., A.Inst.P. Pp. 206. Constable & Co. Ltd., 10-12 Orange Street, London, W.C.2. Price 18s. 6d.

In this small volume the elements of the theory and technology of semiconductors are covered. Inevitably the compression of the large amount of material available has meant that while parts of the book are intelligible to the initiated, beginners may not find it so suitable. To describe the Fermi level, for example, as the maximum energy at zero temperature instead of pointing out its relation to the occupation probability of states does not make for a clear physical grasp.

However, within the compass of the book many topics are usefully described together with references to original papers. The band theory of solids and its relation to metals and semiconductors and conduction through them by majority and minority carriers are discussed. Purification of germanium and silicon and the preparation of single crystals of these materials are described. The theory and manufacture of rectifiers, photocells and transistors are dealt with and in the final chapter properties of intermetallic compounds are examined.

The book will be helpful to beginners in some of its descriptive parts but much of the matter is so compressed that it needs amplifying and qualifying from more advanced texts. M.S.

Nuclear Reactor Shielding

By J. R. HARRISON, M.A.(Oxon). Pp. 68 + viii. Price 10s. 6d.

Nuclear Reactor Control and Instrumentation

By J. H. BOWEN, B.Sc., A.M.I.E.E., and E. F. O. MASTERS, B.Sc., A.C.G.I., A.M.I.E.E. Pp. 78 + x. Price 12s. 6d.

Steam Cycles for Nuclear Power Plant

By W. R. WOOTTON. Pp. 66 + vii. Price 10s. 6d.

These three books are Nos. 4, 5 and 6 respectively of a series of nuclear engineering monographs, published by Temple Press Ltd., Bowling Green Lane, London, E.C.1, in association with their monthly journal, "Nuclear Engineering". Monographs Nos. 1, 2 and 3 were mentioned in the May 1958 *issue of "Electronic & Radio Engineer".*

Brans' Radio Tube Vade-Mecum 1958. 14th Edition

Compiled by P. H. BRANS. Pp. 464. (P. H. Brans Ltd., Antwerp, Belgium.) Bailey Bros. & Swinfen Ltd., Hyde House, West Central St., London, W.C.1. Price 32s.

Tube and Semiconductor Selection Guide 1958–59. 2nd Edition

Compiled by Th. J. KROES. Pp. 160. Philips' Technical Library. Cleaver-Hume Press Ltd., 31 Wright's Lane, London, W.8. Price 9s. 6d.

Radiotron High Fidelity

Pp. 43. Published by Amalgamated Wireless Valve Co. Pty. Ltd., 47 York Street, Sydney, Australia. Price 3s. (Australian).

Reprints of high-fidelity and tape-recorder articles which were published in the A.W.V. journal, *Radiotronics.*

Voice Across the Sea

By ARTHUR C. CLARKE. Pp. 220 + x. Frederick Muller Ltd., 110 Fleet Street, London, E.C.4. Price 18s.

The story of deep-sea cable laying from 1858–1958.

MEETINGS

I.E.E.

10*th April*. "Problems of Sight, Hearing and Touch", discussion to be opened by Professor E. C. Cherry, D.Sc.(Eng.) and "Human Engineering Recording Problems", discussion to be opened by H. C. W. Stockbridge, M.A. at 6 o'clock.

13*th April*. "The Function and Content of an Electric Theory Course", discussion to be opened by F. E. Rogers at 6 o'clock.

21*st April*. "The Problem of Maintenance of Electronic Equipment in the Process Industries", discussion.

23*rd April*. "The Geophysical Year 1957–58", the 50th Kelvin lecture by Sir David Brunt, Sc.D., F.R.S.

24*th April*. "Engineering Aspects of Commercial Television Programme Presentation", by T. C. Macnamara and B. Marsden.

27*th April*. "The Field Strengths Required for the Reception of Television in Bands I, III, IV and V", by G. F. Swann.

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

Brit. I.R.E.

6*th April*. Symposium on "Large Capacity Storage Devices", sessions commence at 3 and 6 o'clock.

22*nd April*. "The Application of Magnetic Resonance to Solid State Electronics", by D. J. E. Ingram, Ph.D., at 6.30.

28*th April*. "Electron Microscopy", by Professor G. Causey and R. S. Page at 6.30.

5*th May*. "An Experimental Diode Parametric Amplifier and its Properties", by I. M. Ross, C. P. Lea-Wilson, A. J. Monk and A. F. H. Thompson, at 6.30.

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

The Television Society

23*rd April*. "The Design of Experimental Tuners for Bands 4 and 5 Television Receivers", by K. H. Smith, to be held at 7 o'clock at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.

The Institute of Navigation

17*th April*. "The Dectra Trials", by Colonel C. Powell, at 5.15 at The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.

28*th*–30*th April*. Convention—"The Place of Automation in Navigational Methods at Sea and in the Air", organized jointly with the French Institute of Navigation and Ausschuss für Funkortung, to be held in Paris.

The British Computer Society

14*th April*. "The Sorting of Data—An Attempt to Measure the Severity of the Task", by Dr. D. A. Bell, at 2.30 at the Northampton College of Advanced Technology, St. John's Street, London, E.C.1.

16*th April*. "The Mechanical Translation of Languages", by Professor Lancelot Hogben, at 6.15 at the Northampton College of Advanced Technology, St. John's Street, London, E.C.1.

The Society of Instrnment Technology

16*th April*. Control Section Annual General Meeting at 6 o'clock, followed by "Self-Optimizing Control System for a Certain Class of Randomly Varying Inputs", by A. P. Roberts, B.Sc., at Manson House, Portland Place, London, W.1.

NEW ELECTRICAL INSTRUMENT TEST SERVICE

The British Scientific Instrument Research Association have announced that a test service for electrical instruments has been established at their Chislehurst Laboratories. The service is available to both members and non-members of the Association and will be operated under the supervision of the National Physical Laboratory.

Using N.P.L. certified equipment, the department will test and issue certificates for instruments up to 'precision grade' accuracy for a fee much lower than that required for N.P.L. certification.

At present, the department will undertake the calibration of d.c. meters between 1 mV and 500 V and between 2 μA and 2 A. At some future date, the latter range is to be extended up to 25 A, and the calibration of a.c. instruments, ranging from 1–500 V and 0·01–25 A will be undertaken.

Further information may be obtained from the Director of Research, The British Scientific Instrument Research Association, South Hill, Chislehurst, Kent.

CORRECTIONS

In the article "Saturable-Transformer Switches" in the March issue, the oscillograms of Fig. 8(a) and (b) were interchanged.

In the article "H.F. Exponential-Line Transformers" in the February issue, the sign of n in the exponent in the denominator of equation (8) should be plus, not minus. Also equation (9) should be $n = -2jTl$.

INTERNATIONAL PLASTICS EXHIBITION

The 1959 plastics exhibition, organized by *British Plastics*, will be held at Olympia, London, from 17th–27th June (excluding Sunday).

Further particulars may be obtained from the organizers at Dorset House, Stamford Street, London, S.E.1.

OBITUARIES

W. J. Picken, O.B.E., M.I.E.E. died on 24th February at the age of 72. He joined Marconi's Wireless Telegraph Co. Ltd. in 1913 and became Deputy Engineer-in-Chief in 1935 and, later, Engineer-in-Chief. In 1938–1940 he combined this position with that of General Manager.

In 1943 he was seconded to C.V.D. and in 1946 he retired from Marconi's to become Secretary to C.V.D. until he retired in 1953. He then joined the English Electric Valve Co. Ltd. as Technical Consultant.

Frederick James Camm died suddenly on 18th February at the age of 63. He was Editor of *Practical Wireless* and *Practical Television*.

STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for February* 1959

Date 1959 February	MSF 60 kc/s 1500 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^8
1	+ 2	NM.
2	+ 2	+ 4
3	+ 2	+ 4
4	+ 2	+ 4
5	+ 2	+ 4
6	+ 3	+ 4
7	N.M.	N.M.
8	N.M.	N.M.
9	N.M.	N.M.
10	+ 2	+ 1
11	+ 2	0
12	+ 2	0
13	+ 2	0
14	+ 2	N.M.
15	+ 2	N.M.
16	+ 2	0
17	+ 3	0
18	+ 3	0
19	+ 3	+ 1
20	+ 3	+ 1
21	N.M.	+ 1
22	N.M.	+ 1
23	+ 3	+ 1
24	+ 3	+ 1
25	+ 4	+ 1
26	+ 4	+ 2
27	+ 4	+ 2
28	+ 4	+ 2

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

New Products

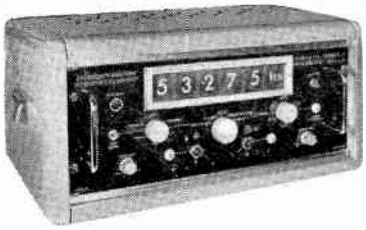
Frequency and Time-Measuring Instrument

This instrument, type TSA53, is completely transistorized and incorporates an optical in-line display by which time and frequency readings are presented in the form of brightly-illuminated 1-in. numerals, readable from a distance of 30 ft. The decimal point and units of measurement are also indicated, as shown in the photograph.

Details from the makers' advance specification for different types of measurement are given below:

Frequency Measurement	
Range	0·1 c/s–100 kc/s.
Input	0·5–500 V (r.m.s.).
Gate time	0·1, 1 and 10 sec.
Time-Interval Measurement	
Range	100 μsec – 278 hours (approx.).
Input	6–250 V (peak-positive pulse).
Period Measurement	
Range	0·00001 c/s–10 kc/s.
Input	0·5–500 V (r.m.s.).
Gate time	1 cycle of unknown frequency.
Count frequency	10 kc/s.
Random Pulse Counting	
Maximum rate	100,000 pulses per sec.
Input	0·5–500 V (r.m.s.).

Six standard output frequencies of 0·1, 1, 10, 100, 1,000 and 10,000 c/s are available and the instrument may be operated from either a 200–250-V, 50 c/s or a 12-V d.c. supply.



Venner Electronics Ltd.,
Kingston By-Pass, New Malden, Surrey.

Elapsed Time Indicator

The elapsed time indicator of Cass & Phillip Ltd. is now available as a fully-sealed and insulated unit for recording the running hours of any moving machinery or electronic equipment under the most adverse conditions of humidity and vibration.

Basically, the new version is designed with a double-skin instrument case, the mechanism being mounted internally on beryllium copper-leaf springs. A very rugged construction is achieved by jewel-mounting the balance staff, balance wheel and armature assembly.

The four leaf-springs are spaced 90° apart and carry the complete mechanism, which is inserted in a polythene case, and a specially-contoured rubber sealing (maintained in position by the bezel) completes the sealing.

Apart from being completely sealed against ingress of water and rust, the instrument is electrically insulated from the case.

Elapsed time can easily be read in hours and minutes to the nearest 15 seconds, and timing accuracy is better than 15 seconds in 24 hours.

The basic timing mechanism remains the

same, but improvements have been made in general constructions. For example, the drive between the centre shaft and the hour counters was previously by means of a 1 : 1 reduction gear, causing unnecessary friction and loss of energy. Now the drive is taken from the centre shaft to the hour counters and the latter are located just above the left centre of the instrument dial instead of at the base.

Cass & Phillip Ltd.,
Mark Road, Adeyfield, Hemel Hempstead, Herts.

Insulated Instrument Wire

Two types of p.t.f.e. insulated instrument wire with r.m.s. voltage ratings of 500 V and 1,000 V have been produced by Siemens Edison Swan Ltd.

Constructed from either single or stranded copper wires (each silver-plated to a radial thickness of not less than 30 × 10^{-6} in.), it is suitable for use where conductor temperatures of between −75 °C and 250 °C are experienced.

The wire, which is available in eleven different colours, can be easily stripped for connections.

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

Electric Counters

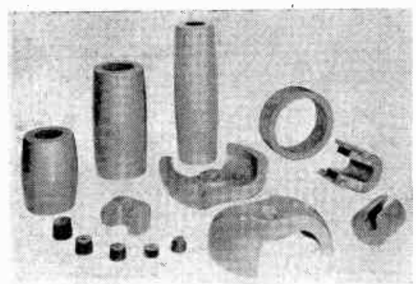
Counting Instruments Ltd., who recently extended their type 100 range of electric counters to include a dozens counter, have announced an improved method of manually re-setting these instruments by means of a push-button located on the left-hand side of the front panel. Previously, the re-setting operation was actuated by means of a centrally-placed lever (as described on p. 320 in the August 1958 issue of *Electronic & Radio Engineer*). The modification is said to make the counter completely dustproof.

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

Precision-Cast Permanent Magnets

A wide range of precision-cast permanent 'Magloy' magnets have been produced by Preformations Ltd.

The magnets, which are made from



precipitation-hardened ferromagnetic alloys, possess excellent magnetic resistance to shock and vibration at any frequency, are stable to within ±0·02% per °C temperature rise up to 500 °C, and are resistant to external magnetic fields.

Die-cast aluminium jackets are provided both for mounting and protection purposes as well as synthetic jackets for absolute protection against magnetic damage.

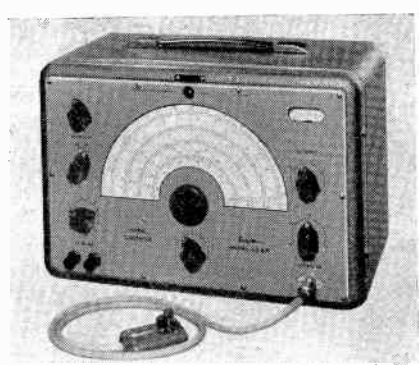
It is claimed that owing to the unique qualities of 'Magloy' materials, more efficient designs are practicable, especially in magnets for microwave applications.

Preformations Ltd.,
Cheney Manor, Swindon, Wilts.

Wide-Range Signal Generator

Taylor Electrical Instruments Ltd. has produced a signal generator, model 68 A/M, which covers the frequency range 100 kc/s to 240 Mc/s in eight bands (all on fundamentals).

Measuring 9 in. × 13 in. × 5 in. and weighing 18 lb., the instrument is specially



suitable for use in the servicing of radio and television receivers.

Details from the makers' specification include:

Calibration accuracy: $\pm$ 1%.
R.F. output: 100 mV (normal).
R.F. leakage: 3 μV (approx.).
Attenuation: Coarse control in 5 steps of -20 dB, and variable fine control to -20 dB (approx.).
A.F. output: Direct, and 1-V level (max.) provided.
Modulation: 400 c/s, 30%.
Output impedance: 75 Ω (approx.).
Scale length: 58 in.

A meter is incorporated for monitoring the r.f. output and a dummy aerial (complete with a coaxial lead and socket) is supplied with each instrument.

Taylor Electrical Instruments Ltd.,
Montrose Avenue, Slough, Bucks.

Phenolic-Dip Protected Capacitors

Johnson, Matthey & Co. Ltd. have announced that their range of silvered mica capacitors has been extended to include phenolic-dip protected capacitors which are suitable for use over the temperature range -60 °C to $+100$ °C. They are available in two main series, 'eyeleted' and 'fired'.

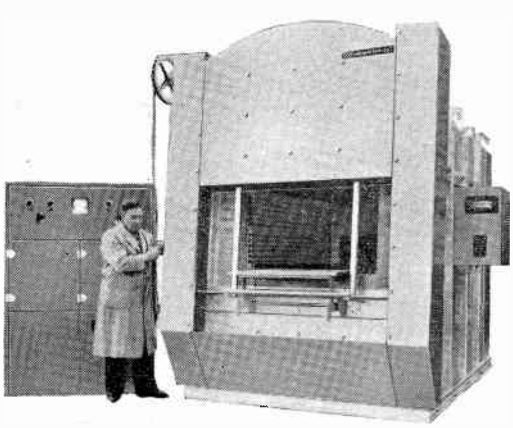
The 'eyeleted' types are assembled from pierced silvered-mica plates by means of composite silver-plated eyelets, the design being such that no inter-plate connecting foils are used. The 'fired' capacitors have plates which have been fired into a solid block, a construction that does not require mechanical joints or connecting foils.

Johnson, Matthey & Co. Ltd.,
73-83 *Hatton Garden, London, E.C.*1.

Electric Furnaces

A new series, E-HDF, of heavy-duty horizontal electric batch furnaces, suitable for general heat-treatment processes such as annealing, normalizing, stress relieving, vitreous enamelling, hardening and tempering, has been introduced by Barlow-Whitney Ltd.

Two basic models are available, the type E-HDF950 for operating temperatures to 950 °C (maximum 1,000 °C), and type E-HDF1100 for operating temperatures to 1,100 °C (maximum 1,150 °C). In the former type, the heavy-gauge nickel-chromium spiral heating elements are housed in specially-moulded carriers and operate at a normal mains voltage whereas, in the higher-temperature models, strip elements are employed which are operated at a much lower voltage.

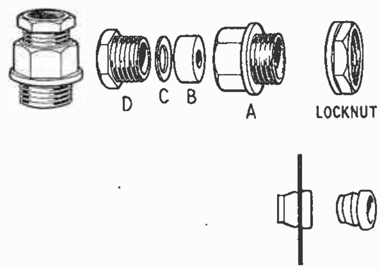


Both types are available with chamber sizes ranging from 18 in. (wide) × 9 in. (high) to 48 in. (wide) × 36 in. (high) with optional front-to-back depths of up to 96 in. Optimum efficiency and consistent performance is ensured by using carefully-graded insulation and a fully-automatic electronic pyrometric temperature controller.

Barlow-Whitney Ltd.,
2 *Dorset Square, London, N.W.*1.

Cable Glands and Insulating Bushes

The 'Elkay' weatherproof nylon cable glands are suitable for screwing into switch boxes, conduit boxes, chassis or wherever a weatherproof cable entry is required. They



can also be used as strain-relief bushings to prevent dragging or chafing.

Each gland comprises a body portion A, a rubber gland B, a metal washer C and a clamping bush D, and is available for cable diameters up to ½ in.

The plastic insulating bush (lower sketch) is designed for press-fitting into metal sheet from 24 s.w.g. up to 10 s.w.g. It is made of white polyethylene and is, at present, available in one size only, to accommodate cables up to $\frac{9}{32}$ in. diameter.

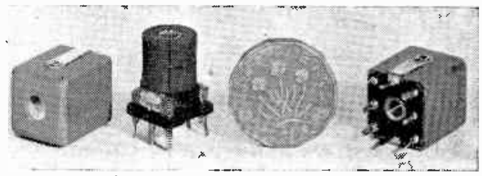
Elkay Electrical Manufacturing Co. Ltd.,
42 *Woburn Place, London, W.C.*1.

Miniature Ferrite Pot-Core Assembly

This assembly, which is enclosed by a half-inch square can, has been developed for transistor circuits especially where printed circuits are employed. The tag arrangement on the base complies with the B.S.I. recommendations for printed circuits, and the can-earthing clips perform the dual function of earthing and retaining of the whole assembly. A rubber compression pad maintains pressure against the mating surfaces of the pot cores and inductance adjustment is obtained by a 4-mm screw core. The components are all selected to operate under tropical conditions without deterioration of performance.

The cores can be supplied in several grades of ferrite according to inductance and frequency requirements. Q factors of 200 are readily obtainable at the lower frequencies and, at 20 Mc/s, a Q of 100 can be obtained.

The low-frequency assemblies can be



wound to give 11 mH inductance up to 4 Mc/s with correspondingly less inductance as the material permeability is reduced for higher-frequency performance.

The temperature coefficient of permeability for the assembly is of the order of less than 100 parts in 10^6 per °C.

Neosid Ltd.,
Stonehills House, Welwyn Garden City, Herts.

110° Television Tubes

Direct-viewing 21-in. and 17-in. television tubes employing electrostatic focusing and 110° magnetic deflection have been produced by Mullard Ltd.

The 21-in. type (AW53-88) is over 8½ in. shorter than its 70° counterpart and 5 in. shorter than a 90° tube. The 17-in. type (AW43-88) shows reductions of approximately 6½ in. and 3 in. respectively, over the equivalent size of the 70° and 90° tubes.

Both tubes, designed for operation up to 16 kV, have high-efficiency metal-backed phosphors, and provide excellent brightness and contrast under high ambient light levels.

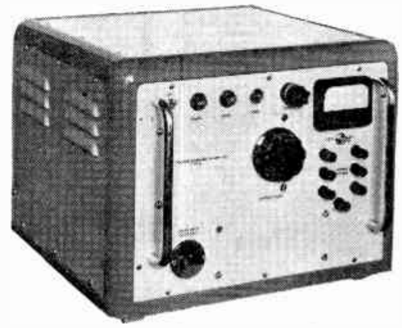
Mullard Ltd.,
*Torrington Place, London, W.C.*1.

