

THE RADIO CONSTRUCTOR

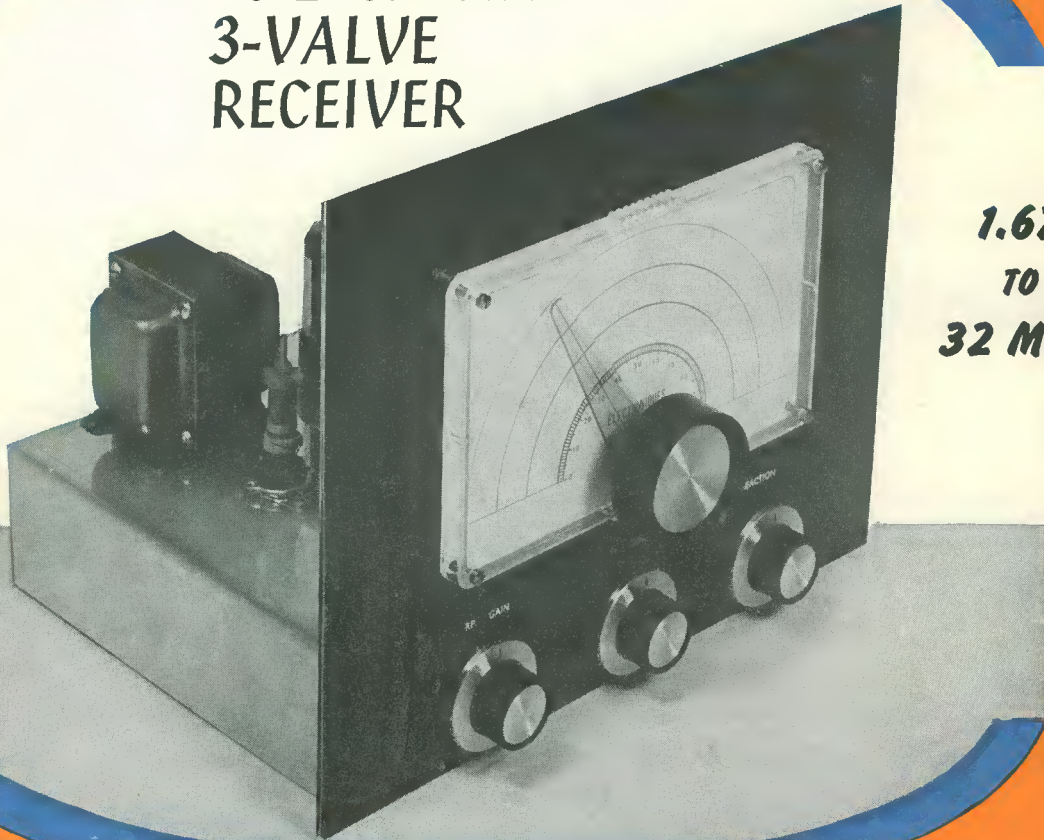
Vol 20 No 8

MARCH 1967
2/6

A DATA PUBLICATION

RADIO · TELEVISION
ELECTRONICS · AUDIO

THE "8-2 SPECIAL" 3-VALVE RECEIVER



1.67
TO
32 Mc/s

a
Short-Wave
Design

- 2-Tone Alarm Generator
- Transmitter V.F.O's
- Semiconductor Device Coding Systems
- Wide-Band Photophone

Scottish Insurance Corporation Ltd

66-67 CORNHILL · LONDON · EC3



TELEVISION
SETS,
RECEIVERS
AND
TRANSMITTERS

Television Sets, Receivers and Short Wave Transmitters are expensive to acquire and you no doubt highly prize your installation. Apart from the value of your Set, you might be held responsible should injury be caused by a fault in the Set, or injury or damage by your Aerial collapsing.

A "Scottish" special policy for Television Sets, Receivers and Short Wave Transmitters provides the following cover:

- (a) Loss or damage to installation (including in the case of Television Sets the Cathode Ray Tube) by Fire, Explosion, Lightning, Theft or Accidental External Means at any private dwelling-house.
- (b) (i) Legal Liability for bodily injury to Third Parties or damage to their property arising out of the breakage or collapse of the Aerial Fittings or Mast, or through any defect in the Set. Indemnity £10,000 any one accident.
- (ii) Damage to your property or that of your landlord arising out of the breakage or collapse of the Aerial Fittings or Mast, but not exceeding £500.

The cost of Cover (a) is 5/- a year for Sets worth £50 or less, and for Sets valued at more than £50 the cost is in proportion. Cover (b) (i) and (ii) costs only 2/6 a year if taken with Cover (a), or 5/- if taken alone.

Why not BE PRUDENT AND INSURE your installation—it is well worth while AT THE VERY LOW COST INVOLVED. If you write to the Corporation's Office a proposal will be submitted for completion.

Write for full details, quoting reference 5304, to:—

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SCOTTISH INSURANCE CORPORATION LTD.,
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Fast Mail Order for the Amateur Radio Enthusiast!

AERIAL EQUIPMENT

TWIN FEEDER. 300 ohm twin ribbon feeder similar K25, 8d. per yard. 75 ohm twin feeder, 6d. per yard. Post on above feeders and cable, 2/- any length.

COPPER WIRE. 14G, H/D, 140 ft., 30/-; 70ft., 14/-, Post and packing 3/3. Other lengths pro rata.

FEEDER SPREADERS. 6" Ceramic type F.S., 10d. each. Postage 2/6 up to 12.