Low-Voltage Stabilized Power Supply

A transistor-stabilized power supply, type 10, suitable for applications where a highly-stable d.c. supply is required, has been produced by the Electronics Division of G.E.C.

Measuring 17½ in. × 17¾ in. × 13 in. and weighing 85 lb, the equipment (which is mains-operated) provides a continuously-variable d.c. output of 6–20 V up to 10 A. The voltage regulation between no load and full load is approximately 1·4% and, at full load, the peak-to-peak ripple is less than 2 mV. It is claimed that after a full-load continuous run lasting five hours, there is less than 1% change in the output voltage while, for a 5% variation in the main supply, the output-voltage variation is approximately 1 : 150. Overload protection is provided.

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



Abstracts and References

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

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

ACOUSTICS AND AUDIO FREQUENCIES

534.1 **1039**

Variation of the Natural Frequencies of Membranes and Resonators with Additional Loads.—Yu. N. Dnestrovskiĭ. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 244–252.)

534.2 **1040**

The Sound Field Generated by a Point Source in a Layer Lying on a Half-Space.—Yu. L. Gazarian. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 233–238.) An investigation of the frequency dependence of the 'lateral-wave' field.

534.2 **1041**

The Effective Dynamic Parameters for Sound Propagation in Inhomogeneous Media.—L. M. Khaĭkovich & L. A. Khalfin. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 275–281.) Equations are derived for calculating the 'effective' parameters of a two-component medium in which the inhomogeneities are distributed in the form of a regular lattice.

534.2 **1042**

Scattering of Sound Waves in Irregular Waveguides.—A. D. Lapin. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 267–274.) A method is proposed for calculating the scattering in waveguides for the case in which the scattered field is not small in comparison with the incident field. The case of small fluctuations in the parameters of the medium inside the waveguide and the case of wall roughness in the waveguide are examined. The scattered field is found in the form of superpositions of the normal waves which would occur in the waveguide if the irregularities were absent.

534.2-14 : 534.6 **1043**

Reverberation Tank Method for the Study of Sound Absorption in the Sea.—V. P. Glotov. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 239–243.) The tank has a capacity of $0 \cdot 5$ m^3 and was designed for ocean measurements at depths not exceeding 50 m. Results of measurements at 20 m depth and frequencies 15–80 kc/s are given.

534.23 **1044**

On the Acoustic Field [caused] **by a Vibrating Source Arbitrarily Distributed on a Ribbon Plate: Part 1.**—N. Kawai. (*J. phys. Soc. Japan*, Nov. 1958, Vol. 13, No. 11, pp. 1374–1384.) Two-dimensional rigorous solutions of the field in air are given for sources of infinitely thin and infinitely long ribbons.

534.232 **1045**

Calculations for a Piston-Type Piezoelectric Radiator when Internal Losses are Neglected.—A. A. Anan'eva. (*Akust. Zh.*, July/Sept. 1958, Vol. 4, No. 3, pp. 223–232.) Analysis of flat piston-type quartz and $BaTiO_3$ transducers operating in water. The importance of choosing materials with a high piezoelectric factor, a high density and large modulus of elasticity is stressed.

534.241 **1046**

The Sound Echo Reflected from a Sphere under Pulse Conditions.—M. Federici. (*Ricerca sci.*, Aug. 1958, Vol. 28, No. 8, pp. 1659–1667.)

534.26-8-14 **1047**

Contribution to the Study of the Scattering of an Ultrasonic Plane Wave by Surfaces with a Periodic Structure.—H. Blons. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 50–52.) Results of experiments made in water at 625–2 500 kc/s on reflections from a periodic surface show good agreement with diffraction theory.

534.26-8-14 : 548.73 **1048**

The Diffraction of Ultrasonic Waves by Multilayer Arrays of Rods.—P. de Villers. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 52–54.) Experiments were made in water at a wavelength of $0 \cdot 57$ mm and with rod diameter $0 \cdot 3$ mm, in order to confirm Bragg's laws for the diffraction of X rays by crystal lattices.

534.522.1 **1049**

Diffraction of Light by Ultrasonic Waves—Oblique Incidences.—S. Parthasarathy & C. B. Tipnis. (*Nature, Lond.*, 18th Oct. 1958, Vol. 182, No. 4642, pp. 1083–1084.) Experimental confirmation of a closed-form expression for diffraction intensity.

534.522.1 **1050**

Investigation of Progressive Ultrasonic Waves by Light Refraction.—M. A. Breazeale & E. A. Hiedemann. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 751–756.) Adaptations of the method used by Loeber & Hiedemann (1941 of 1956) for the study of stationary waves are described, for the detection of waveform distortion due to finite-amplitude effects in liquids.

534.522.1 **1051**

Study of the Intensity Distribution of the Light Diffracted by Ultrasonic Waves.—R. B. Miller & E. A. Hiedemann. (*J. acoust. Soc. Amer.*, Nov. 1958, Vol. 30, No. 11, pp. 1042–1046.) Differences between experimental results and theory are discussed.

534.7 **1052**

Signal Detection as a Function of Frequency Ensemble: Part 1.—F. A. Vaniar. (*J. acoust. Soc. Amer.*, Nov. 1958, Vol. 30, No. 11, pp. 1020–1024.) Results of an experimental investigation of the detection of multiple-tone signals in noise are compared with predictions based on mathematical models.

534.75 **1053**

Intensity and Duration of Noise Exposure and Temporary Threshold Shifts.—W. Spieth & W. J. Trittipoe. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 710–713.) See also 682 of March.

534.75 **1054**

Residual Effects of Low Noise Levels on the Temporary Threshold Shift.—W. J. Trittipoe. (*J. acoust. Soc. Amer.*, Nov. 1958, Vol. 30, No. 11, pp. 1017–1019.)

534.76 **1055**

On the Mechanism of Binaural usion.—E. E. David, Jr, N. Guttman & W. A. van Bergeijk. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 801–802.) The relation between interaural intensity difference and interaural time difference is obtained by a pulse test method. The interpretation of test results is discussed.

534.76 **1056**

An Artificial Stereophonic Effect Obtained from a Single Audio Signal.—M. R. Schroeder. (*J. audio Engng Soc.*, April 1958, Vol. 6, No. 2, pp. 74–79. Discussion.) Report of experimental investigations of the stereophonic effect achieved by feeding a signal to both ears of a listener once directly in phase, and a second time, with a delay of 50–150 msec, in antiphase.

534.78 : 621.39 **1057**

Phonetic Pattern Recognition Vocoder for Narrow-Band Speech Transmission.—H. Dudley. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 733–739.) A phonetic-pattern recognition circuit (1061 below) is combined with a synthesizer to form a phonetic vocoder.

534.781 **1058**

Modern Instruments and Methods for Acoustic Studies of Speech.—G. Fant. (*Acta polyt. scand.*, 1958, No. 246, Ph 1, 81 pp.)

534.782 **1059**

Segmentation Techniques in Speech Synthesis.—G. E. Peterson, W. S. Wang & E. Sivertsen. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 739–742.) In the method described discrete segments of recorded utterances are joined together to produce continuous speech. Techniques for obtaining the segments are discussed.

534.782 **1060**

Segment Inventory for Speech Synthesis.—W. S. Wang & G. E. Peterson. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 743–746.) See 1059 above.

534.784 : 621.395 **1061**

Automatic Recognition of Phonetic Patterns in Speech.—H. Dudley & S. Balashek. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 721–732.) The performance of a ten-word digit recognizer for voice dialling of telephone numbers is discussed. A model of a phonetic pattern recognizer based on a detailed analysis of the power frequency spectrum is described; this provides almost perfect digit recognition when set for a particular voice.

534.79 **1062**

Advantages of the Discriminability Criterion for a Loudness Scale.—W. R. Garner. (*J. acoust. Soc. Amer.*, Nov. 1958, Vol. 30, No. 11, pp. 1005–1012.)

534.846 **1063**

Acoustic Criteria of Outstanding Old and New Concert Halls.—F. Winckel. (*Frequenz*, Feb. 1958, Vol. 12, No. 2, pp. 50–59.) Comparison and discussion of architectural and acoustic features of pre- and post-war auditoria and concert halls.

534.85/.86 : 534.76 **1064**

Stereophonic Sound with Two Tracks, Three Channels by means of a Phantom Circuit (2PH3).—P. W. Klipsch. (*J. audio Engng Soc.*, April 1958, Vol. 6, No. 2, pp. 118–123.) Practical details are given for combining the signals of each channel and using the resultant signal to drive a third central loudspeaker.

534.85/.86 : 534.76 **1065**

Stereo-reverberation.—R. Vermeulen. (*J. audio Engng Soc.*, April 1958, Vol. 6, No. 2, pp. 124–130.) The effect can be achieved using a single-channel system feeding several loudspeakers, distributed around an auditorium, with delayed signals repeated many times at decreasing levels.

534.88 **1066**

Correlation with Similar Uniform Collinear Arrays.—M. J. Jacobson. (*J. acoust. Soc. Amer.*, Nov. 1958, Vol. 30, No. 11, pp. 1030–1034.) An acoustic multiple-receiver correlation system is considered. See also 1959 of 1958.

AERIALS AND TRANSMISSION LINES

621.372.2 **1067**

Approximate Models for Transmission Lines and their Errors.—R. R. Vierhout. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 94–95.)

621.372.2 : 538.569.2/.3 **1068**

The Absorption of Electromagnetic Waves in Absorbers and Lines with Sectionally Homogeneous Distributed Conductance.—K. L. Lenz & O. Zinke. (*Z. angew. Phys.*, Oct. 1957, Vol. 9, No. 10, pp. 481–489.)

621.372.2 : 621.317.34 **1069**

Novel Method for Measuring Impedances on Surface-Wave Transmission Lines.—G. Schulten. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 76–77.)

621.372.22 : 621.372.51 **1070**

H.F. Exponential-Line Transformers.—S. G. Young. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 40–44.) The design and construction of four-wire tapered transmission lines to match balanced impedances are described.

621.372.8 : 621.3.018.75 **1071**

Pulse Waveform Degradation due to Dispersion in Waveguide.—R. S. Elliott. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 254–257.)

621.372.821 **1072**

Optimum Impedance and Dimensions for Strip Transmission Line.—K. S. Packard. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 244–247.)

621.372.824 : 621.372.831 **1073**

Low-Reflection Discontinuities in Diameter of Coaxial Lines.—G. W. Epprecht. (*Tech. Mitt. PTT*, 1st March 1958, Vol. 36, No. 3, pp. 97–103.) Test results obtained on 50-Ω coaxial-waveguide junctions with changes in diameter ratio from 1·15 : 1 to 3 : 1 are given as a function of the distance between the transitions in the inner and outer conductors.

621.372.825 **1074**

Serrated Waveguide.—(*Trans. Inst. Radio Engrs*, July 1957, Vol. AP-5, No. 3, pp. 270–283. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1759.)

Part 1—Theory.—R. S. Elliott (pp. 270–275).

Part 2—Experiment.—K. C. Kelly & R. S. Elliott (pp. 276–283).

621.372.825 **1075**

Wave Propagation in a Diaphragm-Type and a Corrugated Waveguide.—G. Piefke. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 26–34.) The influence of diaphragms and periodic corrugations in a circular loss-free waveguide is investigated theoretically. See also 2625 of 1958.

621.372.826 : 538.614 **1076**

Surface Waves on the Boundary of a Gyrotropic Medium.—M. A. Gintsburg. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1635–1637.) Brief mathematical analysis for the case of (*a*) a gyrotropic plate between two isotropic media and (*b*) a channel between two gyrotropic media, with reference to the wave-retardation properties of the system. See 2564 and 2927 of 1954.

621.372.83 **1077**
Frequency Compensation for Simple Stepped Waveguide Transforming Sections.—D. Wray. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 76–79.) Sections can be 'broad-banded' by varying both broad and narrow waveguide dimensions at each step; this gives a better performance than sections designed on a single mid-band frequency.

621.372.83 **1078**
A Variable-Ratio Microwave Power Divider and Multiplexer.—W. L. Teeter & K. R. Bushore. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 227–229.)

621.372.832.8 **1079**
High-Power Ferrite Circulators.—P. A. Rizzi. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 230–237.)

621.372.837 **1080**
Waveguide Switches and Branching Networks.—J. W. Sutherland. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 64–68.) Barrel, shutter, gas-discharge and ferrite switches are compared. Branching tee properties and applications are discussed and a multiple branching network using hybrids is described.

621.372.852.21 **1081**
Broad-Band Quarter-Wave Plates.—W. P. Ayres. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 258–261.) Analytical expressions are given for the propagation constant for the two orthogonal dominant modes in a square waveguide loaded with a centred dielectric slab.

621.372.855 **1082**
Waveguide Termination with Commercial Film-Type Resistors.—U. v. Kienlin & A. Kürzl. (*Nachrichtentech. Z.*, March 1958, Vol. 11, No. 3, pp. 138–141.) The 20-W termination described consists of an arrangement of 20 1-kΩ, 1-W resistors parallel to the electric field. Methods of compensation and advantages of the system are discussed.

621.372.86 **1083**
End Sections for Matching Waveguide Directional Couplers of Periodic Structure.—I. Lucas. (*Arch. elekt. Übertragung*, Feb. 1958, Vol. 12, No. 2, pp. 91–96.) Rules for determining the matching elements required at either end of the coupler are given. The bandwidth for a given matching condition is calculated for one and two end sections. Measurements on a directional coupler with power evenly shared between outlets indicate that one matching element at either end would be adequate.

621.396.67 + 621.396.11 **1084**
International Colloquium on Current Problems in Radio Wave Propagation, Paris, 17th–21st Sept. 1956.—(See 1333.)

621.396.67 **1085**
The Current Distribution and Input Impedance of Cylindrical Antennas.—E. V. Bohn. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. AP-5, No. 4, pp. 343–348.)

621.396.67-418 **1086**
Currents on Strip Aerials.—T. B. A. Senior. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 60–63.) An exact expression is obtained for the longitudinal distribution of current excited on a perfectly conducting strip by a normally incident plane wave. Computations are made for quarter- and half-wave aerials. Deduced total current distributions do not agree with those for a thin wire, but give a better approximation than the cosine variation.

621.396.67.012 **1087**
Evaluating Aerial Performance.—L. A. Moxon. (*Wireless World*, Feb. & March 1959, Vol. 65, Nos. 2 & 3, pp. 59–65 & 139–144.) Various aspects of dipole, beam array and long-wire aerial design are discussed nonmathematically, using a simplified method for calculating gain and radiation resistance. Problems associated with transmission-line losses and with the maintenance of a high signal/noise ratio at the receiving aerial are considered.

621.396.67.029.62/.63 : 621.397.7 **1088**
Post-Installation Performance Tests of U.H.F. Television Broadcasting Antennas.—D. W. Peterson. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 656–672.) The precautions necessary to reduce errors introduced by reflections from the earth's surface, obstacles, clutter and the earth's curvature, are discussed.

621.396.674.3 **1089**
The Conductance of Dipoles of Finite Length and Thickness.—K. Fränz, P. A. Mann & J. Vocolides. (*Arch. elekt. Übertragung*, Feb. 1958, Vol. 12, No. 2, pp. 49–53.) For a given frequency a family of dipoles is obtained which have the same conductance. The shape and conductance of a dipole are determined by evaluating a number of integrals and solving a first-order differential equation.

621.396.674.3.01 **1090**
Formulation of the Antenna Circuit Theory based on the Variational Method.—K. Furutsu. (*J. Radio Res. Labs, Japan*, Oct. 1958, Vol. 5, No. 22, pp. 315–333.) Radiation power and field strength are expressed as functions of current distribution in a dipole aerial system. These are stationary with respect to small variations in the distribution and are independent of the magnitude of the distribution.

621.396.677 : 629.19 **1091**
Limitations of Satellite Antennas using Spherical Arrays.—K. S. Kelleher & J. P. Shelton. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 74–75.)

621.396.677.3 : 621-526 **1092**
Servo Phase Control shapes Antenna Pattern.—E. W. Markow. (*Electronics*, 2nd Jan. 1959, Vol. 32, No. 1, pp. 50–52.) A design of phase-stabilized u.h.f. amplifier and its application to the directivity control of a multielement stationary array is described. The angular phase error between input and output is less than 2°.

621.396.677.71 **1093**
Convergent Representations for the Radiation Fields from Slots in Large Circular Cylinders.—L. L. Bailin & R. J. Spellmire. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. AP-5, No. 4, pp. 374–383.)

621.396.677.71 : 621.397.61 **1094**
The Travelling-Wave V.H.F. Television Transmitting Antenna.—M. S. Siukola. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 49–58. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 383.)

621.396.677.75 **1095**
A Technique for Controlling the Radiation from Dielectric Rod Waveguides.—J. W. Duncan & R. H. DuHamel. (*Trans. Inst. Radio Engrs*, July 1957, Vol. AP-5, No. 3, pp. 284–289. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1759.)

621.396.677.8.012.12 **1096**
On the Simulation of Fraunhofer Radiation Patterns in the Fresnel Region.—D. K. Cheng. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. AP-5, No. 4, pp. 399–402.)

621.396.677.83 **1097**
The Ideal Plane Reflector.—M. Kummer. (*NachrTech.*, Feb. 1958, Vol. 8, No. 2, pp. 61–65.) The changes of the radiation pattern as a function of reflector distance are calculated for an infinitely large plane with a $\lambda/2$ dipole. The limitations in practical applications are considered.

621.396.677.832 **1098**
A Circularly Polarized Corner-Reflector Antenna.—O. M. Woodward, Jr. (*Trans. Inst. Radio Engrs*, July 1957, Vol. AP-5, No. 3, pp. 290–297. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1759.)

621.396.677.832 **1099**
Corner-Reflector Antennas with Arbitrary Dipole Orientation and Apex Angle.—R. W. Klopfenstein. (*Trans. Inst. Radio Engrs*, July 1957, Vol. AP-5, No. 3, pp. 297–305. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, pp. 1759–1760.)

621.396.677.85 **1100**
Lens-Aerial Design.—P. Foldes & L. Solymar. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 73–75.) A simple geometrical method is described for the design of a lens which will give a prescribed amplitude and phase distribution in an aperture when the aerial is fed by a given primary source. The graphical method yields a polygon as the approximation to the lens contour.

621.396.677.85 **1101**
Scanning-Lens Design for Minimum Mean-Square Phase Error.—E. K. Proctor & M. H. Rees. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. AP-5, No. 4, pp. 348–358.)

AUTOMATIC COMPUTERS

681.142 1102
An Electronic Ratio Calculator.—A. D. Boronkay. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 32–34.) Use is made of the principle that the amplitude ratio of two signals is approximately linearly related to the phase angle of the sum.

681.142 : 621.318.5 : 537.312.62 1103
Digital-Analogue Conversion with Cryotrons.—L. K. Wanlass & L. O. Hill. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 100–101.)

681.142 : 621.318.57 : 538.221 1104
A High-Speed Logic System using Magnetic Elements and Connecting Wire Only.—H. D. Crane. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 63–73.) Square-loop magnetic elements having multi-aperture or multipath variations of the toroid shape are used to provide unilateral information-flow properties. Information-bit rates above ¼ Mc/s have been obtained with commercial ferrite material. Advantages include simplicity of wiring and the inherent nondestructive read-out properties of the system.

681.142 : 621.383 1105
Automatic Data Reduction of Spot Diagram Information.—W. E. Goetz & N. J. Woodland. (*J. opt. Soc. Amer.*, Dec. 1958, vol. 48, No. 12, pp. 965–968.) The redistribution of energy falling on a photocathode during scanning by a mechanical image dissector is evaluated.

CIRCUITS AND CIRCUIT ELEMENTS

621.316.86 : 621.315.56 1106
Measurements of Nonlinearity in Cracked-Carbon Resistors.—G. H. Millard. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 31–34.) Frequencies used were 1·06 kc/s, 3·3 kc/s, and 90 Mc/s. Resistors measured ranged from 50–100 Ω, ⅛ W–55 W. The degree of nonlinearity would be negligible for most purposes.

621.318.57 : 621.318.134 1107
Millimicrosecond Microwave Ferrite Modulator.—A. H. Solomon & F. Sterzer. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 98–100.) Description of a switching device similar to the resonance-type isolator described by Enander (17 of 1957), but with ferrite rings placed inside the helix and with a single straight wire along the axis carrying the modulating current.

621.318.57 : 621.383 1108
Photoelectronic Circuit Applications.—S. K. Ghandhi. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 4–11.) The characteristics of switching circuits using combinations of electroluminescent cells and photoconductors are discussed. Applications are described in computer operations where the highest speed is not essential.

621.318.57 : 621.383.4 : 621.314.7 1109
A Temperature-Stabilized Phototransistor Relay Circuit.—J. C. Anderson & T. Winer. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 36–37.) A phototransistor and an ordinary junction transistor are used in a 'long-tailed pair' arrangement to reduce the variation in dark current.

621.318.57 : 621.387 1110
A Reversible Dekatron Counter.—D. L. A. Barber. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 42–43.)

621.319.45 1111
Recent Advances in the Solid-State Electrolytic Capacitor.—A. V. Fraioli. (*Trans. Inst. Radio Engrs*, June 1958, Vol. CP-5, No. 2, pp. 72–75. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1958, Vol. 46, No. 8, p. 1553.)

621.372.413 : 621.372.54.029.6 1112
Design of Aperture-Coupled Filters.—F. Shnurer. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 238–243.)

621.372.5 1113
Network Synthesis.—J. T. Allanson. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 66–69.) A method is outlined for the synthesis of balanced asymmetrical *RC* networks, which may have a lower pad loss than the corresponding symmetrical types.

621.372.5.029.6 : 621.317.341 1114
Definition of Lossy Quadripoles by Voltage Node Displacements in the Microwave Range.—F. Gemmel. (*Arch. elekt. Übertragung*, Feb. 1958, Vol. 12, No. 2, pp. 76–80.) Equations are derived giving the voltage s.w.r. and the position of the voltage node on the input line of a lossy quadripole in terms of the position of the short-circuit on its output line.

621.372.51(083.57) : 621.314.7 1115
Determination of Transistor Performance Characteristics at V.H.F.—G. E. Theriault & H. M. Wasson. (*Trans. Inst. Radio Engrs*, June 1957, Vol. BTR-3, No. 1, pp. 40–48.) An 'immittance chart', a development of the Smith Chart, is described, and its application in the design of adjustable matching networks for transistor measurements is illustrated.

621.372.54 1116
General Physical Relations in Ladder-Type Filters.—T. Laurent. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 1–8.) The improvement of filter characteristics by mismatch conditions between the ladder sections is considered and a design method is outlined. See also 1931 of 1953.

621.372.54 1117
Minimum-Insertion-Loss Filters.—E. G. Fubini & E. A. Guillemin. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 37–41.) The generalized criterion for Butterworth and Tchebycheff filters of arbitrary bandwidth is chosen for minimum loss in the centre of the band.

621.372.54 1118
The Design of Two-Section Symmetrical Zobel Filters for Tchebycheff Insertion Loss.—W. N. Tuttle. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 29–36.)

621.372.54 1119
A Voltage-Controlled Continuously Variable Low-Pass Filter.—A. J. Seyler & A. Korpel. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 16–22.) The filter has a 6 dB/octave cut-off characteristic and the cut-off frequency, at −3 dB, is continuously variable from 4·5 kc/s to 5 Mc/s by a control-voltage variation of 85 V.

621.372.54 1120
The Exact Design of Two Types of Single-Crystal, Wide-Band Crystal Filters.—T. R. O'Meara. (*Trans. Inst. Radio Engrs*, March 1958, Vol. CP-5, No. 1, pp. 46–52. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 931.)

621.372.54 1121
The Symmetrical Transfer Characteristics of the Narrow-Bandwidth Four-Crystal Lattice Filter.—T. R. O'Meara. (*Trans. Inst. Radio Engrs*, June 1958, Vol. CP-5, No. 2, pp. 84–89. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1958, Vol. 46, No. 8, p. 1553.)

621.372.543.3 1122
The Bifilar-T Circuit.—T. Roddam. (*Wireless World*, Feb. & March 1959, Vol. 65, Nos. 2 & 3, pp. 66–71 & 145–148.) The circuit is critically examined, using lattice network theory. See also 2658 of 1958 (Hendry & McIntosh).

621.372.56.029.64 : 621.317.7.089.6 1123
The Calibration of Microwave Attenuators by an Absolute Method.—E. Laverick. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 250–254.) A bridge method is described and results of measurements at 3·2 cm λ are given.

621.373.4 1124
Alternatives to the Wien Bridge.—J. F. Young. (*Wireless World*, Feb. 1959, Vol. 65, No. 2, pp. 92–95.) Oscillator circuits with two-gang control, having higher output voltages and greater selectivity than the basic Wien bridge circuit are described.

621.373.421 1125
A Wide-Band Voltage-Controlled Swept-Frequency *RC* Oscillator.—R. S. Sidorowicz. (*A.T.E. J.*, April 1958, Vol. 14, No. 2, pp. 88–112.) Oscillators based on the principle of parallel network tuning [see e.g. 2239 of 1955 (Stewart)] are examined. New equipment using this principle is described in detail. Frequency variations can be restricted to any limits in

the range 20 c/s–3 kc/s, and the maximum duration of the sweep cycle can be varied from 4 to 20 min.

621.373.421.11 **1126**

Incremental Frequency Control of *RC* Oscillators.—D. L. A. Smith. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 38–39.) An auxiliary phase-shifting network is introduced to provide a continuously variable incremental frequency control. A practical method is described for systematically correcting the error which exists for large increments.

621.373.421.13 : 529.786 **1127**

Construction and Properties of High-Quality Oscillator Crystals.—G. Becker. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 15–25.) Details are given of the shape, processing and mounting of crystals for use in quartz-clock standards. The influence of various design parameters on crystal characteristics and performance is discussed and a method of frequency control is described. See also 3468 of 1956 (Scheibe et al.).

621.373.43 : 621.314.7 **1128**

A Monostable Circuit using a Transistor.—H. Stinton. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 80–81.) A pulse generating circuit is described using a simple pulse transformer and a transistor. Pulse duration is adjustable over a wide range, greatly exceeding that usually associated with the pulse transformer used.

621.373.52 : 621.372.412 **1129**

Crystal-Controlled Transistor Oscillators.—H. Awender & A. Ludloff. (*Elektronische Rundschau*, March 1958, Vol. 12, No. 3, pp. 75–80.) The design of parallel-resonant circuits is considered.

621.374.3 : 621.387 **1130**

The Use of Dekatrons for Pulse Distribution.—G. H. Stearman. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 69–71.) Specific methods are discussed for increasing the number of output lines available from cold-cathode selector tubes used for distributing pulses in a fixed time sequence.

621.374.32 : 621.314.7 **1131**

A 1-Mc/s Transistor Decade Counter.—C. G. Bradshaw. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 96–97.)

621.374.4 **1132**

Investigations on Harmonic Frequency Dividers to Determine the Prerequisite Conditions for Achieving High Division Ratios and Large Locking Ranges at High Frequencies.—K. Schlichting. (*Z. angew. Phys.*, Sept. 1957, Vol. 9, No. 9, pp. 458–464.)

621.374.43 : 621.314.7 **1133**

A Regenerative Modulator Frequency Divider using Transistors.—F. Butler. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 72–75.)

621.374.5 : 538.652 : 621.396.96 **1134**

Magnetostrictive Delay Line for Video Signals.—G. I. Cohn, L. C. Peach, M. Epstein, H. O. Sorensen & D. P. Kanellakos. (*Trans. Inst. Radio Engrs*, March 1958, Vol. CP-5, No. 1, pp. 53–59. Abstract, *Proc. Inst. Radio Engrs*, May 1958, Vol. 46, No. 5, Part 1, p. 931.)

621.375.018.75 **1135**

Pulse Amplifier with Submillimicrosecond Rise Time.—F. Sterzer. (*Rev. sci. Instrum.*, Dec. 1958, Vol. 29, No. 12, pp. 1133–1135.) The rise time is about $0 \cdot 7$ mμs. The d.c. pulses modulate a 2 750-Mc/s carrier, and the r.f. pulses are amplified by a travelling-wave valve. Demodulation then gives the amplified d.c. pulses. The maximum power gain is 20 dB, and the maximum output voltage $2 \cdot 1$ V.

621.375.018.756 : 537.525.6 **1136**

A Low-Noise Wide-Band Amplifier for the Investigation of Electron Avalanches.—K. J. Schmidt-Tiedemann. (*Z. angew. Phys.*, Sept. 1957, Vol. 9, No. 9, pp. 454–458.)

621.375.024 : 621.317.321.027.21 **1137**

Low-Level D.C. Amplifier with Whole-Loop Feedback.—P. C. Hoell. (*Rev. sci. Instrum.*, Dec. 1958, Vol. 29, No. 12, pp. 1120–1124.) A 60-c/s vibrating-contact modulator and whole-loop negative feedback are used, and a stabilized gain of 10^7 from d.c. to 10 c/s, is obtained. The zero drift is less than $\pm 10^{-9}$ V per hour and less than 2×10^{-9} V per day.

621.375.2.012.6 **1138**

Amplitude and Phase Response of an Amplifier Operating on an Interrupted Cycle.—A. Sabbatini. (*Note Recensioni Notiz.*, May/June 1958, Vol. 7, No. 3, pp. 277–287.) Analysis of class-C amplifier characteristics.

621.375.4 **1139**

Transistor Amplifiers : Common Base versus Common Emitter.—R. F. Purton. (*A.T.E. J.*, April 1958, Vol. 14, No. 2, pp. 157–163.) The gain stability and frequency response of the two circuits are compared theoretically and experimentally. If a generator with a relatively low impedance is used, these characteristics need be no worse for the common-emitter than for the common-base amplifier.

621.375.4 **1140**

Transister Bias Design from Thermal Incremental Properties.—L. M. Vallese. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 88–93.) Explicit formulae for design of single-stage bias networks are given. Two thermal parameters need to be known, one related to dI_{co}/dT and the other to the rate of change of input current. Applications and experimental verification of the theory are included. See also 2078 of 1957.

621.375.4.01 : 512.831 **1141**

An Introduction to Matrices and their Use in Transistor Circuit Analysis.—J. S. Bell & K. Brewster. (*Electronic Engng*, Feb. 1959, Vol. 31, No. 372, pp. 98–102.) Brief details are given for analysing both simple and complex circuits using transistors, with emphasis on application to practical circuits.

621.375.432 **1142**

Low-Noise Transistor Amplifier Stages with Negative Feedback.—E. R. Hauri. (*Tech. Mitt. PTT*, 1st March 1958, Vol. 36, No. 3, pp. 103–108.) Theoretical investigation of the grounded-emitter connection with feedback resistance in the emitter circuit. The relation of source impedance to optimum noise figure is determined.

621.375.9 : 538.569.4.029.64 **1143**

Maser Action in the Region of 60°K.—C. R. Ditchfield & P. A. Forrester. (*Phys. Rev. Lett.*, 15th Dec. 1958, Vol. 1, No. 12, pp. 448–449.) A three-level solid-state maser using a ruby with a $0 \cdot 1\%$ Cr concentration has given a stable gain of 30 dB at 9 518 Mc/s.

621.375.9 : 538.569.4.029.64 **1144**

A Tunable Maser Amplifier with Large Bandwidth.—R. J. Morris, R. L. Khyl & M. W. P. Strandberg. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 80–81.) The bandwidth is 20 Mc/s at 10 dB gain, with a bath temperature of $4 \cdot 2$°K. The centre frequency is tunable in the range 8 400–9 700 Mc/s. Operation depends on a phonon-saturation mechanism in a ruby crystal.

621.375.9 : 538.569.4.029.64 **1145**

A Chromium Corundum Paramagnetic Amplifier and Generator.—G. M. Zverev, L. S. Kornienko, A. A. Manenkov & A. M. Prokhorov. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1660–1661.) Note on the operation of a 3-kMc/s paramagnetic amplifier and generator using a single crystal of $Al_2O_3Cr_2O_3$. Levels characterized by the quantum numbers $M = 3/2, \pm \frac{1}{2}$ when the crystalline axis is oriented parallel to the constant external magnetic field, are used. 3-kMc/s absorption lines for different power levels of the 15-kMc/s auxiliary radiation are shown.

621.375.9 : 538.569.4.029.66 **1146**

Molecular Amplifier and Generator for Submillimetre Waves.—A. M. Prokhorov. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1658–1659.) The rotational transitions of NH_3 molecules lie in the wavelength region below 1 mm. An amplifier can be constructed using a device in which radiation from one horn crosses a number of molecular beams and falls on a second horn. The maximum power obtainable is about 1 μW.

621.375.9 : 621.385.029.6 : 537.533 **1147**

New Method for Pumping a Fast Space-Charge Wave.—Wade & Adler. (See 1412.)

621.375.9.029.62 : 621.3.011.23 : 621.314.63 **1148**

Four-Terminal Parametric Amplifier.—K. K. N. Chang. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 81–82.) A lumped-constant model, using Ge junction diodes at a signal frequency of 214 Mc/s, has a gain of 8 dB, a bandwidth of $0 \cdot 25$ Mc/s and a noise factor of $2 \cdot 5$ dB.

621.375.9.029.64 : 621.3.011.23 : 621.314.63 **1149**

Microwave Parametric Amplifier by means of Germanium Diode.—B. Oguchi, S. Kita, N. Inage & T. Okajima. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 77–78.) Construction and performance of an amplifier for 4 kMc/s, using gold- and silver-bonded Ge diodes.

621.375.9.029.64 : 621.3.011.23 : 621.314.63 **1150**

The Reactatron—a Low-Noise, Semiconductor Diode, Microwave Amplifier.—F. A. Brand, W. G. Matthei & T. Saad. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 42–44.) Two nonlinear-capacitor microwave *p-n*-junction diodes are used in a balanced hybrid system. With a pump frequency of about twice the signal frequency of 2 900 Mc/s, power gains greater than 30 dB with a noise figure of 2·7 dB were obtained for a bandwidth of 0·5 Mc/s.

621.376.2.029.64 : 621.318.134 **1151**

Amplitude Modulation of Centimetre Waves by means of Ferroxcube.—H. G. Beljers. (*Philips tech. Rev.*, 14th Sept. 1956, Vol. 18, No. 3, pp. 82–86.) Losses in a ferroxcube rod, caused by a modulating magnetic field, vary the amplitude of a centimetre wave reflected from it. The nature of the losses and the maximum frequency at which they occur are examined theoretically. An experimental circuit has been constructed, operating at 9 300 Mc/s with a modulation frequency of 1·3 Mc/s; wide bandwidths are not possible.

621.376.23 : 621.317.373 **1152**

Diode Phase Detectors.—S. Krishnan. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 45–50.) The general case of unequal voltages applied to simple and balanced push-pull phase detectors is studied. Curves for design and evaluation of accuracy are given. The use of a signal greater than the reference voltage appears to be desirable in some applications.

621.376.23 : 621.385.029.6 **1153**

The Use of Beam Defocusing to Provide a Microwave Detector.—Castro & Needle. (See 1413.)

GENERAL PHYSICS

537.226 : 539.23 : 538.566.2 **1154**

Stokes' Equations and their Application to the Refractivity of Thin Films.—A. F. Wickersham, Jr. (*J. opt. Soc. Amer.*, Dec. 1958, Vol. 48, No. 12, pp. 958–964.) The problem of determining the microwave optical properties of *n* parallel planes of artificial dielectric from the properties of a single planar array is solved by using an extension of Stokes' equations.

537.523 : 538.56 **1155**

Oscillographic Study of R.F. Oscillations in 'Silent' Electrical Discharges.—P. S. V. Setty. (*Proc. nat. Inst. Sci. India*, Part A, 26th July 1958, Vol. 24, No. 4, pp. 235–239.) Oscillations produced under conditions of light or darkness in (*a*) iodine vapour and (*b*) hydrogen gas, in discharge tubes fitted with external 'sleeve' electrodes are described.

537.525 : 538.569.4 **1156**

Certain Phenomena Observed on Coupling an Oscillator and a Tube of Ionized Gas Introduced in the Oscillator Inductance.—T. V. Ionescu & O. C. Gheorghiu. (*C. R. Acad. Sci., Paris*, 30th June 1958, Vol. 246, No. 26, pp. 3598–3601.) An experimental study is reported in which an oscillator coil carrying a current $\ngtr$ 20 mA surrounds an energized gas-discharge tube. The effects of the h.f. current on the discharge are considered, in particular the luminous effects and the maintenance of the discharge after the energizing voltage is removed. See 1396 of 1958.

537.525.5 : 621.314.65 **1157**

A New Phenomenon of Electron Emission from Thin Mercury Films.—H. von Bertele. (*Nature, Lond.*, 25th Oct. 1958, Vol. 182, No. 4643, pp. 1148–1149.) Report of an experimental investigation of 'film emission' in a specially designed discharge tube, indicating that neither extremely high current densities nor cathode evaporation are intrinsic requirements for emission from liquid metal surfaces.

537.54 : 621.396.822.029.6 **1158**

Spectral Distribution of Thermal Noise in a Gas Discharge.—S. M. Bergmann. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 237–238.) "By means of thermodynamic considerations it is shown under which conditions microwave noise power generated by a gas discharge can be considered thermal. A critical analysis of Mumford's hypothesis is made."

537.56 **1159**

The Statistics of Plasma.—P. S. Pütter & F. Sauter. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 4–15.) Fundamental dynamic equations are derived comprising a system of partial differential equations.

537.56 **1160**

Calculation of Fields on Plasma Ions by Collective Coordinates.—A. A. Broyles. (*Z. Phys.*, 23rd April 1958, Vol. 151, No. 2, pp. 187–201. In English.)

537.56 : 537.311.3 **1161**

The Electrical Conductivity of a Plasma.—H. Schirmer & J. Friedrich. (*Z. Phys.*, 23rd April & 8th May 1958, Vol. 151, Nos. 2 & 3, pp. 174–186 & 375–384.) The exact theory of electrical conductivity of a plasma in the presence of a weak electric field is obtained by the integration of Boltzmann's equation for ionized gases, taking account of electron interaction.

537.56 : 538.63 **1162**

The Structure of Strong Collision-Free Hydrodynamic Waves.—J. H. Adlam & J. E. Allen. (*Phil. Mag.*, May 1958, Vol. 3, No. 29, pp. 448–455.) "A theoretical study has been made of the structure of strong 'hydromagnetic' waves which are propagated, across a magnetic field, in a low-density plasma where collisions can be neglected."

538.56 **1163**

Oblique Shock Waves in a Plasma with Finite Conductivity.—M. I. Kiselev & V. I. Tseplyaev. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1605–1609.)

538.56 : 537.56 **1164**

The 'Fourth Reflection Condition' for Electromagnetic Waves in a Plasma.—K. Rawer & K. Suchy. (*C. R. Acad. Sci., Paris*, 23rd June 1958, Vol. 246, No. 25, pp. 3428–3430.) The introduction of a term generally neglected in applying Boltzmann's equation modifies the expression for refractive index normally used in magneto-ionic theory.

538.561 : 539.2 **1165**

Coherent Spontaneous Microwave Emission by Pulsed Resonance Excitation.—L. E. Norton. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. MTT-5, No. 4, pp. 262–265.) "This paper describes an investigation of the coherent microwave emission from pulse-excited ammonia molecules. Coherent and periodic pulses of near resonance frequency and 1-μs duration excited the gas from its initial thermal condition. Self-induced coherent emission (molecular ringing) continued after the excitation field was removed. This radiation was observed during a period of 10 μs. In an actual experiment performed, a new Doppler bandwidth reduction method was used in the gas cell. The observed spectral width of the ammonia 7,7 line was about 5 kc/s. The emission was used to stabilize the excitation signal source to a short-term frequency stability of 2×10^{-10}."

538.566 **1166**

Propagation of an Electromagnetic Field in a Medium with Spatial Dispersion.—V. D. Shafronov. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1475–1489.) Investigation of the propagation of a transverse wave along a magnetic field in a plasma taking account of the thermal motion of electrons. Strong absorption occurs in the region at which Cherenkov radiation is possible.

538.566 : 535.4] + 534.2 **1167**

On the Diffraction and Reflection of Waves and Pulses by Wedges and Corners.—F. Oberhettinger. (*J. Res. nat. Bur. Stand.*, Nov. 1958, Vol. 61, No. 5, RP 2906, pp. 343–365.) "Various problems arising in the theory of the excitation of a perfectly reflecting wedge or corner by a plane, cylindrical, or spherical wave, are dealt with. The incident wave is represented by a line source (acoustic or electromagnetic) parallel to the edge. The spherical wave is emitted by an acoustic point source or by a Hertz dipole with its axis parallel to the edge. The case of an incident plane wave field is obtained as the limiting case (for large distances of the source from the edge) of the cylindrical or spherical wave excitation."

538.566 : 535.42 1168

The Diffraction of an Electromagnetic Plane Wave by a Perfectly Conducting Half-Plane.—P. Poincelot. (*C. R. Acad. Sci., Paris*, 23rd June 1958, Vol. 246, No. 25, pp. 3418–3419.)