CERAMIC CENTRE PIECE for dipoles, Type AT, 1/6 each. P. & P. 1/-.

2 METRE BEAM, 5 ELEMENT W.S. YAGI. Complete in box with 1" to 2 1/2" masthead bracket. Price 56/-, P. & P. 4/-.

SUPER AERAXIAL. 70/80 ohm coax. 300 watt very low loss, 2/3 per yard. 50 ohm 300 watt, 2/6 per yard. P. & P. 2/6.

TOUGH POLYTHENE LINE, type ML1 (100lb.), 2d. per yd. or 12/6 per 100 yds. Type ML2 (220lb.), 4d. per yd. or 25/- per 100 yds., ML4 (400lb.), 6d. per yd. ideal for Guys, L.V. Supports, Halyards, etc. Postage 1/6 on all line.

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- GREEN E.C.E.
- T.W. ELECTRONICS
- CODAR RADIO
- C.S.E.

NEW BOXED VALVES

3/6 each, 4 for 10/-, P. & P. 2/-.

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6N7GT	6SF7	6AB7	6SH7
6AC7	6F7	6SK7	956
6K7			

ABSORPTION WAVEMETERS, 3.00 to 35.00 Mc/s in 3 Switched Bands. 3.5, 7, 14, 21 and 28 Mc/s. Ham Bands marked on scale. Complete with indicator bulb. A MUST for any Ham Shack. ONLY 25/- EACH. Post free.

BANDCHECKER MONITOR, 3.00-35.00 Mc/s in 3 switched bands. 0-1 mA indicator. Monitor socket. Very sensitive, £3/13/6. P. & P. 3/6.

VARIABLE CONDENSERS. All brass with ceramic end plates and ball race bearings. 50pF, 5/9; 100, 6/6; 160, 7/6; 240, 8/6; and 300pF, 10/6. Extension for ganging. P. & P. 2/-.

SEALED RELAYS, 12v. 105Ω Coil. Type A. 4 Pole, C.O. 15/-, Type B.2 Pole, C.O. + 2 Pole Norm. on, 12/6. P. & P. 1/6.



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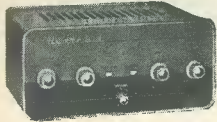
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HI-FI AMPLIFIERS ~~~~~ TUNERS ~~~~~ RECORD PLAYERS



**3+3W
STEREO
AMP.
S-33H**



**10W
POWER
AMP.
MA-12**

10W POWER AMPLIFIER. Model MA-12. 10W output, wide freq. range, low distortion. For use with control unit. Kit **£12.18.0** Assembled **£16.18.0**

3 + 3W STEREO AMPLIFIER. Model S-33. An easy-to-build, low cost unit. 2 inputs per channel. Kit **£13.7.6** Assembled **£18.18.0**

DE LUXE STEREO AMPLIFIER. Model S-33H. De luxe version of the S-33 with two-tone grey perspex panel, and higher sensitivity necessary to accept the Decca Deram pick-up. Kit **£15.17.6** Assembled **£21.7.6**

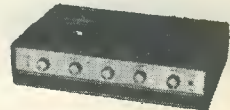
HI-FI STEREO AMPLIFIER. Model S-99. 9 + 9W output. Ganged controls. Stereo/Mono gram., radio and tape inputs. Push-button selection. Printed circuit construction. Kit **£28.9.6** Assembled **£38.9.6**

TRANSISTOR PA/GUITAR AMPLIFIER, PA-2. 20W amplifier. Four inputs. Variable tremolo. **New Low Price** Kit **£39.19.0** Assembled **£54.10.0**

50W VALVE PA/GUITAR AMP., PA-1. Kit **£54.15.0** Assembled **£74.0.0**



**GARRARD
PLAYER
AT-60**



**20+20W
STEREO
AMP.
AA-22U**

TRANSISTOR MIXER. Model TM-1. A must for the tape enthusiast. Four channels. Battery operated. Similar styling to Model AA-22U Amplifier. With cabinet. Kit **£11.16.6** Assembled **£16.17.6**

20+20W TRANSISTOR STEREO AMPLIFIER. Model AA-22U. Outstanding performance and appearance. Kit **£39.10.0** (less cabinet). Assembled **£57.10.0** Attractive walnut veneered cabinet **£2.5.0** extra.

GARRARD AUTO/RECORD PLAYER. Model AT-60. less cartridge **£13.1.7** With Decca Deram pick-up **£17.16.1** incl. P.T. Many other Garrard models available, ask for Lists.

HI-FI MONO AMPLIFIER. Model MA-5. A general purpose 5W Amplifier, with inputs for Gram., Radio. Attractive modern styling. Kit **£11.9.6** Assembled **£15.15.0**



SEE HEATHKIT AT THE GRAND HOTEL SOUTHAMPTON ROW, LONDON (concurrent with the Audio Fair)

INSTRUMENTS

3" LOW-PRICED SERVICE OSCILLOSCOPE. Model OS-2. Compact size 5" x 7 1/2" x 12" deep. Wt. only 9 1/2 lb. "Y" bandwidth 2 c/s-3 Mc/s ± 3dB. Sensitivity 100mV/cm. T/B 20 c/s-200 kc/s in four ranges, fitted mu-metal CRT Shield. Modern functional styling. Kit **£23.18.0** Assembled **£31.18.0**

5" GEN-PURPOSE OSCILLOSCOPE. Model 10-12U. An outstanding model with professional specification and styling. "Y" bandwidth 3 c/s-4.5 Mc/s ± 3dB. T/B 10 c/s-500 kc/s. Kit **£35.17.6** Assembled **£45.15.0**