538.566.2 : 548 1169

Electromagnetic Waves in Isotropic and Crystalline Media Characterized by Permittivity with Spatial Dispersion.—V. L. Ginzburg. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1593–1604.) The necessity of taking account of spatial dispersion in non-gyrotropic media in relation to the permittivity tensor and also in the analysis of longitudinal (plasma) waves propagating in an isotropic medium or along the principal dielectric axes in crystals is stressed. Collective energy losses and the Cherenkov effect are also considered.

538.569.4.029.64 1170

Attempt to Interpret the Shape of Paramagnetic Resonance Signals by the Introduction of a High-Frequency Demagnetizing Field.—A. Charru. (*C. R. Acad. Sci., Paris*, 23rd June 1958, Vol. 246, No. 25, pp. 3445–3447.) The theory accounts for the asymmetry of the resonance curve of organic radicals such as diphenyl picryl hydrazyl at 3 kMc/s.

538.569.4.029.64 : 061.3 1171

Microwave Physics.—K. W. H. Stevens. (*Nature, Lond.*, 25th Oct. 1958, Vol. 182, No. 4643, pp. 1121–1123.) Discussion of resonance techniques, particularly as applied for microwave amplification, with reference to three papers read at the British Association meeting (Section A) held in Glasgow, September 1958.

538.569.4.029.64 : 535.34 1172

Comparison of the Sensitivities of Paramagnetic-Electron-Resonance Spectrographs in the 3-cm Band.—J. Roch. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 59–62.) Modification of the coupling between a waveguide and a cavity is discussed for the following cases: (*a*) negligible background noise; (*b*) signal intensity of the same order as background noise.

539.2 : 548.0 1173

The Elastic Model of Lattice Defects.—J. D. Eshelby. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 116–121. In English.) It is shown that between rigid misfitting spheres in an elastic continuum there is a repulsive interaction if the spheres are bonded to the medium. The effect of other boundary conditions is discussed, and a simplified method is given for calculating the interaction between various types of defect.

539.2 : 548.0 1174

Energy States of a One-Dimensional Crystal with Vacant Lattice Sites.—A. Mozumder. (*Proc. nat. Inst. Sci. India*, Part A, 26th Sept. 1958, Vol. 24, No. 5, pp. 288–294.)

539.2 : 548.4 1175

Relations between the Concentrations of Imperfections in Solids.—F. A. Kröger & H. J. Vink. (*J. Phys. Chem. Solids*, May 1958, Vol. 5, No. 3, pp. 208–223.) Imperfections having opposite effective charges tend to increase each other's concentrations. Foreign atoms are incorporated in a manner dependent both on energy levels caused by them and on the type of imperfection prevailing in the crystal in their absence.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164.3 1176

On a Feature of Galactic Radio Emission.—H. Tunmer. (*Phil. Mag.*, April 1958, Vol. 3, No. 28, pp. 370–376.) "Observations of the intensity distribution of radio emission have shown a bright belt approximately normal to the galactic plane and passing through the anti-centre. An explanation is suggested in terms of the highly anisotropic radiation from relativistic electrons moving in the magnetic field of the local spiral arm. This suggestion avoids the supposition that the sun is in a special position in the galaxy."

523.164.3 : 535.417 1177

A Phase-Sensitive Interferometer Technique for the Measurement of the Fourier Transforms of Spatial Brightness Distributions of Small Angular Extent.—R. C. Jennison. (*Mon. Not. R. astr. Soc.*, Sept. 1958, Vol. 118, No. 3, pp. 276–284.) The r.f. interferometer described consists of three aerial systems tuned to 127 Mc/s.

523.164.32 : 550.385.4 1178

On the Power Spectrum of Solar Radio Outburst and its Relation to S.W.F. (Dellinger Phenomenon) and Geomagnetic Storm.—Y. Hakura. (*J. Radio Res. Labs, Japan*, Oct. 1958, Vol. 5, No. 22, pp. 283–293.) Solar radio outbursts are classified according to their power spectra. Those with power spectra increasing towards lower frequencies are associated with magnetic storms but not with S.I.D's, while those with spectra increasing towards higher frequencies are associated with S.I.D's without an accompanying magnetic storm. Outbursts with flat power spectra appear to combine the attributes of the other two types.

523.75 : 523.165 1179

The Possibility of Forecasting Solar Phenomena and their Terrestial Repercussions by the Study of the Intensity of Cosmic Radiation.—J. P. Legrand. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 70–73.) An analysis of observations made from 12th April 1957 to 31st March 1958 shows that a decrease in cosmic-ray intensity of 1–3% occurs 24–48 h before chromospheric eruptions which are followed by intense geomagnetic and ionospheric disturbances. This decrease also precedes or coincides with an increase in 200-Mc/s solar r.f. noise. See also 2396 of 1958 (Brown).

523.75 : 550.385 1180

Discrimination between Chromospheric Eruptions associated with Geomagnetic Disturbances during the Present Solar Maximum.—M. Lépineux. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 109–111.) An analysis of observations from 15th April 1957 to 31st March 1958 shows that some but not all solar flares are accompanied by a marked increase of r.f. noise at about 200 Mc/s.

523.75 : 551.510.535 1181

A Simple Method of Detecting Solar Flare Effects in the Lower Ionosphere.—E. Lauter & K. Sprenger. (*Z. Met.*, July 1958, Vol. 12, No. 7, pp. 205–210.) Field-strength records of transmitters operating at frequencies 150–300 kc/s over a distance of 1 300 km provide a sensitive means of detecting and identifying solar flare effects.

523.755 1182

Intensities, Polarization and Electron Density of the Solar Corona from Photographs taken at the Total Solar Eclipse of 1952 February 25.—H. von Klüber. (*Mon. Not. R. astr. Soc.*, Sept. 1958, Vol. 118, No. 3, pp. 201–223.)

550.385.4 1183

Sudden Commencements of Magnetic Storms at Tamanrasset.—J. L. Bureau. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 112–114.) An analysis of observations made at Tamanrasset (22° 48′ N, 5° 31′ E) from 1950 to 1956 inclusive.

550.389.2 : 629.19 1184

Scale Height of the Upper Atmosphere.—C. H. Bosanquet. (*Nature, Lond.*, 11th Oct. 1958, Vol. 182, No. 4641, pp. 1010–1011.) The effect of air drag on the orbit of an earth satellite is calculated allowing for the ellipticity of the earth.

550.389.2 : 629.19 1185

Sputnik as a Tool for Securing Geodetic Information.—L. Gold. (*J. Franklin Inst.*, Aug. 1958, Vol. 266, No. 2, pp. 103–107.)

550.389.2 : 629.19 1186

A Possible Cause of Air Resistance Changes in Satellite Orbits.—J. Bartels. (*Naturwissenschaften*, April 1958, Vol. 45, No. 8, p. 181.) The variation in the extent of the atmosphere in accordance with geomagnetic activity, particularly in the auroral zones, is suggested as a possible cause of observed changes in satellite orbital period.

550.389.2 : 629.19 1187

The Rotation of the First Russian Earth Satellite.—J. H. Thomson. (*Phil. Mag.*, Aug. 1958, Vol. 3, No. 32, pp. 912–916.) Fading records, obtained during transits over Cambridge at about 0330 U.T. 16th–22nd October 1957, have been examined. The 40-Mc/s records show a discontinuous change in fading rate at the time of nearest approach. This is attributed to changes in polarization in certain aspects of the satellite relative to the observer, caused by the generation of a cone by the rotating aerial. It is concluded that the satellite rotated at 6·5 rev/min, generating a cone of semi-angle 45°, the direction of the rotating axis being at 23° to the direction of the orbit at perigee.

550.389.2 : 629.19 **1188**

Determination of the Orbit of an Artificial Satellite.—N. Carrara, P. F. Checcacci & L. Ronchi. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, p. 75.) Theory of a method depending on Doppler measurements at four stations. No initial assumptions about the orbit are required.

550.389.2 : 629.19 **1189**

A New Method of Tracking Artificial Earth Satellites.—A. P. Willmore. (*Nature, Lond.*, 11th Oct. 1958, Vol. 182, No. 4641, pp. 1008–1010.) A photoelectric method of optical tracking has been developed which appears to have a precision comparable with astronomical observations. A record of observations of the rocket of Sputnik III on 23rd July 1958 at Tatsfield, Surrey, is shown.

550.389.2 : 629.19 **1190**

Tracking Orbits of Man-Made Moons.—C. A. Schroeder, C. H. Looney, Jr, & H. E. Carpenter, Jr. (*Electronics*, 2nd Jan. 1959, Vol. 32, No. 1, pp. 33–37.) Description of the Minitrack interferometer-type system designed to measure satellite position to an accuracy within 20 seconds of arc with a time precision of one millisecond.

550.389.2 : 629.19 **1191**

Doppler Satellite Measurements.—H. D. Tanzman, G. A. MacLeod & W. T. Scott. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 75–76.) A note on records obtained during successful and unsuccessful launchings.

550.389.2 : 629.19 **1192**

Further Radio Observations on Artificial Satellites.—P. F. Checcacci & C. Carreri. (*Ricerca sci.*, Sept. 1958, Vol. 28, No. 9, pp. 1817–1831.) Report on data obtained in Florence, Italy, from radio observations of satellites 1957 α, 1957 β, 1958 α, 1958 β 2 (Vanguard) and 1958 γ.

550.389.2 : 629.19 **1193**

Signal Strength Recordings of the Satellite 1958 δ2 (Sputnik III) at College, Alaska.—R. Parthasarathy & G. C. Reid. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 78–79.)

550.389.2 : 629.19 : 002.6 **1194**

World Data Centre for Rockets and Satellites, Slough.—(*Nature, Lond.*, 11th Oct. 1958, Vol. 182, No. 4641, p. 988.) It was agreed at the recent Moscow meeting of the Special Committee for the I.G.Y. to accept a British offer of a third centre where data from British investigations would be collected and exchanged with the centres in Washington and Moscow.

551.508.8 **1195**

Radiosonde Measurement of Vertical Electric Field and Polar Conductivity.—O. C. Jones, R. S. Maddever & J. H. Sanders. (*J. sci. Instrum.*, Jan. 1959, Vol. 36, No. 1, pp. 24–28.) A field mill and a Gerdian-type conductivity apparatus are described for use with a modified Kew Mk II radiosonde. These are used for measuring the electrical properties of the lower atmosphere and clouds.

551.510.535 **1196**

On the Semi-diurnal Lunar Variations in the Critical Frequency, Semi-thickness and Height of Maximum Ionization of the F_2 Layer in the Daytime.—T. Yonezawa & Y. Arima. (*J. Radio Res. Labs, Japan*, Oct. 1958, Vol. 5, No. 22, pp. 303–314.) Results obtained for f_0F_2 show reasonable agreement with those of other workers but contradict a theory of F_2-layer formation previously postulated [128 of January (Yonezawa)]. Results for other parameters are doubtful, owing to statistical errors, but appear to be exactly opposite in phase to those of other workers.

551.510.535 **1197**

Back-Scatter Ionospheric Sounding Experiments.—I. Ranzi. (*Note Recensioni Notiz.*, March/April 1958, Vol. 7, No. 2, pp. 201–212. English summary, pp. 213–215.) Report on tests at 18·6 Mc/s made near Rome from August to October 1957. Large discrepancies are found with results derived from vertical soundings. Subsequent tests are proceeding at 21·64 and 22·3 Mc/s.

551.510.535 : 551.513 **1198**

Ionospheric Drift Measurements in the Long-Wave Range as a Contribution to the Problem of General Circulation of the Upper Atmosphere.—K. Sprenger. (*Z. Met.*, July 1958, Vol. 12, No. 7, pp. 211–218.) Spaced-receiver measurements at 245 kc/s of ionospheric drift at altitude 90 km indicate distinct seasonal reversals in the circulation of the upper atmosphere. Small diurnal variations are superimposed on the normal drifts and an additional N–S component is observed during magnetic storms.

551.510.535"1958" : 621.396.11 **1199**

Ionosphere Review 1958.—T. W. Bennington. (*Wireless World*, Feb. 1959, Vol. 65, No. 2, pp. 72–73.) Solar activity, F_2-layer ionization and ionospheric and magnetic disturbances were less than those recorded in 1957.

551.510.535 **1200**

The Ionosphere. [Book Review]—K. Rawer. Publishers: Crosby Lockwood & Son, London, 1958, 202 pp., 42s. (*Nature, Lond.*, 25th Oct. 1958, Vol. 182, No. 4643, p. 1116.) A translation of the original German book (see 421 of 1954).

551.594 **1201**

Atmospheric Electricity. [Book Review]—J. A. Chalmers. Publishers: Pergamon Press, London, 1957, 327 pp., 63s. (*Met. Mag., Lond.*, Nov. 1958, Vol. 87, No. 1037, pp. 343–344.) Enlarged and revised edition of an earlier work (see 98 of 1950).

LOCATION AND AIDS TO NAVIGATION

621.396.93 : 061.3 **1202**

Radio Aids to Navigation.—(*Nature, Lond.*, 11th Oct. 1958, Vol. 182, No. 4641, pp. 978–980.) Summaries of three papers read at the British Association meeting (Section G, Engineering) held in Glasgow, September 1958.

621.396.933.2 : 621.396.677 **1203**

An Investigation of High-Frequency Direction-Finding Errors caused by Nearby Vertical Reradiators.—C. W. McLeish & R. S. Roger. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 58–60.) An expression for the maximum errors caused by reradiators is derived and the susceptibilities to error of three types of d.f. aerial are compared.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 **1204**

A New Type of Titanium Getter Pump.—L. Holland, L. Laurenson & J. T. Holden. (*Nature, Lond.*, 27th Sept. 1958, Vol. 182, No. 4639, pp. 851–852.) Description of recent developments in the design of vapour sources in which refractory supports are avoided.

533.5 : 621.385.029.6 **1205**

The Titanium Pump: a Device for the Maintenance of Vacuum in Electron Tubes.—H. Huber & M. Warnecke. (*Le Vide*, March/April 1958, Vol. 13, No. 74, pp. 84–90. In French & English.) A new type of getter-ion pump is described.

535.215 **1206**

The Influence of Defect Levels on Photoemission.—W. E. Spicer. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 555–563.) Minimum thermionic emission, maximum conductivity and optimum effect of band-loading are produced by p-type defects near the top of the valence band. If photoemission from defect levels is desired, n-type levels near the bottom of the conduction band are needed. The maximum quantum efficiency from defect levels is about 1%.

535.215 : 535.37 **1207**

The Intensity-Dependence of Photoconduction and Luminescence of Photoconductors in the Stationary State.—H. A. Klasens. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 175–200.) Experimental data are reviewed and two models for the electron-hole recombination processes are considered. The one-state model involving recombination either direct or via one impurity state, is shown to be generally inadequate, but the two-state model, with two discrete levels having different capture cross-sections for electrons and holes, can explain quantitatively many aspects of photoconduction and luminescence. 60 references.

535.215+535.37] : 546.482.21 **1208**

Fundamental Absorption Edge in Cadmium Sulphide.—D. Dutton. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 785–792.) The absorption and reflection spectra of CdS have been determined in the temperature range 90°–340 °K by photoelectric measurements on single crystals, using polarized light.

535.215 : 546.863.221 **1209**

Optical Properties of Antimony Trisulphide Films.—C. Kunze. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 165–172.) Measurements of absorption coefficient and refractive index on vapour-deposited Sb_2S_3 films were made at wavelengths of 400–1 000 mμ.

535.215 : 546.863.221 **1210**

Photoelectric Properties of Antimony Trisulphide Films.—C. Kunze. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 173–182.) The spectral distribution of photoconductivity in Sb_2S_3 films of thickness 200–400 mμ was measured using a vidicon-type camera tube.

535.37 **1211**

Transient Response of Phosphors.—G. I. Cohn & H. M. Musal. (*Trans. Inst. Radio Engrs*, June 1958, Vol. CP-5, No. 2, pp. 90–101. Abstract, *Proc. Inst. Radio Engrs*, Aug. 1958, Vol. 46, No. 8, p. 1553.)

535.37 : 546.472.21 **1212**

ZnS : Sn,Li Phosphor.—A. Wachtel. (*J. electrochem. Soc.*, July 1958, Vol. 105, No. 7, pp. 432–433.)

535.376 **1213**

The 'Memory' Effect in the Enhancement of Luminescence by Electric Fields.—G. Destriau. (*Z. Phys.*, 10th March 1958, Vol. 150, No. 4, pp. 447–455.) See also 2752 of 1958.

535.376 : 546.472.21 **1214**

Electron Traps and the Electroluminescence Brightness and Brightness Waveform.—F. F. Morehead, Jr. (*J. electrochem. Soc.*, Aug. 1958, Vol. 105, No. 8, pp. 461–468.) "This work describes an investigation of the effect of temperature, voltage, and rise time on the size and shape of the brightness waves of electroluminescence; sawtooth and square pulse voltage waveforms were used to excite Cu-activated ZnS phosphors in a slightly conducting medium."

535.376 : 546.472.21 **1215**

Electroluminescence of Insulated Particles.—K. Maeda. (*J. phys. Soc. Japan*, Nov. 1958, Vol. 13, No. 11, pp. 1352–1361.) The local high electric field is attributed to disturbance of the applied field by conductive substances in the phosphor. Distributions of local field are calculated and the relation between brightness and applied voltage is discussed. The estimated luminous efficiency of a ZnS cell agrees with the observed value.

535.376 : 546.482.21 **1216**

The $L_{2,3}$ Emission Spectrum and the Position of the $L_{2,3}$ Absorption Edge of S in CdS.—G. Eichhoff. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 55–69.)

537.227 **1217**

Energy Loss Processes in Ferroelectric Ceramics.—B. Lewis. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 17–24.) Measurements of permittivity, elastic compliance, and the electrical and mechanical loss coefficients of $Ba(Ti_{0.95}Zr_{0.05})O_3$ and $(Ba_{0.98}Ni_{0.02})TiO_3$, and their changes with time, are described. The observed losses can be accounted for entirely as macro- and micro-hysteresis loss associated with movement of domain boundaries.

537.227 : 547.476.3 **1218**

Upper Curie Temperature and Domain Structure in Various Regions of Rochelle Salt Crystals.—H. E. Müser & H. Flunkert. (*Z. Phys.*, 21st Dec. 1957, Vol. 150, No. 1, pp. 21–32.)

537.228.1 : 549.514.51 **1219**

Mechanical Resonance Dispersion in Quartz at Audio Frequencies.—E. R. Fitzgerald. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 765–784.) Measurements of the complex shear compliance of single crystals of quartz and fused quartz at frequencies from 100 to 5 000 c/s have resulted in the discovery of sharp resonances in the compliance. Analysis of the data on the basis of a generalized stress/strain relation involving a linear combination of strain and its first and second time derivatives gives a close fit to the experimental curves.

537.311.3 : 538.22 **1220**

Electrical Resistivity of Compounds with Ordered Spin Arrangements.—R. Parker. (*Phil. Mag.*, Aug. 1958, Vol. 3, No. 32, pp. 853–861.) "A summary is given of the published data of electrical resistivity of semiconductors which exhibit ordered spin arrangements below a Curie or Néel temperature. It is found generally that electrical anomalies accompany the advent of magnetic order only if the specific resistivity at the transformation temperature is less than a critical value. The implications of this rule are discussed in the light of recent theories of electrical conduction in spontaneously magnetized materials."

537.311.31 **1221**

The Quantum Theory of the High-Frequency Surface Impedance of a Metal.—M. Ya. Azbel'. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 105–117.)

537.311.31 : [546.86+546.87 **1222**

The Thermal and Electrical Resistivity of Bismuth and Antimony at Low Temperatures.—G. K. White & S. B. Woods. (*Phil. Mag.*, April 1958, Vol. 3, No. 28, pp. 342–359.)

537.311.33 **1223**

A Theory of Impurity Conduction: Part 1.—T. Kasuya. (*J. phys. Soc. Japan*, Oct. 1958, Vol. 13, No. 10, pp. 1096–1110.) Two ranges of impurity concentration in semiconductors such as Ge and Si are considered: (*a*) a fairly large concentration, with impurity levels merged into the conduction bands; (*b*) a fairly small concentration, with impurity levels separated from the conduction bands.

537.311.33 **1224**

A Theory of Impurity Conduction: Part 2.—T. Kasuya & S. Koide. (*J. phys. Soc. Japan*, Nov. 1958, Vol. 13, No. 11, pp. 1287–1297.) The case is considered of very low impurity concentration, where the fluctuation of local potential energy is much larger than the translational energy. In this case electron-phonon interaction is important since it enables electrons to move between localized states. Agreement between theory and experiment is fairly satisfactory. Part 1: 1223 above.

537.311.33 **1225**

Mobility of Electrons in Non-degenerate Semiconductors considering Electron-Electron Scattering.—M. S. Sodha & P. C. Eastman. (*Z. Phys.*, 27th Jan. 1958, Vol. 150, No. 2, pp. 242–246. In English.)

537.311.33 **1226**

The Crystal Structure and Properties of the Group VB to VIIB Elements and of Compounds Formed between them.—E. Mooser & W. B. Pearson. (*J. Phys. Chem. Solids*, Oct. 1958, Vol. 7, No. 1, pp. 65–77.)

537.311.33 **1227**

Semiconductor Compounds of Predominantly Homopolar Character.—H. Welker & H. Weiss. (*Z. Metallkde*, Nov. 1958, Vol. 49, No. 11, pp. 563–570.) Predominantly homopolar semiconductors are considered and the pronounced rectifier and transistor effects, as well as thermoelectric and photoelectric effects, exhibited by compounds such as InSb are discussed.

537.311.33 **1228**

Time-Dependent Changes in Excess Carrier Concentrations in the Presence of Surface Recombination.—J. D. Nixon & P. C. Banbury. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 54–58.) The decay of excess carriers in a thin slab of semiconductor is considered, assuming that surface recombination takes place through Shockley-Read type centres (420 of 1953). Numerical calculations for Ge show that the time constants differ from the steady-state lifetime except in the case of low trap densities.

537.311.33 : 535.215 **1229**

Role of Traps in the Photoelectromagnetic and Photoconductive Effects.—R. N. Zitter. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 852–855.) Expressions for the steady-state photoelectromagnetic and photoconductive currents are obtained which show that in certain cases the photoelectromagnetic current is determined by a lifetime different from the one determining the photoconductive current. This theory has been applied to p-type InSb.

537.311.33 : 535.215 **1230**

Optical Method for Determining Carrier Lifetimes in Semiconductors.—L. Huldt. (*Phys. Rev. Lett.*, 1st Jan. 1959, Vol. 2, No. 1, pp. 3–5.) Photoelectrically liberated carriers cause an increase in the infrared absorption of semiconductors. This increase is used to measure carrier lifetime in Ge.

537.311.33 : 537.32 **1231**

The Power Limits of Thermoelectric Effects in Semiconductors.—O. Böttger. (*Z. Phys.*, 8th May 1958, Vol. 151, No. 3, pp. 296–306.) The efficiency of a loaded thermocouple and the maximum temperature reduction of a Peltier element are calculated. A criterion for assessing the suitability of materials for thermoelectric devices is defined. Suitable semiconductors are restricted to the conductivity range $10^2 - 3 \times 10^3 \Omega^{-1} cm^{-1}$.

537.311.33 : 538.615 **1232**

Theory of Impurity Photo-ionization Spectrum of Semiconductors in Magnetic Fields.—R. F. Wallis & H. J. Bowlden. (*J. Phys. Chem. Solids*, Oct. 1958, Vol. 7, No. 1, pp. 78–89.) The theory shows two types of photo-ionization spectrum depending on the impurity ionization for zero field being larger or smaller than $heH/4\pi m^*c$. Detailed calculations are presented of the high-field spectrum for donor impurities in InSb.

537.311.33 : 538.63 **1233**

Surface Transport Theory.—J. N. Zemel. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 762–765.) A theory is presented for the dependence of the galvanomagnetic parameters on the surface potential of a semiconductor. The expressions for the conductivity reduce to that given by Schrieffer (2322 of 1955) when the magnetic field is zero. Formal equations for the magnetoconductivity and Hall coefficient are derived.

537.311.33 : 538.632 **1234**

Influence of the Geometry on Hall Effect and Magnetoresistance Effect in Rectangular Semiconductor Specimens.—H. J. Lippmann & F. Kuhrt. (*Naturwissenschaften*, April 1958, Vol. 45, No. 7, pp. 156–157.)

537.311.33 : 538.632 **1235**

Hall Effect in High Electric Fields.—J. F. Gibbons. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, p. 102.) Discussion of the limitation of Hall voltage in different semiconductor materials due to the terminal velocity of the charge carriers.

537.311.33 : 539.2 **1236**

Energy-Band Interpolation Scheme based on a Pseudopotential.—J. C. Phillips. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 685–695.) The effective potential for electrons near the Fermi level is split into two parts, the part due to the core and the part due to the other valence electrons. It is assumed that the relative effect of the core is small enabling two-parameter pseudopotentials to be constructed for diamond and silicon that give good agreement with orthogonalized plane-wave calculations and experiment at special points of the Brillouin zone, and also yield reasonable results for the bands at other points of the zone.

537.311.33 : 546.24 : 538.632 **1237**

Effect of Hydrostatic Pressure on the Anomalous Hall-Coefficient Reversal in Single-Crystal Tellurium.—A. Nussbaum, J. Myers & D. Long. (*Phys. Rev. Lett.*, 1st Jan. 1959, Vol. 2, No. 1, pp. 6–7.)

537.311.33 : [546.26 + 546.28 + 546.289 **1238**

The Electronic Structure of Diamond, Silicon and Germanium.—G. G. Hall. (*Phil. Mag.*, May 1958, Vol. 3, No. 29, pp. 429–439.) "The equivalent orbital method of describing the electronic structure of valence crystals, introduced in an earlier discussion of diamond [see *Proc. roy. Soc. A*, 7th Aug. 1950, Vol. 202, No. 1070, pp. 336–344], is extended in various ways."

537.311.33 : 546.26-1 : 538.63 **1239**

Magnetic Field Dependence of the Hall Effect and Magnetoresistance in Graphite Single Crystals.—D. E. Soule. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 698–707.)

537.311.33 : 546.26-1 : 538.63 **1240**

Analysis of Galvanomagnetic de Haas-van Alphen Type Oscillations in Graphite.—D. E. Soule. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 708–714.)

537.311.33 : 546.26-1 : 538.63 **1241**

Analysis of Multicarrier Galvanomagnetic Data for Graphite.—J. W. McClure. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 715–721.)

537.311.33 : [546.28 + 546.289 **1242**

Theoretical Calculation of Distribution Coefficients of Impurities in Germanium and Silicon, Heats of Solid Solution.—K. Weiser. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 118–126.) Distribution coefficients are calculated for group III, IV and V impurities in Ge and Si by estimating the change in energy involved in transferring the impurity atoms from a crystalline reservoir to the solid and molten host material respectively. Good agreement with experimental data is obtained.

537.311.33 : [546.28 + 546.289 **1243**

The Effect of Pressure on the Optical Absorption Edge of Germanium and Silicon.—T. E. Slykhouse & H. G. Drickamer. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 210–213.)

537.311.33 : 546.28 **1244**

Absorption Spectrum of Arsenic-Doped Silicon.—H. J. Hrostowski & R. H. Kaiser. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 236–239.) Experimental results agree well with results obtained from the effective mass theory. Bands not obeying the calculated term scheme are shown to arise from other donor impurities. Optical and thermal ionization energies appear to be very similar for both group III and group V impurities.

537.311.33 : 546.28 **1245**

Electropolishing Silicon in Hydrofluoric Acid Solutions.—D. R. Turner. (*J. electrochem. Soc.*, July 1958, Vol. 105, No. 7, pp. 402–408.)

537.311.33 : 546.28 : 535.215 **1246**

Measurement of Surface Recombination Velocity in Silicon by Steady-State Photoconductance.—H. M. Bath & M. Cutler. (*J. Phys. Chem. Solids*, May 1958, Vol. 5, No. 3, pp. 171–179.) In the method described, excitation can be accomplished by penetrating as well as non-penetrating light, so that the components of lifetime which depend respectively on the bulk and surface properties of the sample can be emphasized. Application of the method to Si indicates that surface recombination processes are different from those of Ge.

537.311.33 : 546.289 **1247**

New Observations on Dislocations in Germanium.—H. Dorendorf. (*Z. angew. Phys.*, Oct. 1957, Vol. 9, No. 10, pp. 513–519.) The origin and extent of dislocations and their elimination by annealing is discussed with reference to series of photographs.

537.311.33 : 546.289 **1248**

High-Vacuum Studies of Surface Recombination Velocity for Germanium.—H. H. Madden & H. E. Farnsworth. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 793–800.) Details are given of the observed changes in the value of the surface recombination velocity of minority carriers for (100) faces of Ge crystals under various vacuum conditions.

537.311.33 : 546.289 **1249**

Free-Carrier Absorption in *n*-Type Ge.—R. Rosenberg & M. Lax. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 843–852.) The structure of the conduction band of Ge is completely taken into account in calculating the cross-section in second-order Born approximation. An excellent fit to data at 450 °K is obtained.

537.311.33 : 546.289 **1250**

Comparison of Radio-Copper and Hole Concentrations in Germanium.—K. Wolfstirn & C. S. Fuller. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 141–145.) The changes in carrier concentration for four ^{64}Cu concentrations were determined in Ga-doped and As-doped Ge, by resistivity measurements. The results are in reasonable agreement with calculations based on published values of the ionization energy levels for Cu in Ge [2192 of 1957 (Woodbury & Tyler)].

537.311.33 : 546.289 **1251**

Melted-Layer Crystal Growth and its Application to Germanium.—F. H. Horn. (*J. electrochem. Soc.*, July 1958, Vol. 105, No. 7, pp. 393–395.) Single-crystal material of constant impurity distribution may be grown conveniently from a doped melted layer maintained above solid retained in a crucible.

537.311.33 : 546.289 **1252**

Etching and Polishing of Germanium Surfaces.—D. Geist & E. Preuss. (*Z. angew. Phys.*, Oct. 1957, Vol. 9, No. 10, pp. 526–531.) Comparison of the effectiveness of various methods.

537.311.33 : 546.289 : 535.215 : 538.639 **1253**

A Quadratic Photoelectromagnetic Effect in Germanium.—M. Cardona & W. Paul. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 127–140.) The quadratic effect occurs when the magnetic field is rotated about an axis lying in the illuminated face and perpendicular to the original direction of the field. Its relation to the usual linear photoelectromagnetic effect is investigated theoretically and experimentally, and satisfactory agreement obtained. The ratio of the quadratic to the linear photoelectromagnetic field is, for small light intensities, a function only of the Hall angles, the magnetoresistance coefficients of electrons and holes, the impurity concentration, and the magnetic field. See also 2460 of 1958 (Kikoin & Bykovskiĭ).

537.311.33 : 546.289 : 539.164 **1254**

Alpha-Particle Irradiation of Ge at 4·2 °K.—G. W. Gobeli. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 732–739.) The radiation removed carriers from the samples, no measureable thermal recovery being observed below 22 °K. Two distinct regions of thermal recovery were observed, the maximum rates of change occurring near 33 °K and 67 °K.

537.311.33 : 546.289 : 621.396.822 **1255**

Study on the Correlation between the Noise by Hole Generation and Surface Recombination Velocity at Ge Fused Junction.—K. Komatsubara. (*J. phys. Soc. Japan*, Nov. 1958, Vol. 13, No. 11, pp. 1409–1410.)

537.311.33 : 546.289 : 621.396.822 **1256**

1/*f* Noise in Germanium Devices.—T. B. Watkins. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 59–68.) Simultaneous field-effect and noise measurements were made on specially prepared Ge junction diodes to test current theories of 1/*f* noise. As a result a mechanism is proposed which is based on Petritz's theory involving local barrier breakdown.

537.311.33 : 546.289 : 621.396.822 **1257**

Excess Noise in Narrow Germanium *p-n* Junctions.—T. Yajima & L. Esaki. (*J. phys. Soc. Japan*, Nov. 1958, Vol. 13, No. 11, pp. 1281–1287.) Alloyed junctions less than 200Å wide, showing inverted rectification and negative resistance in the forward direction, gave considerable 1/*f* noise in a certain forward-bias range. The noise was stable and surface-insensitive and mainly associated with excess forward current. Measurements were made at temperatures down to 77 °K. The interpretation of results is discussed.

537.311.33 : 546.492.241 **1258**

Preparation and Electrical Properties of Mercury Telluride.—T. C. Harman, M. J. Logan & H. L. Goering. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 228–235.) Details are given of the preparation, by direct reaction of Hg vapour with liquid Te, zone melting and annealing of HgTe. The composition, near stoichiometry, could be altered by the anneal. Hall coefficient and resistivity were measured as a function of temperature.

537.311.33 : 546.681.19 **1259**

Diffusion of Zinc in Gallium Arsenide.—J. W. Allen & F. A. Cunnell. (*Nature, Lond.*, 25th Oct. 1958, Vol. 182, No. 4643, pp. 1158.) A theoretically derived curve showing impurity concentration as a function of depth for a diffusion temperature of 1 000 °C, is confirmed experimentally.

537.311.33 : 546.682.86 **1260**

Electrical Conduction in *p*-Type InSb between 100° and 2° K.—E. H. Putley. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 128–131.)

537.311.33 : 546.682.86 **1261**

Nuclear Magnetic Resonance in Indium Antimonide: Part 1 — The Effect of Impurities.—E. H. Rhoderick. (*Phil. Mag.*, June 1958, Vol. 3, No. 30, pp. 545–563.) In *n*-type (Te-doped) and *p*-type (Zn-doped) samples of InSb the amplitude and width of the ^{115}In nuclear magnetic resonance are very sensitive to the presence of donors or acceptors. This is ascribed to broadening arising from the interaction of the nuclear quadrupole moment with the Coulomb field of the ionized impurity atoms.

537.311.33 : 546.682.86 **1262**

Nuclear Magnetic Resonance in Impure Indium Antimonide.—M. H. Cohen. (*Phil. Mag.*, June 1958, Vol. 3, No. 30, pp. 564–566.) An analysis of experimental results (1261 above) on nuclear quadrupole broadening in InSb based on a detailed theory of line shape.

537.311.33 : 546.682.86 : 538.63 **1263**

Magnetically Induced Impurity Banding in *n*-InSb.—R. J. Sladek. (*J. Phys. Chem. Solids*, May 1958, Vol. 5, No. 3, pp. 157–170.) "A detailed study of the effect of a strong magnetic field on donor levels in *n*-InSb has been made by means of Hall-effect and resistivity measurements down to 1·5 °K, using field strengths up to 28 000 G. The results of these measurements indicate that donor levels which are split off from the conduction band under the influence of the magnetic field form a narrow impurity band with finite mobility. The position of the impurity band (i.e. donor ionization energy) and the mobility of electrons in the impurity band have been determined as a function of magnetic field strength."

537.311.33 : 546.682.86 : 538.63 **1264**

An Oscillatory Transverse Magnetoresistance Effect in *n*-Type InSb.—E. H. Putley. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 131–133.)

537.311.33 : 546.682.86 : 538.63 **1265**

Interpretation of the Transverse Magnetoresistance in *p*-Type Indium Antimonide at Liquid-Nitrogen Temperature.—C. H. Champness. (*Phys. Rev. Lett.*, 15th Dec. 1958, Vol. 1, No. 12, pp. 439–440.)

537.311.33 : 546.714-31 **1266**

Study of the Semiconducting Properties of Pyrolusite.—J. N. Das. (*Z. Phys.*, 8th May 1958, Vol. 151, No. 3, pp. 345–350. In English.)

537.311.33 : 546.817.221 **1267**

Magnetoresistance in PbS, PbSe, and PbTe at 295°, 77·4°, and 4·2 °K.—R. S. Allgaier. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 828–836.) Magnetoresistance measurements at 77·4 °K conformed to the general phenomenological weak-field theory. Saturation effects and deviations from the sinusoidal behaviour were observed at 4·2 °K. The data generally favoured a model with considerable mass anisotropy.

537.311.33 : 546.863.241 : 537.32 **1268**

Thermoelectric Properties of Antimony Telluride and the Solid Solutions Sb_2Te_3-Bi_2Te_3.—H. Benel. (*C. R. Acad. Sci., Paris*, 4th Aug. 1958, Vol. 247, No. 5, pp. 584–587.) Thermoelectric power and electrical resistivity were measured and the effect of doping was investigated.

537.311.33 : 548.0 **1269**

Ternary Semiconducting Compounds with Sodium-Chloride-Like Structure: $AgSbSe_2$, $AgSbTe_2$, $AgBiS_2$, $AgBiSe_2$.—S. Geller & J. H. Wernick. (*Acta cryst.*, 10th Jan. 1959, Vol. 12, Part 1, pp. 46–54.) Crystallographic data on high- and low-temperature modifications are discussed.

537.311.33 : 669.018.13 **1270**

Constitution of the $AgSbSe_2$-$AgSbTe_2$-$AgBiSe_2$-$AgBiTe_2$ System.—J. H. Wernick, S. Geller & K. E. Benson. (*J. Phys. Chem. Solids*, Nov. 1958, Vol. 7, Nos. 2/3, pp. 240–248.)

537.311.4 **1271**

Induced Conductivity at the Surface of Contact between Metals.—S. D. Chatterjee & S. K. Sen. (*Phil. Mag.*, Aug. 1958, Vol. 3, No. 32, pp. 839–852.) A report is given of experimental investigations into the variations in contact resistance between potassium and aluminium surfaces when the junction is under the influence of an electrical stress.

537.32 : 546.23-17 **1272**

Thermoelectric Observations on Grey Selenium.—J. I. Carasso & R. W. Pittman. (*Nature, Lond.*, 11th Oct. 1958, Vol. 182, No. 4641, p. 1011.) Experiments on spectrographically pure Se showed that the thermoelectric power of single crystals is independent of temperature while that of oriented polycrystals increases with temperature.

538.22 **1273**

The Influence of Hydrostatic Pressure on the Magnetizability of

Au_2Mn.—K. H. v. Klitzing & J. Gielessen. (*Z. Phys.*, 10th March 1958, Vol. 150, No. 4, pp. 409–414.)

538.22 : 537.311.3 1274

The Magnetic Susceptibility and Electrical Resistivity of some Transition-Metal Silicides.—D. A. Robins. (*Phil. Mag.*, April 1958, Vol. 3, No. 28, pp. 313–327.)

538.221 1275

Thermal Activation of Ferromagnetic Domains.—F. D. Stacey. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 136–138.) By similar analysis to that used in the theory of electrical noise it is shown that the factor C in the expression $C \exp(-E/kT)$ for the probability of energy E being available to overcome potential barriers in domain wall movements, is proportional to the absolute temperature T.

538.221 1276

Orientational Superstructure arising from Mechanical Deformation of an Fe-Ni Alloy.—R. Vergne. (*C. R. Acad. Sci., Paris*, 16th July 1958, Vol. 247, No. 2, pp. 197–200.) Samples of an Fe-Ni alloy under tensile stress were heated above the Curie point for 75 min and then quenched. The energy of magnetization subsequently observed was found to be a function of the tension.

538.221 : 538.632 1277

Theory of the Hall Effect in Ferromagnetic Substances.—J. M. Luttinger. (*Phys. Rev.*, 1st Nov. 1958, Vol. 112, No. 3, pp. 739–751.) The Hall effect is computed on the basis of a simple model making use of the transport theory of Kohn & Luttinger (1105 of 1958). The calculation is rigorous, but assumes a slowly varying scattering potential, a simple band, and very few conduction electrons. None of these assumptions is very realistic, but they enable a discussion to be given of the types of contribution which can occur.

538.221 : 538.65 1278

The Damping of High-Frequency Elastic Waves in Cubic Ferromagnetic Single Crystals.—G. Simon. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 23–35.) The inhomogeneity of magnetization due to domain formation is shown to give rise to a damping maximum, when the depth of skin-effect penetration becomes comparable with the mean dimensions of the domains.

538.221 : 539.23 : 538.632 1279

Ferromagnetic Thin Films. Hall Effect in Thin Films of Nickel.—G. Goureaux, P. Huet & A. Colombani. (*C. R. Acad. Sci., Paris*, 16th July 1958, Vol. 247, No. 2, pp. 189–193.)

538.221 : 539.23 : 538.632 1280

Measurements of the Hall Effect in Ferromagnetic Nickel Films.—L. Reimer. (*Z. Phys.*, 14th Feb. 1958, Vol. 150, No. 3, pp. 277–286.) Results obtained with vapour-deposited films of thickness 20–2 000 Å are discussed.

538.221 : 539.234 : 538.61 1281

The Observation of Weiss Domains in Polycrystalline Material using the Magnified Magneto-optical Kerr Rotation.—J. Kranz & W. Drechsel. (*Z. Phys.*, 8th April 1958, Vol. 150, No. 5, pp. 632–639.) Application of the technique described in 841 of 1957 (Kranz) to observation of the specimen at an angle of about 55°.