DE LUXE LARGE-SCALE VALVE VOLTMETER. Model IM-13U. Circuit and specification based on the well-known model V-7A but with many worth-while refinements. 6" Ernest Turner meter. Unique gimbal bracket allows operation of instrument in many positions. Modern styling. Kit **£18.18.0** Assembled **£26.18.0**

AUDIO SIGNAL GENERATOR. Model AG-9U. 10 c/s to 100 kc/s, switch selected. Distortion less than 0.1%, 10V sine wave output metered in volts and dB's. Kit **£23.15.0** Assembled **£31.15.0**

VALVE VOLTMETER. Model V-7A. 7 voltage ranges d.c. volts to 1,500. A.c. to 1,500 r.m.s. and 4,000 peak to peak. Resistance 0.1Ω to 1,000MΩ with internal battery. D.c. input resistance 11MΩ. dB measurement, has centre-zero scale. Complete with test prods, leads and standardising battery. Kit **£13.18.6** Assembled **£19.18.6**

MULTIMETER. Model MM-1U. Ranges 0-1.5V to 1,500V a.c. and d.c.; 150μA to 15A d.c.; 0.2Ω to 20MΩ. 4 1/2" 50μA meter. Kit **£12.18.0** Assembled **£18.11.6**

R.F. SIGNAL GENERATOR. Model RF-1U. Up to 100 Mc/s fundamental and 200 Mc/s on harmonics. Up to 100mV output. Kit **£13.18.0** Assembled **£20.8.0**

SINE/SQUARE GENERATOR. Model IG-82U. Freq. range 20 c/s-1 Mc/s in 5 bands less than 0.5% sine wave dist. less than 0.15μ sec. sq. wave rise time. Kit **£25.15.0** Assembled **£37.15.0**

TRANSISTOR POWER SUPPLY. Model IP-20U. Up to 50V, 1.5A output. Ideal for Laboratory use. Compact size. Kit **£35.8.0** Assembled **£47.8.0**



OS-2



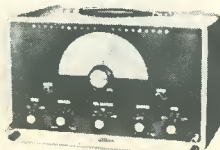
IM-13U



V-7A



RF-1U



IG-82U

**MO Prices quoted
Retail Prices in
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Complete your motoring pleasure with this outstanding CAR RADIO, Model CR-1



Will give you superb LW and MW entertainment wherever you drive. Tastefully styled to harmonise with any car colour scheme. Available for your convenience in two separate parts, RF Amp. Kit **£1.13.6** incl. P.T. IF/AF Amp. Kit **£11.3.6**

Total Price Kit (excl. Loudspeaker) £12.17.0 incl. P.T.
8" x 5" Loudspeaker **£1.16.1** incl. P.T.

TRANSISTOR RADIOS

"OXFORD" LUXURY PORTABLE. Model UXR-2. Specially designed for use as a domestic, car or personal portable receiver. Many features, including solid leather case. Kit **£14.18.0** incl. P.T.



UXR-2

TRANSISTOR PORTABLE. Model UXR-1. Pre-aligned I.F. transformers, printed circuit. Covers L.W. and M.W. Has 7" x 4" loudspeaker. Real hide case.



UXR-1

Kit **£12.11.0** incl. P.T.
JUNIOR EXPERIMENTAL WORKSHOP Model EW-1. More than a toy! Will make over 20 exciting electronic devices, incl: Radios, Burglar Alarms, etc. 72 page Manual. The ideal present! Kit **£7.13.6** incl. P.T.

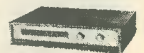
STEREO TRANSISTOR FM TUNER.
(Mono version also available)

Designed to match the AA-22U Amplifier. Available in separate units, can be built for a total price.

Kit (Stereo) **£24.18.0** incl. P.T.

Kit (Mono) **£20.19.0** incl. P.T.

Cabinet extra **£2.5.0**.



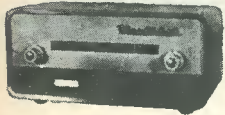
FM Tuner

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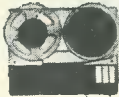
We open **MONDAY-SATURDAY** 9 a.m. - 5.30 p.m.
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WHEN YOU ARE IN TOWN, WE HOPE YOU WILL VISIT US THERE

TAPE AMPLIFIERS TAPE DECKS CONTROL UNITS



**FM
TUNER
FM-4U**



**MAGNAVOX
DECK**

HI-FI FM TUNER. Model FM-4U. Available in two units. R.F. tuning unit (£2,15.0 incl. P.T.) with I.F. output of 10.7 Mc/s, and I.F. amplifier unit, with power supply and valves (£13.13.0). May be used free standing or in a cabinet. Total Kit **£16.8.0**
(Multiplex adapter available, as extra.)

HI-FI AM/FM TUNER. Model AFM-1. Available in two units which, for your convenience, are sold separately. Tuning heart (AFM-T1—£4.13.6 incl. P.T.) and I.F. amplifier (AFM-A1—£22.11.6). Printed circuit board, 8 valves. Covers L.W., M.W., S.W., and F.M. Built-in power supply. Total Kit **£27.5.0**
(Multiplex adapter available, as extra.)