538.221 : 621.3.042.15 : 538.662 1282

Investigations on the Temperature Coefficient of the Initial Permeability of Dust Core Materials.—H. Henniger. (*NachrTech.*, Feb. 1958, Vol. 8, No. 2, pp. 66–75.) Test equipment is described and the results of measurements on cores prepared from various materials are analysed in order to improve manufacturing techniques.

538.221 : [621.318.124+621.318.134 1283

Domain Configurations on Ferrites.—D. J. Craik & P. M. Griffiths. (*Proc. phys. Soc.*, 1st Jan. 1959, Vol. 73, No. 469, pp. 1–13, plates.) The domain structures of Mn, MnZn, Ni, Cu and Ba ferrite crystals were examined by conventional and modified Bitter figure techniques. The Mn, Ni and Ba ferrites gave patterns as expected from their known magnetic constants. The Cu ferrite patterns could not be interpreted exactly. The MnZn ferrite gave anomalous patterns considered to be due to a strain-induced uniaxial anisotropy.

538.221 : 621.318.124 1284

Magnetic Materials with Perminvar Effect.—A. v. Kienlin. (*Z. angew. Phys.*, Oct. & Dec. 1957, Vol. 9, Nos. 10 & 12.)

Part 1—Contribution to the Understanding of the Perminvar Effect and of its Dependence on Temperature (pp. 520–526).

Part 2—The Significance of Composition for the Perminvar Effect in Ferrites with Low Cobalt Content (pp. 631–640).

538.221 : 621.318.134 1285

Phenomenological Theory of Kinetic Processes in Ferromagnetic Dielectrics.—M. I. Kaganov & V. M. Tsukernik. (*Zh. eksp. teor. Fiz.*, June 1958, Vol. 34, No. 6, pp. 1610–1618.) Calculation of the relaxation time resulting from the interaction of spin waves.

538.221 : 621.318.134 1286

Electromagnetic Fields in Ferrite Ellipsoids.—R. A. Waldron. (*Brit. J. appl. Phys.*, Jan. 1959, Vol. 10, No. 1, pp. 20–22.) The field in a ferrite immersed in a uniform magnetic field is calculated with the aid of the permeability tensor for a number of ferrite shapes.

538.221 : 621.318.134 1287

Effect of Oxygen Pressure on Microstructure and Coercive Force of Magnesium Ferrite.—R. E. Carter. (*J. Amer. ceram. Soc.*, Dec. 1958, Vol. 41, No. 12, pp. 545–550.)

538.221 : 621.318.134 1288

Dispersion in the Complex Transverse Susceptibility of Lithium Ferrite in the Frequency Range 10–10 000 Mc/s.—F. Voigt. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 86–101.) The results of measurements on $LiFe_5O_8$ ring cores are discussed.

538.221 : 621.318.134 1289

The Growth of Single Crystals of Magnetic Garnets.—J. W. Nielson & E. F. Dearborn. (*J. Phys. Chem. Solids*, May 1958, Vol. 5, No. 3, pp. 202–207.) Crystals of magnetic garnets, $Y_3Fe_5O_{12}$, $Sm_3Fe_5O_{12}$, $Er_3Fe_5O_{12}$ and $Gd_3Fe_5O_{12}$ suitable for many research purposes have been grown from iron-rich melts of lead oxide.

538.221 : 621.318.134 1290

Interpretation of the Magnetic Properties of Erbium Garnet in which the Al^{3+} and Cr^{3+} Ions have been Substituted for Fe^{3+} Ions.—G. Villers, J. Loriers & R. Pauthenet. (*C. R. Acad. Sci., Paris*, 4th Aug. 1958, Vol. 247, No. 5, pp. 587–590.)

538.221 : 621.318.134 1291

Initial and Remanent Permeability Spectra of Yttrium Iron Garnet.—R. D. Harrington & A. L. Rasmussen. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, p. 98.)

538.221 : 621.318.134 : 534.232 1292

Effect of Divalent Ion Substitutions on the Magnetomechanical Properties of Nickel Ferrite.—S. F. Ferebee & C. M. Davis, Jr. (*J. acoust. Soc. Amer.*, Aug. 1958, Vol. 30, No. 8, pp. 747–750.) Magnetostrictive properties of Co-substituted Ni ferrous ferrite are investigated to determine its suitability for use in electromechanical transducers.

538.221 : 621.318.134 : 621.317.411 1293

Measurements of the Permeability Tensor for 'Ferroxcube' at 24 000 Mc/s.—G. G. Robbrecht & J. L. Verhaeghe. (*Nature, Lond.*, 18th Oct. 1958, Vol. 182, No. 4642, p. 1080.) Measurements have been made on spherical and disk samples of ferroxcube IV*B* in a TE_{112} resonant cavity excited by circularly polarized waves.

538.23 : 538.221 1294

The Influence of Temperature in the Creep of Asymmetric Hysteresis Cycles.—Nguyen-Van-Dang. (*C. R. Acad. Sci., Paris*, 7th July 1958, Vol. 247, No. 1, pp. 56–59.) Experiments at 290°, 77° and 20 °K show that temperature has little influence on creep.

539.234 : 546.72 1295

Stability of Evaporated Films.—M. W. Roberts. (*Nature, Lond.*, 25th Oct. 1958, Vol. 182, No. 4643, pp. 1151–1152.) Thin Fe films show a rapid reduction of surface area to 23% of the original value as a result of the chemisorption of O_2.

621.315.61-41 1296

Tape and Film Insulation for Electronic Equipment.—G. Sideris. (*Electronics*, 2nd Jan. 1959, Vol. 32, No. 1, pp. 42–43.) A brief review of the properties and applications of some of the newer forms of insulating material.

MATHEMATICS

512 : 621.316.5 **1297**

The Solution of Equations in Switching Algebra.—H. Zemanek. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 35–44.)

517.512.2 **1298**

Fourier Analysis, a New Numerical Method.—L. Hyvärinen. (*Acta polyt. scand.*, 1958, No. 248, Ma 2, 19 pp.) Evaluation is facilitated by the use of templates.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) **1299**

A Very Precise Short-Period Comparator for 100-kc/s Frequency Standards.—A. M. Thompson & R. W. Archer. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 61–64.) The instrument described uses addition, amplification and phase detection; an accuracy of frequency comparison within 1 part in 10^{13} has been obtained.

621.3.018.41(083.74) **1300**

A Microwave Frequency Standard.—B. H. L. James & M. T. Stockford. (*Electronic Engng*, Jan. & Feb. 1959, Vol. 31, Nos. 371 & 372, pp. 2–7 & 82–87.) Reference frequencies within the range 7–20 kMc/s are derived as harmonics from a 100-kc/s temperature-controlled crystal oscillator with a short-term stability better than 1 in 10^7. Several lower-frequency references are provided by frequency division down to 50 c/s. Full descriptions of the circuits are given.

621.3.018.41(083.74) : 529.786 : 525.35 **1301**

The Random Variation in the Speed of Rotation of the Earth.—A. Stoyko & N. Stoyko. (*C. R. Acad. Sci., Paris*, 16th July 1958, Vol. 247, No. 2, pp. 182–185.) A comparison of U.T.2 with the Cs resonator of the National Physical Laboratory is made for 1955–1958 based on the work of Essen et al. (3195 of 1958).

621.3.089.6 : 061.6 **1302**

Services and Facilities of the Electronic Calibration Centre.—(*Tech. News Bull. nat. Bur. Stand.*, Nov. 1958, Vol. 42, No. 11, pp. 223–229.) Details of the department recently opened at Boulder, Colorado, for the calibration of laboratory standards in terms of the national standards maintained by the N.B.S. See 858 of 1958.

621.317(083.7) : 061.3 **1303**

Conference on Electronic Standards and Measurements.—(*Tech. News Bull. nat. Bur. Stand.*, Nov. 1958, Vol. 42, No. 11, pp. 209–217.) Report on the conference held at Boulder, Colorado, 13th–15th August 1958. The technical program consisted of the following sessions:

(*a*) The Relationship of Standards to Physical Constants;
(*b*) Frequency and Time Interval Standards;
(*c*) Direct-Current and Low-Frequency Standards;
(*d*) Radio-Frequency Standards;
(*e*) Microwave Standards;
(*f*) Organization and Operation of Standards Laboratories.

621.317.3 : 621.314.7 **1304**

Measurement of Transistor Characteristics in the 3–250-Megacycle Frequency Range.—J. H. O'Connell & T. M. Scott. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 598–616.) Description of apparatus and techniques developed for the measurement of the input, output and unilateralization impedances, transconductance and current gain.

621.317.335 **1305**

Simple Apparatus for Measuring Dielectric Constants and Losses from 10 c/s to 50 kc/s.—J. C. S. Richards. (*J. sci. Instrum.*, Jan. 1959, Vol. 36, No. 1, pp. 22–23.) A bridge circuit with a battery-operated detector is described in which no coupling transformer or Wagner earth is used. Measurements are made to an accuracy within about ± 1%.

621.317.336 : 537.311.62 **1306**

Surface Impedance.—J. C. Anderson. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 56–60.) The surface impedance of a conducting material is measured at v.h.f. by the use of a coaxial transmission line terminated by a disk of the material. Theory, experimental method, and results for pure Ni are given. A calibrated r.f. meter is not required.

621.317.34 : 621.372.2 **1307**

Novel Method for Measuring Impedances on Surface-Wave Transmission Lines.—G. Schulten. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 76–77.)

621.317.34.029.64 : 621.317.755 **1308**

A Microwave Reflectometer Display System.—J. C. Dix & M. Sherry. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 24–29.) The instrument described gives a continuous display of the matching of an X-band waveguide component over the frequency band 7 500–11 000 Mc/s.

621.317.35 : 621.396.822 **1309**

Spectrum Analysis of Random Noise Generators.—L. G. Polimerou. (*Commun. & Electronics*, Nov. 1958, No. 39, pp. 672–679.) A method is described using a selective filter and a variable chopper-carrier. Possible limitations and errors are discussed.

621.317.351 : 621.317.619 : 621.396/.397 **1310**

A Method of Measurement and Oscillographic Recording of Phase Characteristics for Wide-Band Transmission Systems in the Range 0·1–10 Mc/s.—W. Rietz. (*Z. angew. Phys.*, Oct. 1957, Vol. 9, No. 10, pp. 489–495.) The equipment described [see also 219 of 1958 (Kroebel & Wegner)] has a phase error of ± 0·05° at 13 kc/s for input amplitude changes of 20 dB.

621.317.361 : 529.786 **1311**

Efficiency of Frequency Measurements with an Atomic Clock.—M. Peter & M. W. P. Strandberg. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 92–93.)

621.317.382.029.64 : 538.632 **1312**

Power Measurement at 4 Gc/s by the Application of the Hall Effect in a Semiconductor.—L. M. Stephenson & H. E. M. Barlow. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 27–30.) Description of an instrument developed from an earlier design [see 1488 of 1956 (Barlow & Stephenson)]. The range at 4 kMc/s is 30 mW–20 W with a ± 3 % error at any s.w.r. between unity and 0·1. Only about 3·4 % of the power measured is absorbed. Measurements made using a crystal within the power-carrying waveguide are also described.

621.317.44 **1313**

A New Method of Measuring Susceptibility.—C. Hilsum & A. C. Rose-Innes. (*Nature, Lond.*, 18th Oct. 1958, Vol. 182, No. 4642, p. 1082.) A magnetic analogue of a Wheatstone bridge with a permanent magnet as the source of magnetomotive force is described. It may be used to measure a very weak susceptibility such as that of deionized water.

621.317.44 : 538.248 : 539.23 **1314**

The Measurement of Residual Magnetism of Thin Films.—S. Yamaguchi. (*Naturwissenschaften*, May 1958, Vol. 45, No. 9, p. 205.) A method based on the Lorentz effect is described.

621.317.7.089.6 : 621.372.56.029.64 **1315**

The Calibration of Microwave Attenuators by an Absolute Method.—Laverick. (See 1123.)

621.317.733.011.4 **1316**

The Influence of the Leads on the Capacitance of Capacitors.—G. Zickner & W. Wiessner. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 70–85.) Error sources in capacitance-bridge measurements by substitution methods are discussed. The elimination of the error due to the separation of the lead from the capacitor is described, and the magnitude of the effect is investigated by means of measurements on two- and three-plate capacitors.

621.317.755 **1317**

Simple, High-Sweep-Speed, Single-Stroke Oscilloscope.—W. P. Baker. (*J. sci. Instrum.*, Jan. 1959, Vol. 36, No. 1, pp. 30–31.) The advantages of applying the accelerating voltage to the c.r. tube in the form of a long-tailed pulse instead of using a steady d.c. source and grid bias are noted, and circuit arrangement are shown.

621.317.755 : 621.372.5 **1318**

Image Distortion by *RC* Quadripoles of a Cathode-Ray Oscillograph and its Correction at Low Frequencies.—H. Wittke. (*Elektronische Rundschau*, March

1958, Vol. 12, No. 3, pp. 89–93.) Correction formulae are derived for reconstituting the original signal waveform from the output of a *RC* network. Difficulties of correcting circuit distortion in c.r. oscilloscopes are outlined.

621.317.77.029.4 1319

A Low-Frequency Phasemeter.—N. Hambley. *Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 13–15.) The phasemeter described gives a quick and accurate measurement in the frequency range 1–100 c/s.

621.317.784 1320

A Precision Thermoelectric Wattmeter for Power and Audio Frequencies.—J. J. Hill. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, p. 64.) Discussion on 1226 of 1958.

621.317.79.029.6 : 535.322.4 1321

Compact Microwave Refractometer for Use in Small Aircraft.—M. C. Thompson, Jr, & M. J. Vetter. (*Rev. sci. Instrum.*, Dec. 1958, Vol. 29, No. 12, pp. 1093–1096.) The refractometer operates on the principle developed by Birnbaum (see *ibid.*, Feb. 1950, Vol. 21, No. 2, pp. 169–176) using a frequency-modulated klystron oscillator. A two-channel recording system is described which provides a simultaneous indication of profile and turbulence data. The r.m.s. noise levels are approximately 0·05 and 0·01 *N* units respectively.

621.317.794.029.62/.63 1322

An Aperiodic Barretter Probe for Power Measurements at $f = 30$ to 1 500 Mc/s, and its Application.—H. Rieck. (*Nachr Tech.*, Feb. 1958, Vol. 8, No. 2, pp. 50–55.) A balanced twin-barretter mount is described for use as a bolometer at powers up to 10 mW.

621.317.799 : 621.397.6 1323

A Television Waveform Generator using Transistors.—Rozner. (See 1373.)

621.317.799 : 621.397.62 : 535.623 1324

A One-Tube Crystal-Filter Reference Generator for Colour TV Receivers.—R. H. Rausch & T. T. True. (*Trans. Inst. Radio Engrs*, June 1957, Vol. BTR-3, No. 1, pp. 2–7.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

611.84.001.57 : 621.372.029.64 1325

Characteristics of a Model Retinal Receptor Studied at Microwave Frequencies.—J. M. Enoch & G. A. Fry. (*J. opt. Soc. Amer.*, Dec. 1958, Vol. 48, No. 12, pp. 899–911.) A simplified microwave model of the human eye is described. Tests were carried out at 2·42 and 3·20 cm λ.

612.3 : 621.3.083.7 1326

New Method of Exploring the Digestive System by means of an Ingestible Radio Capsule.—M. Marchal & M. T. Marchal. (*C. R. Acad. Sci., Paris*, 23rd June 1958, Vol. 246, No. 25, pp. 3519–3520.)

616 : 621.3.083.7 : 621.314.7 1327

Experiments with an Ingestible Intestinal Transmitter.—M. v. Ardenne & H. B. Sprung. (*Naturwissenschaften*, April 1958, Vol. 45, No. 7, pp. 154–155.) The transmitter, which is 26 mm long and 10 mm in diameter, operates at 1·5 kc/s; this frequency can increase up to 5 kc/s in accordance with pressure. See also 3611 of 1957 (Mackay & Jacobson).

621.384.62 1328

Radio-Frequency Aspects of Electronuclear Accelerators.—A. F. Harvey. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 43–57.) Survey of various machines with emphasis on r.f. aspects of design and performance. 176 references.

621.384.622.2 1329

Anomalous Attenuation in Linear Electron Accelerators.—V. J. Vanhuyse. (*Nature, Lond.*, 18th Oct. 1958, Vol. 182, No. 4642, pp. 1081–1082.) Anomalous attenuation is explained on the basis of the 'multipactor' effect.

621.385.833 1330

Emission Microscopy with Secondary Electrons (15-keV Primary Electrons).—W. Bayh. (*Z. Phys.*, 21st Dec. 1957, Vol. 150, No. 1, pp. 10–15.)

621.385.833 1331

Method of Obtaining Images in Electron Diffraction and Microscopy without Photographic Emulsion.—J. J. Trillat & L. Tertian. (*C. R. Acad. Sci., Paris*, 4th Aug. 1958, Vol. 247, No. 5, pp. 582–584.) The photographic plate in an electron microscope or diffractograph is replaced by a sheet of insulating material which is treated with a fine positively charged powder to reveal the image.

621.387.4 1332

The Spreading of the Discharge in Self-Quenching Counter Tubes: Part 2.—E. Huster & E. Ziegler. (*Z. Phys.*, 25th Nov. 1957, Vol. 149, No. 5, pp. 583–593.) Experimental results are interpreted on the basis of the theory suggested in Part 1 (1561 of 1957).

PROPAGATION OF WAVES

621.396.11 + 621.396.67 1333

International Colloquium on Current Problems in Radio Wave Propagation, Paris, 17th–21st Sept. 1956.—(*Ann. Télécommun.*, May 1957, Vol. 12, No. 5, pp. 140–216.)

Ionospheric Propagation Problems:

(*a*) Theory of the Interaction between Two Waves in an Ionized Gas.—M. Bayet, J. L. Delcroix & J. F. Denisse (pp. 140–141).

(*b*) Graphical Method for the Routine Determination of the Vertical Distribution of Electron Density in Ionospheric Layers from Frequency-Sweep Recordings.—W. Becker (pp. 141–145). See 115 of 1957.

(*c*) Study of Ionospheric Propagation by Hyperfine Recording of the Field of Standard Transmitters.—J. Bouchard (pp. 146–150).

(*d*) The Self-Demodulation of Waves in the Ionosphere.—M. Cutolo (pp. 150–155).

(*e*) Pulse Propagation Experiments at Oblique Incidence.—W. Dieminger (pp. 155–159).

(*f*) Ionospheric Wind Observations.—F. Harnischmacher (pp. 159–161).

(*g*) The Transmission of Solar Radio Bursts.—C. de Jager (pp. 161–164).

(*h*) A Proposal for Improving the Accuracy of Ionospheric Forecasts.—C. M. Minnis (pp. 164–168).

(*i*) Some Current Problems in Ionospheric Forecasting.—K. Rawer (pp. 169–172).

(*j*) The Influence of the Auroral Zone on Radio Communications.—J. Rybner & E. Ungstrup (pp. 172–173).

(*k*) On the Apparent Velocity of Short Waves.—A. Stoyko (pp. 173–174).

(*l*) Influence of Propagation on Radio-navigation Systems based on Phase Measurement.—E. Vassy (pp. 175–176).

Radio Meteorology:

(*m*) Influence of Meteorological Factors on the Propagation of Ultra High Frequencies.—M. Anastassiades & L. Carapiperis (pp. 177–180).

(*n*) The Refractive Index of Humid Air for Microwaves.—A. Battaglia, G. Boudouris & A. Gozzini (pp. 181–184).

(*o*) The Reflection of V.H.F. Waves by Sharp Variations of Humidity in the Troposphere.—P. Beckmann (pp. 184–186).

(*p*) Study of the Interaction of Two Adjacent Spheres Placed in an Electromagnetic Field.—J. Mevel (pp. 186–188).

(*q*) Influence of Frontal Discontinuities on the Propagation of Decimetric and Centimetric Waves.—P. Misme (pp. 189–194).

Influence of Irregular Surfaces; Aerial Problems:

(*r*) New Types of Aerial for Long-Range Radio Links.—G. Broussaud (pp. 195–197).

(*s*) Optimum Apertures of Aerials for Randomly Distributed Fields.—W. C. Hoffman (pp. 198–200).

(*t*) Measurements of Gain across Mountainous Obstacles in California (U.S.A.).—R. E. Lacy (pp. 200–204).

(*u*) The Centipede Aerial.—D. K. Reynolds, J. Lignon & P. A. Szente (pp. 205–210).