MAGNAVOX "363" TAPE DECK. The finest buy in its price range. Operating speeds: 1½", 3¼" and 7½" p.s. Two tracks, "wow" and "flutter" not greater than 0.15% at 7½" p.s. **£13.10.0**
MAGNAVOX deck with TA-IM Kit **£31.5.6**



**TRUVOX
DECK**



**AM/FM
TUNER**

TRUVOX D-93 TAPE DECKS. High quality stereo/mono tape decks. D93/2, ½ track, **£36.15.0** D93/4, ¼ track, **£36.15.0**

TAPE RECORDING/PLAYBACK AMPLIFIER
Mono Model TA-IM kit **£19.18.0** Assembled **£28.18.0**
Stereo Model TA-IM kit **£25.10.0** Assembled **£35.18.0**

HI-FI CABINETS. A wide range available for example: Malvern Kit **£18.1.0** incl. P.T.

MONO CONTROL UNIT. Model UMC-1. Designed to work with the MA-12 or similar amplifier requiring 0.25V or less for full output. 5 inputs. Baxandall type controls. Kit **£9.2.6** Assembled **£14.2.6**

STEREO CONTROL UNIT. Model USC-1. Push-button selection, accurately matched ganged controls to ±1dB. Rumble and variable low-pass filters. Printed circuit boards. Kit **£19.19.0** Assembled **£27.5.0**

EXHIBITION, Demonstrations of Hi-Fi (Admission free) MARCH 30th-APRIL 2nd—11 a.m.—9 p.m. (8 p.m. Sunday).



SSU-1

SPEAKER SYSTEMS

HI-FI SPEAKER SYSTEM. Model SSU-1. Ducted-port bass reflex cabinet "in the white". Two speakers. Vertical or horizontal models with legs, Kit **£12.12.0**, without legs, Kit **£11.17.6** incl. P.T.

THE BERKELEY Slim-line SPEAKER SYSTEM, fully finished walnut veneered cabinet for faster construction. Special 12" bass unit and 4" mid/high frequency unit. Range 30-17,000 c/s. Size 26" x 17" x only 7½" deep. Modern attractive styling. Excellent value.

Kit **£19.10.0** Assembled **£24.0.0**

COTSWOLD SPEAKER SYSTEMS. Outstanding performance for price.

MFS: Size 36" x 16½" x 14" deep.

Kit **£25.12.0** Assembled **£33.17.0**

STANDARD: Size 26" x 23" x 14½" deep.

Kit **£25.12.0** Assembled **£33.17.0**



Berkeley

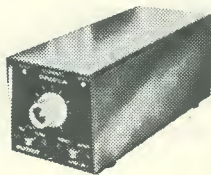
MULTIPLEX DECODER, SD-1

for receiving Stereo FM.

Convert your existing FM Mono receiver to stereo with this low-cost, self-powered unit. Fully transistorised. Styled to match Heathkit models FM-4U and AFM-1 Tuners.

Kit **£8.10.0** Assembled **£12.5.0**

Styled to match British Heathkit models.



SD-1

"AMATEUR" EQUIPMENT

THE "MOHICAN" GENERAL COVER-AGE RECEIVER. Model GC-1U. With 4 piezo-electric transmitters, variable tuned B.F.O. and Zener diode stabiliser, this is an excellent fully transistorised general purpose receiver for Amateur and Short wave listeners. Printed circuits, telescopic aerial, tuning meter and large slide-rule dial.

Kit **£37.17.6** Assembled **£45.17.6**

AMATEUR BANDS RECEIVER. Model RA-1. To cover all the Amateur Bands from 160-10 metres. Many special features, including: half-lattice crystal filter; 8 valves; signal strength "S" meter; tuned R.F. Amp. stage.

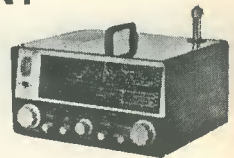
Kit **£39.6.6** Assembled **£52.10.0**

160-10M TRANSMITTER. Model DX-100U. Careful design has achieved high performance and stability. Completely self-contained.

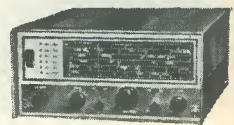
Kit **£81.10.0** Assembled **£106.15.0**

COMMUNICATIONS TYPE RECEIVER. Model RG-1. A high performance, low cost receiver for the discriminating listener. Frequency coverage: 600 kc/s-1.5 Mc/s and 1.7 Mc/s-32 Mc/s.

Kit **£39.16.0** Assembled **£53.0.0**



GC-1U



RG-1

NEW! Hi-Fi Models. 'AVON' The Mini Speaker System with Hi-Fi performance.

Only 7½" wide x 13½" high x 8½" deep. Specially designed.

Walnut veneered finished cabinet kit **£8.18.0**.

Two speaker system and crossover network **£4.18.0**.

Total price kit £13.16.0.



Low-cost 3 + 3W Transistor Stereo Amplifier, TS-23

Has all the essential features for good quality reproduction from gram, radio and other sources.

Kit (less cabinet) **£17.15.0**.

Kit (with handsome finished cabinet) **£18.19.0**.

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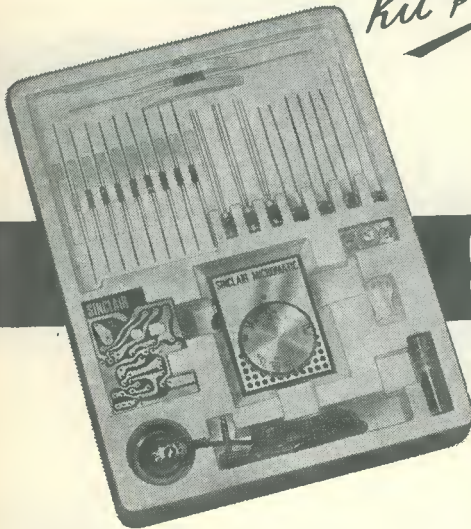
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*World's smallest radio
comes to you in the new
"See for Yourself" sealed
Kit pack.*



SINCLAIR MICROMATIC

SIX STAGE TRANSISTOR RECEIVER

Now when you buy your Sinclair Micromatic Kit, you see every component instantly in the new sealed polystyrene Kit pack. Never before in the history of Radio has any Kit been so elegantly presented. Your Micromatic reaches your factory fresh and guaranteed. Check for yourself at once that every component is in its proper place in the new case made and shaped specially to take the Micromatic kit. Your Micromatic kit

does not cost you a penny more bought this way. Such is the appearance and performance of this brilliant new Sinclair design that you will want to build and use one immediately. There is no other set in the world as small, efficient and dependable as this. Reception both of home and overseas programmes is often much easier than with larger conventional radios. In fact, your Micromatic will virtually play anywhere. It is also available ready built.

- ★ MEASURES ONLY 14¹/₅" x 13¹/₁₀" x 1¹/₂".
- ★ BEAUTIFULLY STYLED CASE
- ★ POLISHED ALUMINIUM FRONT PANEL
- ★ CALIBRATED SPUN ALUMINIUM TUNING DIAL
- ★ AMAZING POWER, RANGE AND SENSITIVITY
- ★ NEW CIRCUITRY
- ★ BANDSPREAD TUNING PLUS A.G.C.
- ★ 5 YEAR GUARANTEE

TECHNICAL DESCRIPTION

The Sinclair Micromatic is housed in a neat plastic case with attractive aluminium front panel and aluminium tuning dial to match, calibrated in metres and kc/s. Special Sinclair transistors are employed in a six-stage circuit of exceptional power and sensitivity. Two stages of powerful R.F. amplification are followed by a double diode detector from which the signal tuned in is passed to a high gain three stage audio amplifier. Automatic Gain control counteracts fading from distant stations. The set is powered by two Mallory ZM.312 Cells obtainable from radio shops, Boots Chemists, etc., for 1/7 each. They give approximately 70 hrs. working life. Inserting the earpiece plug switches the set on, withdrawing it switches off. Complete kit of parts including lightweight earpiece, instructions and solder.

SINCLAIR MICROMATIC, ready built, tested and guaranteed, with earpiece, complete in presentation case.

59/6

79/6

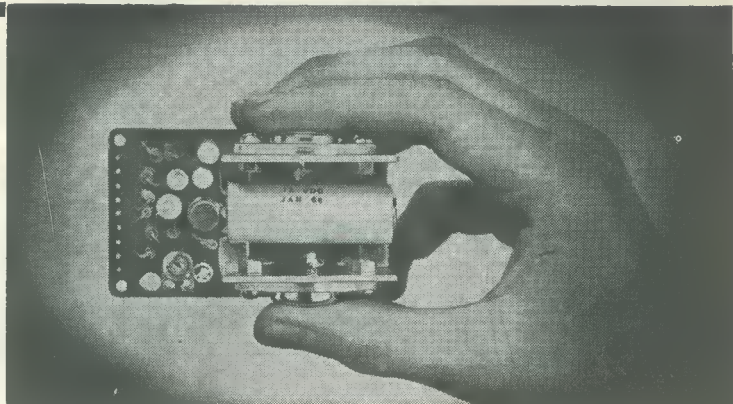
SINCLAIR RADIONICS LIMITED
22 NEWMARKET ROAD, CAMBRIDGE. Telephone OAC3-52731

SINCLAIR Z12

INTEGRATED 12 WATT
AMPLIFIER AND PRE-AMP

A MASTER OF POWER,
COMPACTNESS AND
VERSATILITY

For size alone, the Z.12 marks an important advance in quality design, for its amazing compactness opens up exciting new vistas in amplifier housing and application. Combined with this are fantastic power and superb quality which can provide an effortless output of 12 watts R.M.S. continuous sine wave from the unique eight transistor circuit used. Basically intended as the heart of any good mono or stereo hi-fi system, the size and efficiency of this Sinclair unit make it equally useful for a car radio (with the Micro-6 for example), a high quality radio with the Micro FM, in a guitar, P.A. or intercom system, etc. Other applications are certain to suggest themselves to constructors. The manual included with the Z.12 details mono and stereo tone and volume control circuits by which inputs can be matched (and switched in) to the pre-amp. The size, performance and price of the Z.12 all favour the constructor seeking the finest in transistorised audio reproduction—it is in fact today's finest buy in top grade high fidelity.



12 WATTS R.M.S. OUTPUT CONTINUOUS SINE WAVE
(24 W. PEAK)
15 WATTS R.M.S. MUSIC POWER (30 WATTS PEAK)

- ★ Ultra-linear class B output and generous neg. feedback.
- ★ Response—15 to 50,000 c/s ± 1 dB.
- ★ Output suitable for 3, 7.5 and

- 15 ohms loads. Two 3 ohm speakers may be used in parallel.
- ★ Input—2mV into 2K ohms.
- ★ Signal to noise ratio—better than 60dB.

Built, tested
and
guaranteed.