(*v*) The Reflection of a Plane Electromagnetic Wave by a Perfectly Conducting Rough Surface.—J. P. Schouten & A. T. De Hoop (pp. 211–214).

(*w*) Calculation of Reflection Coefficients and Differential Scattering Cross-Sections for Irregular Surfaces.—V. Twersky (pp. 214–216).

621.396.11 1334

The Sommerfeld Ground Wave.—B. Kockel. (*Ann. Phys., Lpz.*, 1958, Vol. 1, Nos. 1–3, pp. 145–156.) A discussion of Sommerfeld's treatment of surface-wave propagation.

621.396.11 **1335**

Some Observations on Scattering by Turbulent Inhomogeneities.—M. Balser. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. AP-5, No. 4, pp. 383–390.)

621.396.11 : 551.510.52 **1336**

Propagation at Great Heights in the Atmosphere.—G. Millington. (*Marconi Rev.*, 4th Quarter 1958, Vol. 21, No. 131, pp. 143–160.) A theoretical study is made of tropospheric propagation through an atmosphere which is standard at small heights, but in which the refractive index approaches unity asymptotically at great heights. This leads, in the geometrical-optics region, to a reduced horizon distance, and in the diffraction region, to a reduction in height-gain.

621.396.11 : 551.510.52 **1337**

Diurnal Influences in Tropospheric Propagation.—M. W. Gough. (*Marconi Rev.*, 4th Quarter 1958, Vol. 21, No. 131, pp. 198–212.) Signal-strength recordings at 80 Mc/s combined with meteorological soundings, have been maintained for six months over a 137-km non-optical path close to the Persian Gulf. Results show that there are three main types of propagation. Examples are given of the association between high signal levels and surface ducts, particularly of the intense nocturnal type, most of which are capable of trapping the first transmission mode.

621.396.11 : 551.510.535 **1338**

The Propagation of Electromagnetic TE(H) Waves Generated by a Horizontal Current-Carrying Conductor Loop in the Cavity between Two Concentric Spheres.—O. Rösner. (*Z. angew. Phys.*, Sept. 1957, Vol. 9, No. 9, pp. 448–453.) The system earth-air-ionosphere is considered.

621.396.11 : 551.510.535 **1339**

Ionospheric Propagation: the C.C.I.R. at Warsaw 1956 and the Method of Professor Gea.—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, Sept. 1958, Vol. 12, No. 53, pp. 13–20.) A further comparison of I.F.R.B. predictions with those made by Gea's method. See also 3952 of 1958.

621.396.11 : 551.510.535 **1340**

The Determination of Skip Distances by Back-Scatter Sounding.—I. Ranzi & A. Porreca. (*Note Recensioni Notiz.*, May/June 1958, Vol. 7, No. 3, pp. 294–299. English translation, pp. 300–305.) Back-scatter minimum equivalent paths for 22·3 Mc/s are compared with vertical-sounding data obtained about half-way along the path at the time when reception of the 21·64-Mc/s transmission from Daventry fails in Rome. Skip distances determined by back-scatter sounding agree with observations; those deduced from vertical soundings appear to be excessive. See also 1197 above.

621.396.11 : 551.510.535 **1341**

Determination of H.F. Sky-Wave Absorption.—G. L. Pucillo. (*Trans. Inst. Radio Engrs*, July 1957, Vol. EP-5, No. 3, pp. 314–315.) A graphical method is described.

621.396.11 : 551.510.535 **1342**

Characteristics of F_2-Layer Multiple Reflections: Part 2.—Y. Nomura, S. Katano, Y. Echizenya, R. Nishizaki, K. Ishizawa, T. Mori & T. Kokaku. (*J. Radio Res. Labs, Japan*, Oct. 1958, Vol. 5, No. 22, pp. 295–302.) Focusing effects due to the curvature of the ionosphere are considered to be the cause of multiple reflections. For Part 1 see 894 of 1956 (Echizenya et al.).

621.396.11 : 551.510.535 **1343**

Round - the - World Echoes.—G. A. Isted. (*Marconi Rev.*, 4th Quarter 1958, Vol. 21, No. 131, pp. 173–183.) Information on round-the-world echoes, including those attributed to transmissions from satellite 1957α, is reviewed. Investigations show that circulating signals can be propagated through areas of low ionization density, and it is suggested that the mechanism is one in which transmission is made possible by launching a wave at an extremely small angle of incidence against a tilted layer, to produce a succession of reflections which are confined solely to the ionosphere.

621.396.11 : 551.510.535 : 523.5 **1344**

Meteor Activity as a Factor in Ionospheric Scatter Propagation.—G. A. Isted. (*Marconi Rev.*, 4th Quarter 1958, Vol. 21, No. 131, pp. 161–172.) An analysis has been made of the results of experimental work on transient echoes and of the characteristics of the received signal. Weather-cloud discharges or some other mechanism may prove to have a greater influence on forward scatter propagation than meteor activity.

621.396.11 : 551.510.535 : 629.19 **1345**

Measurement of the Scattering Matrix with an Intervening Ionosphere.—H. Brysk. (*Commun. & Electronics*, Nov. 1958, No. 39, pp. 611–612.) An expression for the received field is derived for radar scattering from earth satellites when Faraday rotation occurs.

621.396.11.029.63 **1346**

Propagation Measurements at 858 Mc/s over Paths up to 585 km.—G. C. Rider. (*Marconi Rev.*, 4th Quarter 1958, Vol. 21, No. 131, pp. 184–197.) Results are given of measurements of signal strength and fading parameters which have been accompanied by polarization, height-gain and space-diversity tests, and an investigation of multipath time delays. These results give an attenuation ratio of 0·103 dB/km, show no dependence of fading-rate upon distance and diversity distances larger than expected.

621.396.11.029.64 : 621.396.677 **1347**

Some Observations of Antenna-Beam Distortion in Trans-horizon Propagation.—A. T. Waterman, Jr, N. H. Bryant & R. E. Miller. (*Trans. Inst. Radio Engrs*, July 1957, Vol. AP-5, No. 3, pp. 260–266. Abstract, *Proc. Inst. Radio Engrs*, Dec. 1957, Vol. 45, No. 12, p. 1759.)

RECEPTION

621.396.62 : 621.376.3 **1348**

The R.F. Protection Ratios Required for Modern V.H.F. F.M. Receivers.—B. Gramatke, R. Netzband & E. Paulsen. (*Rundfunktech. Mitt.*, April 1958, Vol. 2, No. 2, pp. 41–53.) Subjective and objective methods of measuring interference on commercial receivers are described. Measurements were made for transmitters modulated with different programs and for transmitters radiating the same program. The reduction of protection ratios for shared-channel operation is discussed.

621.396.621 : 621.314.7 **1349**

Special Circuits for Transistor Receivers.—W. E. Sheehan & W. H. Ryer. (*Electronics*, 9th Jan. 1959, Vol. 32, No. 2, pp. 56–57.) Design data on four portable transistor receiver circuits are summarized in tabular form.

621.396.621 : 621.376.3 **1350**

Dynamic Trap Captures Weak F.M. Signals.—E. J. Baghdady & G. J. Rubissow. (*Electronics*, 9th Jan. 1959, Vol. 32, No. 2, pp. 64–66.) The use is described of a high-Q trap in a reactance-valve circuit for the suppression of a stronger undesired signal.

621.396.81.029.62 **1351**

220-Mc/s Radio-Wave Reception at 700–1 000 Miles.—L. A. Ames & T. F. Rogers. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, p. 86.) Fields measured on the ground and in the air were higher than would be expected from theories of tropospheric scatter, and showed rapid fading. They are attributed to ionospheric scatter, and reasons are given for the increase in field strength above that estimated by extrapolation from lower-frequency measurements.

621.396.812.3 **1352**

Theoretical Considerations of A1-Telegraphy Reception Disturbed by Fading.—B. Betzenhammer. (*Arch. elekt. Übertragung*, Feb. 1958, Vol. 12, No. 2, pp. 81–90.)

621.396.82.029.64 : 621.396.96 **1353**

Radar Interference to Microwave Communication Services.—R. D. Campbell. (*Commun. & Electronics*, Nov. 1958, No. 39, pp. 717–722.) The radiated frequency spectrum of high-power radars has been studied and the effect of these emissions on microwave systems has been determined. Suggestions are made for avoiding interference, including the suppression of spurious radar frequencies.

621.396.821 **1354**

Average Power of Impulsive Atmospheric Radio Noise.—S. V. C. Aiya. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, p. 92.)

STATIONS AND COMMUNICATION SYSTEMS

621.39 1355

Comprehensive Comparisons in the Planning of Telecommunication Systems.—R. Krzyczkewski. (*A.T.E. J.*, April 1958, Vol. 14, No. 2, pp. 120–128.) A method is described for evaluating, by weighting, the relative merits of radio, coaxial-cable and balanced-pair carrier cable systems when applied to a hypothetical trunk telecommunications problem.

621.391 1356

Sampling of Signals without D.C. Components.—A. R. Billings. (*Electronic Radio Engr*, Feb. 1959, Vol. 36, No. 2, pp. 70–72.) A signal contained in a finite frequency band can only be sampled unambiguously at frequencies in certain permitted bands. The Shannon-Hartley law restricting permissible sampling frequencies is generalized.

621.391 1357

Maximum Number of Signals of Fixed Total Energy among which a Discrimination to within ϵ can be made in the Presence of White Gaussian Noise.—P. Béthoux. (*C. R. Acad. Sci., Paris*, 4th Aug. 1958, Vol. 247, No. 5, pp. 573–575.) A note on the determination of $n(E, \epsilon)$, the maximum number of different signals of energy $\leqslant$ E in a limited spectrum which it is possible to distinguish with an error probability $\leqslant \epsilon$.

621.391 : 621.396.3 1358

Notes on Error-Correcting Techniques: Part 1—Efficiency of Single-Error-Correcting Codes with a Constant Bit Rate of Transmission.—J. Dutka. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 628–641.)

621.391 : 621.396.8 1359

On the 'Minimum Loss Operation Time' for Short-Wave Communication: Part 2.—H. Shibata. (*J. Radio Res. Labs, Japan*, Oct. 1958, Vol. 5, No. 22, pp. 335–340.) This concept is not only a useful one for prediction purposes, but it also allows the integration of all possible modes of propagation. The importance of the E_s layer is stressed. Part 1 : 258 of January.

621.396.41 : 551.510.52(083.57) 1360

Tropospheric-Scatter System Design.—L. P. Yeh. (*Commun. & Electronics*, Nov. 1958, No. 39, pp. 707–716.) Generalized curves have been obtained from experimental and theoretical data for channel signal/noise ratio, scatter loss, terrain effects and fading. Calculations show that for a 200-mile path a s.s.b. system will be more reliable than a f.m. system. See also 1552 of 1958.

621.396.41 : 621.396.65 1361

Design of Tropospheric-Scatter Multichannel Telephone Links.—B. Peroni & G. Bianconi. (*Note Recensioni Notiz.*, March/April 1958, Vol. 7, No. 2, pp. 147–180.) A method of calculating the system parameters for f.m. radio links is given with design curves.

621.396.41 : 621.396.65 1362

New Techniques Allow Wider Use of Radio Carrier in Telephone Systems.—E. V. Hird. (*Canad. Electronics Engng*, Aug. 1958, Vol. 2, o. 8, pp. 28–33.) Methods of high-frequency bridging, branching and interconnecting of multichannel R/T systems are described.

621.396.5 : 534.76 1363

A Compatible System of Stereo Transmission by F.M. Multiplex.—M. G. Crosby. (*J. audio Engng Soc.*, April 1958, Vol. 6, No. 2, pp. 70–73. Discussion.) A sum-and-difference technique is described in which an additive combination of the microphone outputs is transmitted on the main channel and a subtractive combination on the subcarrier channel. Advantages include a balanced program for the monaural listener and equal signal/noise ratios for the stereophonic system, with an effective gain for the subcarrier channel of 6 dB.

621.396.65 1364

New Aspects in the Planning of Radio Links.—G. Megla. (*Nachr Tech.*, Feb. 1958, Vol. 8, No. 2, pp. 50–55.) Choice of route and range, and the use of special aerials are discussed on the basis of recent experimental findings.

SUBSIDIARY APPARATUS

621.3.078 : 621.318.3 1365

Electronic Magnetic-Field Stabilizer with Feedback.—C. Fric. (*C. R. Acad. Sci., Paris*, 30th June 1958, Vol. 246, No. 26, pp. 3602–3605.) Fluctuations in the magnetizing current of an electromagnet are integrated using a vibrator-amplifier in a circuit with a time constant of 10 ms, and the derived signal energizes a compensating coil.

621.311.69 : 537.311.33 : 535.215 1366

The Effect of Radiation on Silicon Solar-Energy Converters.—J. J. Loferski & P. Rappaport. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 536–554.) The solar cells are unaffected by irradiation with ultraviolet light and X rays. High-energy electrons, protons and α-particles produce a time decay in such characteristics as maximum power output, open-circuit voltage and short-circuit current. The lifetime of the cells, to 75% of the initial power output, is estimated to be 10^5 years when exposed to the radiations assumed to exist at earth-satellite orbit heights.

621.314.63 : 621.3.014.36 1367

Forward Current Surge Failure in Semiconductor Rectifiers.—F. E. Gentry. (*Commun. & Electronics*, Nov. 1958, No. 39, pp. 746–750. Discussion, pp. 754–755.)

621.316.721.078 : 621.314.7 1368

A Semiconductor Current Limiter.—R. M. Warner, Jr, W. H. Jackson, E. I. Doucette & H. A. Stone, Jr. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 44–56.) A constant-current device is described which uses the field effect at a junction. The reverse bias applied controls the conductance of a thin sheet of semiconductor. Various models and a particular design are discussed and some applications to switching and waveform generation are given.

621.316.721.078 : 621.385.3 1369

Compensation for Fluctuations in the Anode Current of a Valve due to Heating.—C. Curie & Y. Descamps. (*C. R. Acad. Sci., Paris*, 21st July 1958, Vol. 247, No. 3, pp. 278–280.) In the method described, an antiphase rectified voltage proportional to heater current variations is applied to the control grid of the valve to stabilize the anode current.

TELEVISION AND PHOTOTELEGRAPHY

621.397.331.2 : 771.35 1370

Zoom Lenses for Closed-Circuit Television.—F. G. Back. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1958, Vol. 67, No. 9, pp. 598–600. Discussion.) The use of zoom lenses is described and the optical and mechanical properties of various lenses are compared.

621.397.5 : 535.623 : 778.5 1371

Colour TV Recording on Black and White Lenticular Film.—J. M. Brumbaugh, E. D. Goodale & R. D. Kell. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 71–75. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 383.) See also 2562 of 1958 (Kingslake).

621.397.6 : 535.623 1372

Colour Television Experiments.—N. Mayer. (*Rundfunktech. Mitt.*, April 1958, Vol. 2, No. 2, pp. 75–85.) Test equipment for the N.T.S.C. system is described.

621.397.6 : 621.317.799 1373

A Television Waveform Generator using Transistors.—F. Rozner. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 8–12.) A small light-weight instrument designed for use with transportable television broadcasting equipment is described. Four outputs are provided: composite synchronizing waveform, line drive, field drive and composite blanking waveform.

621.397.61 : 535.88 1374

The Television Transmission of Non-transparent Still Pictures by means of the Flying-Spot Scanning System (Television Episcope).—R. Theile & F. Pilz. (*Rundfunktech. Mitt.*, April 1958, Vol. 2, No. 2, pp. 54–63.) An experimental episcope is described which has facilities for variable magnification and automatic focusing. Suggestions are made for the design of a television epidiascope.

621.397.611.2 : 771.35 **1375**

Vidicon Camera Lenses.—G. H. Cook. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1958, Vol. 67, No. 9, pp. 596–598.) The suitability of standard cinematographic lenses is discussed and a new range of lenses is described.

621.397.611.2 : 771.35 **1376**

A New Series of Lenses for Vidicon-Type Cameras.—J. D. Hayes. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1958, Vol. 67, No. 9, pp. 593–595.)

621.397.62 **1377**

The Importance of the Video Receiver in Black-Level Transmission in Television.—H. Grosskopf. (*Rundfunktech. Mitt.*, April 1958, Vol. 2, No. 2, pp. 64–74.) The importance of the picture monitor in controlling the black level in transmission is outlined and the need for standardization of characteristics and adjustments is stressed. Improvements in black-level reproduction are suggested taking account of the subjective assessment of picture quality.

621.397.62 **1378**

Synchronous and Exalted - Carrier Detection in Television Receivers.—J. Avins, T. Brady & F. Smith. (*Trans. Inst. Radio Engrs*, Feb. 1958, Vol. BTR-4, No. 1, pp. 15–23. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 801.)

621.397.62 : 535.623 **1379**

Magnetic Demodulators for Colour TV.—M. Cooperman. (*Electronics*, 2nd Jan. 1959, Vol. 32, No. 1, pp. 56–58.) The basic design of a magnetic demodulator is given and its performance is discussed.

621.397.62 : 535.623 : 621.317.799 **1380**

A One-Tube Crystal-Filter Reference Generator for Colour TV Receivers.—R. H. Rausch & T. T. True. (*Trans. Inst. Radio Engrs*, June 1957, Vol. BTR-3, No. 1, pp. 2–7.)

621.397.62 : 621.373.52 **1381**

Transistorized Television Vertical-Deflection System.—W. F. Palmer & G. Schiess. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 98–105. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, pp. 383–384.)

621.397.62 : 621.374.33 **1382**

A New Noise-Gated A.G.C. and Sync System for TV Receivers.—(*Trans. Inst. Radio Engrs*, June 1957, Vol. BTR-3, No. 1.)

Part 1—J. G. Spracklen & W. Stroh (pp. 28–31).

Part 2—G. C. Wood (pp. 32–34).

See 2928 of 1957.

621.397.62 : 621.375.1 **1383**

Design Considerations of a Developmental U.H.F. Tuner using an R.F. Amplifier.—J. B. Quirk. (*Trans. Inst. Radio Engrs*, Feb. 1958, Vol. BTR-4, No. 1, pp. 5–11.) Investigation of the performance of a ceramic triode incorporated in an existing u.h.f. tuner.

621.397.62 : 621.375.1 **1384**

A Constant-Input-Impedance R.F. Amplifier for V.H.F. Television Receivers.—H. B. Yin & H. M. Wasson. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 65–70. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 383.)

621.397.62 : 621.375.4 **1385**

Transistor Design for Picture I.F. Stages.—R. J. Turner & P. Hermann. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 76–80. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 383.)

621.397.621 : 535.623 : 621.385.832 **1386**

Techniques of Colour Purity Adjustments in Receivers employing the 'Apple' Cathode-Ray Tube.—R. C. Moore, A. Hopengarten & P. G. Wolfe. (*Trans. Inst. Radio Engrs*, June 1957, Vol. BTR-3, No. 1, pp. 23–27.)

621.397.621 : 621.318.4 : 621.373.43 **1387**

A Line Transformer with Tuned High-Voltage Winding.—H. Reker. (*Nachrichtentech. Z.*, March 1958, Vol. 11, No. 3, pp. 147–153.) A method of preventing parasitic oscillation in line-deflection transformers is described. See 3673 of 1957 (Beauchamp).

621.397.621 : 621.397.82 **1388**

The Synchronization Characteristics of Television Receivers in the Presence of Interference.—E. Lüdicke. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 8–14.) Methods of reducing the susceptibility to noise of the sweep control systems in television receivers are described.

621.397.621.2 : 621.385.032.263 **1389**

Drive Factor and Gamma of Conventional Kinescope Guns.—Gold & Schwartz. (See 1417.)

621.397.621.2.001.4 **1390**

A Method of Measuring the Optical Sine-Wave Spatial Spectrum of Television Image Display Devices.—O. H. Schade. (*J. Soc. Mot. Pict. Telev. Engrs*, Sept. 1958, Vol. 67, No. 9, pp. 561–566.) The resolution characteristic of kinescopes has been measured with electrically generated sine-wave patterns giving consistent results with varying beam current. Measurements are independent of phosphor decay time.

621.397.8 : 535.623 **1391**

On the Quality of Colour-Television Images and the Perception of Colour Detail.—O. H. Schade, Sr. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 495–535.) The contrast range and colour saturation of commercial colour kinescopes in the N.T.S.C. system give a colour space larger than that of commercial colour motion pictures. More than 60% of the fine-detail colour information is transmitted and reproduced.

621.397.8 : 535.623 **1392**

Colour-Signal Distortions in Envelope Type of Second Detectors.—B. D. Loughlin. (*Trans. Inst. Radio Engrs*, Oct. 1957, Vol. BTR-3, No. 2, pp. 81–93. Abstract, *Proc. Inst. Radio Engrs*, Jan. 1958, Vol. 46, No. 1, p. 383.)