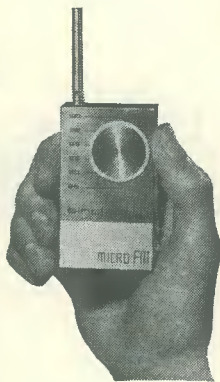
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SINCLAIR MICRO FM

Combined FM Tuner Receiver

- ★ Use it as a Tuner
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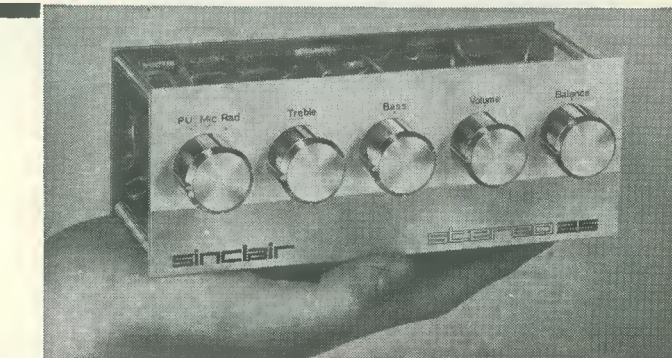
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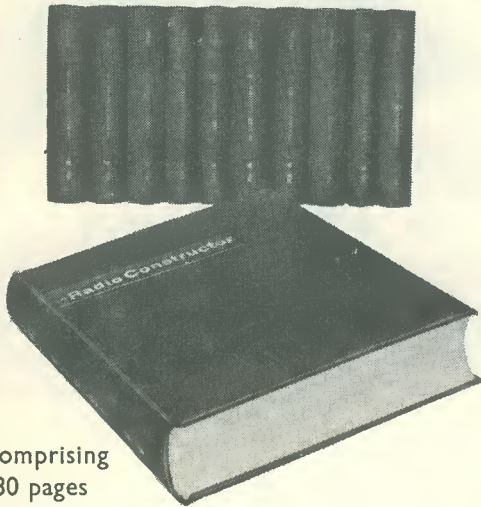
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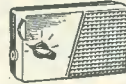
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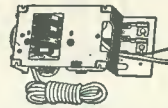
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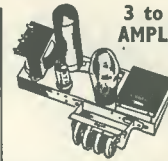
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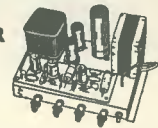
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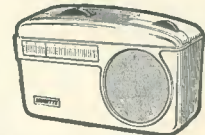
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Annual Subscription 36s. (U.S.A. and Canada \$5) including postage. Remittances should be made payable to "Data Publications Ltd". Overseas readers please pay by cheque or International Money Order.

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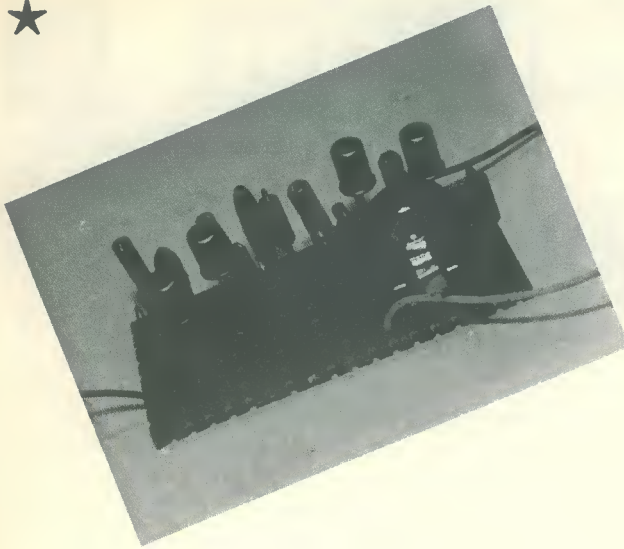
Correspondence should be addressed to the Editor, Advertising Manager, Subscription Manager or the Publishers as appropriate.

Opinions expressed by contributors are not necessarily those of the Editor or proprietors.

Production.—Letterpress/contact litho.

Published in Great Britain by the Proprietors and Publishers
Data Publications Ltd. 57 Maida Vale, London W9
Printed by A. Quick & Co. (Printers) Ltd.
Clacton-on-Sea England

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Transistorised 2-Tone Alarm Generator

by F. L. Thurston

Attention is immediately caught by the distinctive signal that this neat little assembly generates. It may be coupled to any a.f. amplifier.

THE CONVENTIONAL ALARM CALL GENERATOR, in the form of a continually sounding bell, suffers from a number of distinct disadvantages. These disadvantages do not result from any technical inadequacy in the alarm bell itself, but from the psychological characteristics of the human target at whom the alarm is aimed. To be effective

in attracting attention a constant tone or signal must be produced at considerable volume, as it is a natural characteristic of the brain to suppress all constant level signals; our hearing has, in effect, a built-in a.g.c. system. Since these alarm calls are transmitted at considerable volume they tend, if the listener is within close proximity, to attract attention in a very jarring and upsetting way. Most readers will, no doubt, have had that feeling of distinct shock as a bell goes off within a few feet of the ear when they have walked past in a mood of deep concentration!

Changing Tones

Scientific tests have shown that the brain responds

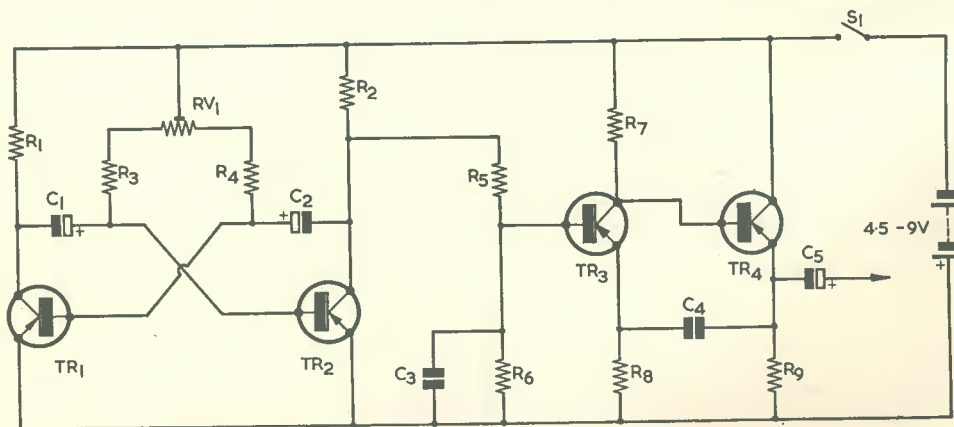


Fig. 1. The circuit of the transistorised 2-tone alarm call generator

most readily to audio tones that are continually changing either in frequency or in volume, and that, with such signals, the attention can be attracted quite easily even at very low volume levels. Such signals are thus of considerable advantage in applications where they may be used to attract attention without creating undue disturbance, i.e., in "doctor calling" services in hospitals, in the homes of old folk, etc. This high response of the brain to changing, rather than constant level signals, applies to all of the senses. It is most striking in sight where, for example, the movement of a single rabbit in a large field, or the flicker of a single bulb amongst thousands of lights, immediately attracts the attention.

The unit that forms the basis of this article is a fully transistorised 2-tone alarm call generator, and may be used to feed the call signal to either a power amplifier or radio transmitter, or some other distribution system. Amongst the applications which are of interest to the amateur, it may be used as a front door "visitor announcer", as an extension "phone call indicator", as an intercom "call" generator, or as an alarm generator for use with a rain sensor, heat or smoke indicator or burglar detector.

Circuit Operation

The full circuit diagram of the 2-tone generator is shown in Fig. 1. It is possible to build a 2-tone generator using fewer transistors than shown here, but this will usually involve the use of a transformer which, in many cases, may be difficult to obtain. The circuit of Fig. 1 calls for no specialised components, is tolerant of quite wide variations in component values, and presents no constructional problems.

The circuit can be divided into two distinct sections, TR₁-TR₂ being a low frequency pulsing or switching generator, whilst TR₃-TR₄ is the audio frequency generator.

TR₁ and TR₂ are wired as a conventional stable multivibrator. Here, one or other of the two transistors is biased hard on while the other is biased off, this condition being held for a pre-determined length of time until, suddenly, the states of the two transistors change over. This second condition is also held for a pre-determined length of time, after which the circuit states suddenly change back again, and another cycle commences.

The output of the TR₁-TR₂ circuit, taken from TR₂ collector, has the form of a rectangular waveform, the mark-space, or on-off ratio of the waveform depending on the individual time constants in the circuit. In Fig. 1, this mark-space ratio can be altered by varying the setting of RV₁. With the circuit values shown, the circuit operated in the prototype at a frequency of slightly greater than 1 c/s, but this can be increased or decreased as required, by either decreasing or increasing the values of C₁ and C₂.*

The second part of the circuit, TR₃-TR₄, is the audio oscillator section. Here, TR₃ is wired as a grounded base amplifier, with collector load R₇ and emitter bias resistor R₈, whilst TR₄ is wired as an emitter follower, with emitter load R₉. The base bias of the grounded base amplifier is determined by R₅ and R₆, the input to the transistor being applied to the emitter. Thus, if a negative-going signal is applied to the emitter the effective base-bias is reduced and the emitter (and collector) current is decreased; the voltage drop across R₇, therefore decreases also, and the collector voltage of TR₃ goes negative. The input and output signals of TR₃ are thus in phase. The signal at TR₃ collector is fed to TR₄ base, and this transistor, being an

* Due to the relatively wide tolerance on electrolytic capacitor values, it may be found that some components with the nominal values specified for C₁ and C₂ offer a slightly different switching frequency to that mentioned here.—EDITOR.

Resistors

(All fixed values $\frac{1}{4}$ watt 10%)

R ₁	5.6k Ω
R ₂	5.6k Ω
R ₃	8.2k Ω
R ₄	8.2k Ω
R ₅	33k Ω
R ₆	5.6k Ω
R ₇	4.7k Ω
R ₈	1.2k Ω
R ₉	4.7k Ω
RV ₁	250k Ω , preset, skeleton potentiometer

Capacitors

(All capacitors miniature types. C₁, C₂ and C₄ may require adjustment—see text)

C ₁	2 μ F electrolytic, 12V wkg.
C ₂	2 μ F electrolytic, 12V wkg.
C ₃	0.01 μ F
C ₄	0.1 μ F
C ₅	4 μ F electrolytic, 12V wkg.

COMPONENTS

Transistors

TR ₁	OC75
TR ₂	OC75
TR ₃	OC71
TR ₄	OC71

Switch

S ₁	s.p.s.t. switch (see text)
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Battery

4.5 to 9 volts

Miscellaneous

Veroboard, 0.15in matrix, 2 $\frac{3}{4}$ x 1 $\frac{1}{8}$ in (7 strips x 18 holes)
Wire, etc.

emitter follower, gives an output at its emitter which is also in phase with its input, but at a low impedance level. The signal at TR₄ emitter is thus in phase with, but of greater amplitude than, the signal at TR₃ emitter input. In consequence, we have the required conditions for oscillation and, when

operating frequency of the two transistors, and this bias is derived from TR₂ collector. The frequency of the TR₃-TR₄ oscillator is thus varied automatically by the switching oscillator given by TR₁ and TR₂.