621.397.813 **1393**

Test Pattern for Assessing the Quality of Television Images and Projection Systems.—W. Kroebel. (*Naturwissenschaften*, May 1958, Vol. 45. No. 9, pp. 205–206.) The pattern consists of a series of equidistant vertical grey wedges decreasing in contrast along the abscissa, It is designed for subjective assessment o- picture quality.

621.397.9 **1394**

Industrial Television Installations.—W. Mayer. (*Frequenz*, Feb. 1958, Vol. 12, No. 2, pp. 45–49.) Outline of the basic principles of design and operation.

621.397.9 **1395**

New Applications of Industrial Tele vision.—E. F. Spiegel. (*Frequenz*, Feb. 1958, Vol. 12, No. 2, pp. 33–38.) Accessories and special equipment of German manufacture are briefly described.

TRANSMISSION

621.396.61.001.4 : 621.396.3 **1396**

I.R.E. Technical Committee Report: Methods for Testing Radiotelegraph Transmitters (below 50 Mc/s).—(*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 57–63.) Report 58 I.R.E. 15. TR1.

621.396.712.2 : 621.396.66 **1397**

A New Method of Automatic Monitoring of Radio Broadcast Channels.—F. Enkel. (*Nachrichtentech. Z.*, March 1958, Vol. 11, No. 3, pp. 142–147.) An automatic system is described for the supervision of a transmission system by monitoring the radiated signal. Measurements of level, noise and distortion are carried out during natural breaks in modulation. Tests of nonlinear distortion are made using the interval signal, and facilities for recording interference on magnetic tape are provided.

VALVES AND THERMIONICS

621.314.63 **1398**

The Influence of Adjacent Connections on the D.C. and A.C. Characteristics of *p-n* Junctions.—H. Beneking. (*Z. angew. Phys.*, Dec. 1957, Vol. 9, No. 12, pp. 626–631.) The effect of the external connections closely approaching the barrier layer is calculated; the result is in agreement with experimental findings.

621.314.7 **1399**

Transistors Reach for Higher Frequencies.—A. L. Barry & P. J. Coppen.

(*Canad. Electronics Engng*, Aug. 1958, Vol. 2, No. 8, pp. 19–27.) A review is given of progress in the development of amplifying devices for high-frequency operation.

621.314.7 : 621.317.3 **1400**
Measurement of Transistor Characteristics in the 3–250-Megacycle Frequency Range.—O'Connell & Scott. (See 1304.)

621.314.7 : 621.318.57 **1401**
The Theory of the Switching Transistor.—W. v. Münch. (*Z. angew. Phys.*, Dec. 1957, Vol. 9, No. 12, pp. 621–625.) Discussion of the characteristics of the special transistor described in 3244 of 1956 (Salow & v. Münch).

621.314.7.002.2 **1402**
Selective Electrolytic Etching of Germanium and Silicon Junction Transistor Structures.—I. A. Lesk & R. E. Gonzalez. (*J. electrochem. Soc.*, Aug. 1958, Vol. 105, No. 8, pp. 469–472.) Selective etching may be used to expose part of the base region and thereby facilitate attachment of the transistor base lead.

621.314.7.01 **1403**
On the Usefulness of Transconductance as a Transistor Parameter.—H. L. Armstrong. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 83–84.)

621.383.2 **1404**
Image Converters for Quantity Production, and the Feasibility of Defining their Emission Gain Characteristic.—P. Görlich, A. Krohs, H. J. Pohl & G. Zerbst. (*Z. angew. Phys.*, Nov. 1957, Vol. 9, No. 11, pp. 561–566.) Three types of converter, two with Cs-Sb and one with caesium oxide cathode, and their applications are described.

621.383.4 **1405**
A Simple Photoelectric Amplifier.—H. Oswald & H. Straubel. (*Z. angew. Phys.*, Sept. 1957, Vol. 9, No. 9, pp. 438–442.) A differential arrangement of two photoconducting cells in a bridge circuit is described. The two cells are formed by dividing a single CdS crystal after mounting.

621.383.4 **1406**
The Conduction Mechanism of CdS Sandwich Photocells.—G. Ecker & J. Fassbender. (*Z. Phys.*, 25th Nov. 1957, Vol. 149, No. 5, pp. 571–582.) A theory is developed based on a model of photoconductivity in semiconductors and compared with results of measurements on photocells.

621.383.4 **1407**
Behaviour of Lead Sulphide Photocells in the Ultraviolet.—A. Smith & D. Dutton. (*J. opt. Soc. Amer.*, Dec. 1958, Vol. 48, No. 12, pp. 1007–1009.) Spectral response was measured in the wavelength range $0{\cdot}2$–$2{\cdot}0\,\mu$. The increase in quantum yield observed below $0{\cdot}6\,\mu$ is attributed to electron multiplication by secondary internal emission.

621.383.4 : 535.371.07 **1408**
A Feedback Light-Amplifier Panel for Picture Storage.—B. Kazan. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 12–19.) A panel incorporating optical feedback is described in which separate phosphor layers having 40 elements per inch are used for viewing and feedback. The effect of varying the feedback factor is shown.

621.385.029.6 **1409**
Generation of Second Harmonic in a Velocity-Modulated Electron Beam of Finite Diameter.—F. Paschke. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 617–627.) The analysis, taking into account the beam fringe fields, shows that the amplitudes of the second-harmonic current and velocity do not vary periodically with distance.

621.385.029.6 **1410**
Development of High-Power Pulsed Klystrons for Practical Applications.—M. Chodorow, E. L. Ginzton, J. Jasberg, J. V. Lebacqz & H. J. Shaw. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 20–29.) The design procedure is described for sealed-off, tunable klystrons giving peak output powers of 1–2 MW and operating in the X, S and L frequency bands. The electron beam and r.f. structure are considered and cavity tuning methods appropriate to the three bands are discussed. The measured performance is shown by curves of power output, gain and efficiency.

621.385.029.6 **1411**
The Effect on Gain and Stability of a Travelling-Wave Valve of the Choice of Surface Resistance of its Attenuating Layer.—W. Eichin & G. Landauer. (*Nachrichtentech. Z.*, March 1958, Vol. 11, No. 3, pp. 131–137.) Maximum attenuation in coaxial cylindrical attenuators with normal helix dimensions is obtained with surface resistance about $2{\cdot}5\,\text{k}\Omega$. Non-coaxial arrangements have attenuation maxima corresponding to lower surface resistances.

621.385.029.6 : 537.533 : 621.375.9 **1412**
A New Method for Pumping a Fast Space-Charge Wave.—G. Wade & R. Adler. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 79–80.) A technique for low-noise amplification by interaction between a fast wave and an electron beam is illustrated by a mechanical analogy of a pendulum with periodically applied vertical forces.

621.385.029.6 : 621.376.23 **1413**
The Use of Beam Defocusing to Provide a Microwave Detector.—P. S. Castro & J. S. Needle. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 82–83.) A note on the detection action which occurs when a density-modulated or velocity-modulated electron beam is allowed to drift through a field-free space inside a cylindrical collector of resistive material. Wide-band characteristics are obtainable, with sensitivity comparable to that of a crystal diode. See also 2259 of 1956 (Mendel).

621.385.029.63 **1414**
Electrostatically Focused Travelling-Wave Tube.—D. J. Blattner & F. E. Vaccaro. (*Electronics*, 2nd Jan. 1959, Vol. 32, No. 1, pp. 46–48.) Details are given of the construction and performance of a light-weight valve for operation at frequencies between about 2 and 3 kMc/s.

621.385.029.65 **1415**
Helix-Type Travelling-Wave Amplifier for 48 kMc/s.—T. Miwa, M. Mishima & I. Yanaoka. (*Proc. Inst. Radio Engrs*, Jan. 1959, Vol. 47, No. 1, pp. 89–90.)

621.385.032.26 **1416**
The Focal Length of a Diaphragm of Finite Aperture for Electron Beams with Finite Space Charge.—K. Pöschl & W. Veith. (*Arch. elekt. Übertragung*, Jan. 1958, Vol. 12, No. 1, pp. 45–48.)

621.385.032.263 : 621.397.621.2 **1417**
Drive Factor and Gamma of Conventional Kinescope Guns.—R. D. Gold & J. W. Schwartz. (*RCA Rev.*, Dec. 1958, Vol. 19, No. 4, pp. 564–583.) A new expression for the beam-current/drive-voltage relations is derived which gives values for gamma and drive factor that agree more closely with experimental results than the values derived from previous expressions.

621.385.2.032.213.1 **1418**
Surface Structure of Saturated-Diode Filaments.—F. A. Benson & M. S. Seaman. (*Electronic Engng*, Jan. 1959, Vol. 31, No. 371, pp. 40–41.)

621.385.3.029.63 **1419**
Design and Performance of a New Low-Noise Triode for Use up to 1 000 Mc/s.—A. D. Williams & D. C. Gore. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 35–42.) A glass-base valve Type A2521 is described which is stable up to 1 kMc/s. Measured noise factor in optimum conditions increases linearly from 6 dB at 300 Mc/s to $11{\cdot}5$ dB at 900 Mc/s. Mutual conductance is 15 mA/V and amplification factor is 60. Used as an oscillator the valve has a power output of 150 mW at 1 250 Mc/s.

MISCELLANEOUS

621.3-71 : 629.13 **1420**
Cooling Airborne Electronic Equipment.—L. A. Williamson. (*Wireless World*, Feb. 1959, Vol. 65, No. 2, pp. 87–91.) Conditions under which equipment must operate in aircraft and the limitations of air-cooling are examined. Experimental chassis to facilitate cooling by liquids have been constructed using either an electroforming honeycomb process or roll-bonded ducted aluminium sheet. The choice of coolants and heat-exchangers is discussed and the advantages of the system are summarized.

621.396 **1421**
[I.E.E.] **Radio and Telecommunication Section : Chairman's Address.**—G. Millington. (*Proc. Instn elect. Engrs*, Part B, Jan. 1959, Vol. 106, No. 25, pp. 11–14.) Topics discussed include technical literature, 'rigour' in mathematics, systems of notation, the limitations of statistics, and the use of 'speculative reasoning'.

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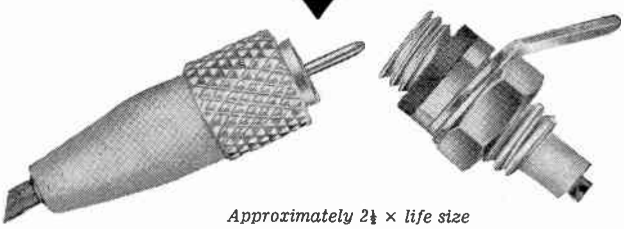

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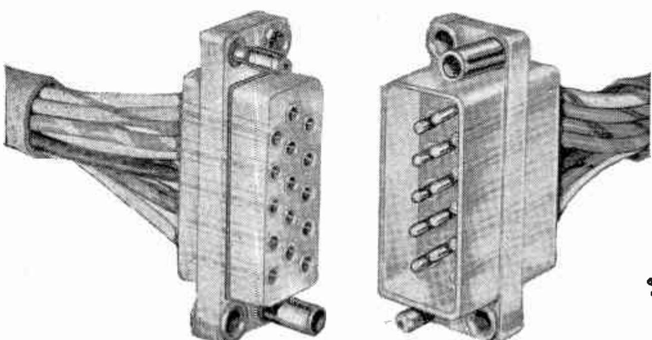

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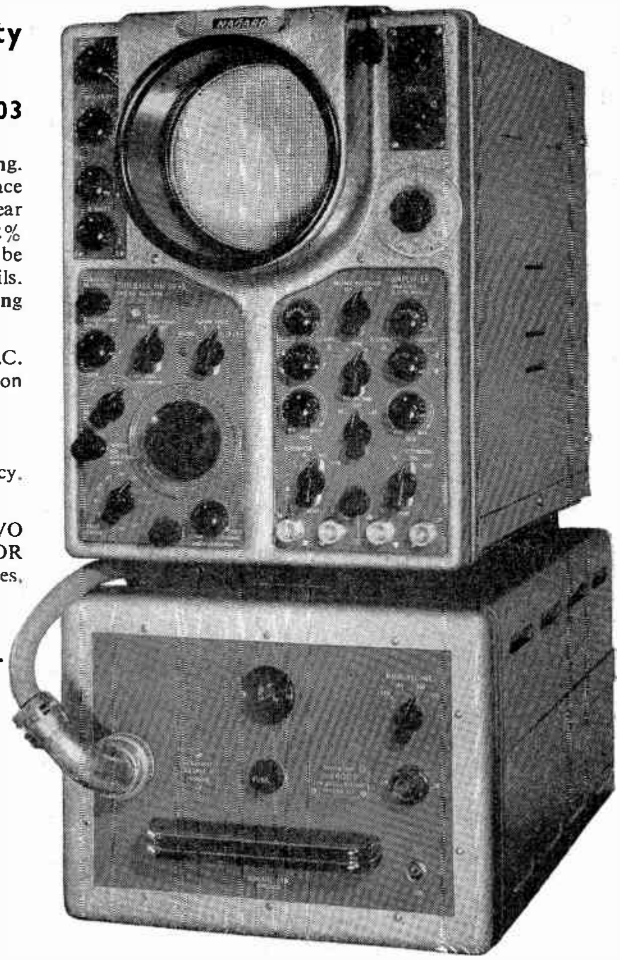

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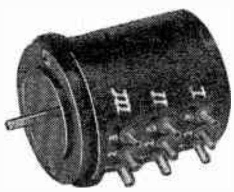

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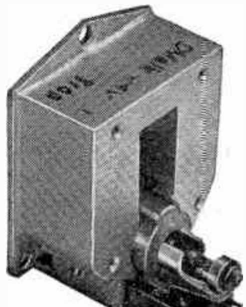

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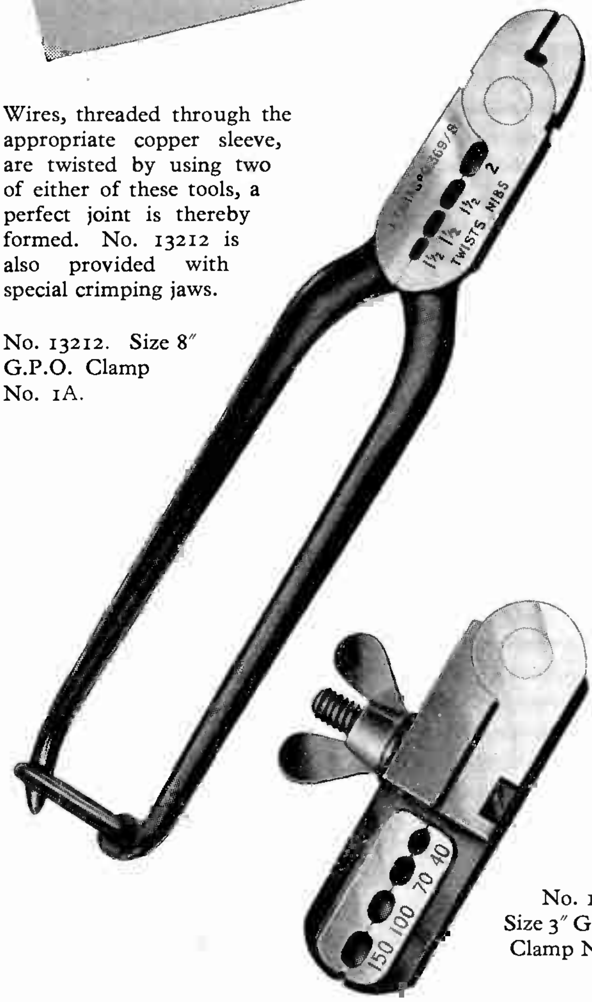

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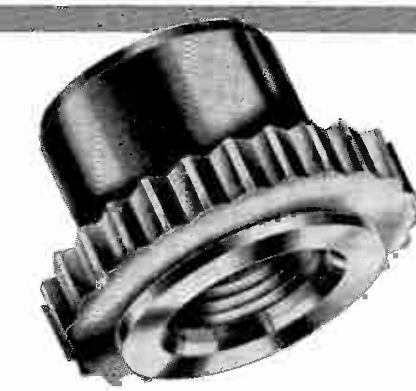

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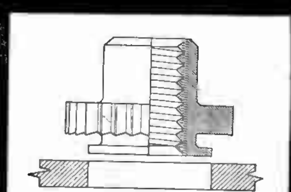

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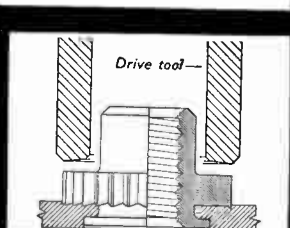

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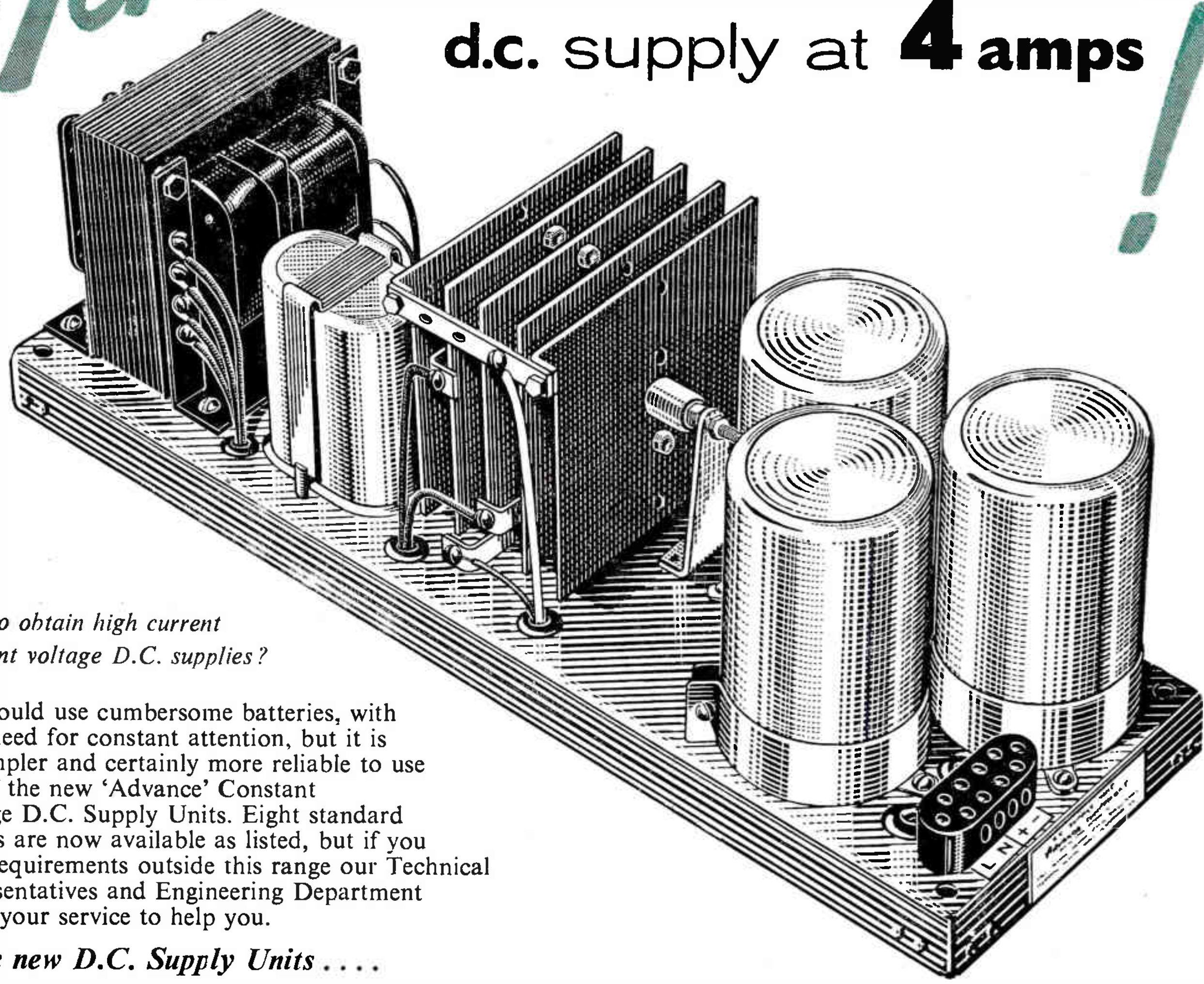

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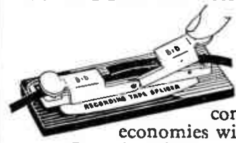

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