The output of the unit is taken, via C₅, from the

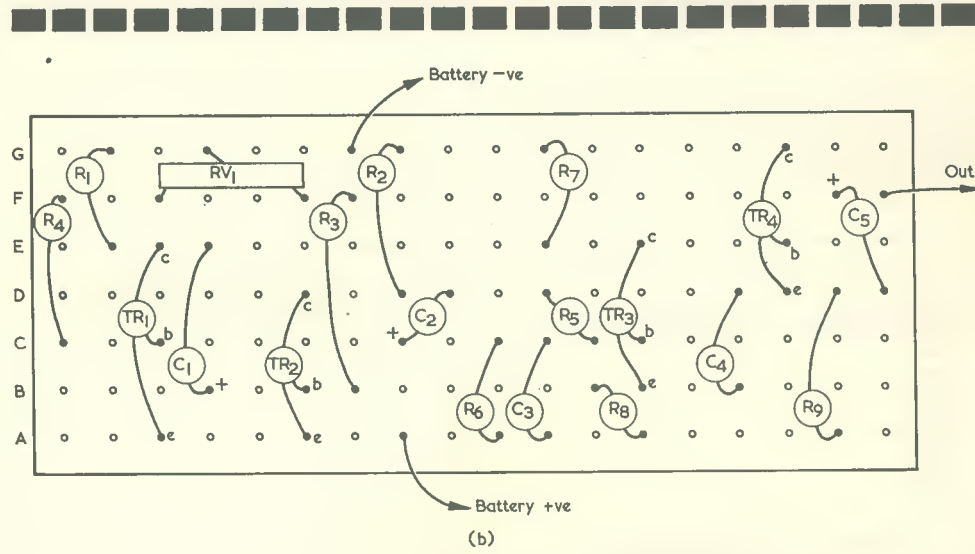
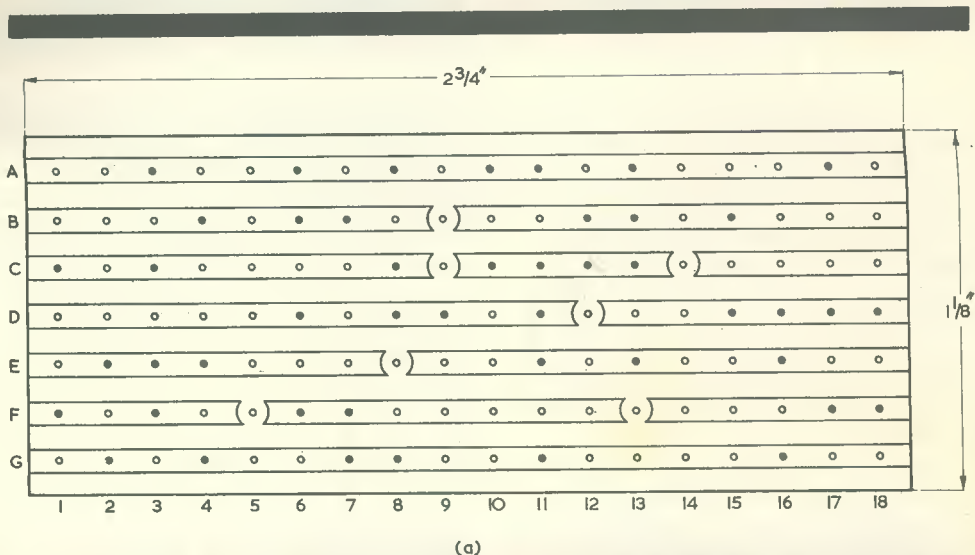


Fig. 2 (a). The copper side of the Veroboard
(b). The components fitted to the non-copper side of the board

the coupling capacitor C₄ is inserted, the circuit does in fact oscillate. The operating frequency is determined mainly by the values of R₇ and C₄. In practice, this frequency is adjusted by altering the value of C₄.

The base-bias to TR₃ also has some effect on the

emitter of TR₄. Any supply between approximately 4.5 and 9 volts may be used.

Construction

The circuit is wired up on a small piece of Vero-
(Continued on page 489)

Tetrode F.E.T. Improves A.G.C. Circuits

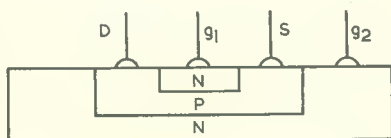
by

M. J. DARBY

IF ONE APPLIES AN A.G.C. CONTROL VOLTAGE TO a valve or transistor i.f. amplifier stage, there is a change in the effective input capacitance owing to Miller effect considerations. This inevitably leads to some detuning.

This difficulty may be avoided by the use of one of the 3N89 field effect transistors manufactured by the Siliconix Company. These components have two completely separate gate electrodes. (For a discussion of field effect transistors, see the article "Field Effect Transistors" in *The Radio Constructor*, February 1966.) In the 3N89 the two gates are physically in parallel rather than in tandem as in a multi-grid valve. As shown in Fig. 1, the second gate does not screen the first gate from the drain and application of a.g.c. to the second gate does not therefore affect the input capacitance associated with the first gate.

An i.f. amplifier circuit for 45 Mc/s using the Siliconix 3N89 is shown in Fig. 2. The gain of this amplifier may be altered over a range of 40dB in a typical case by an a.g.c. control voltage of 0 to +6 volts. The variation in the resonant frequency of the input and output tuned circuits was un-



D = drain
S = source
g₁ = first gate
g₂ = second gate

Fig. 1. Structure of the triode F.E.T.

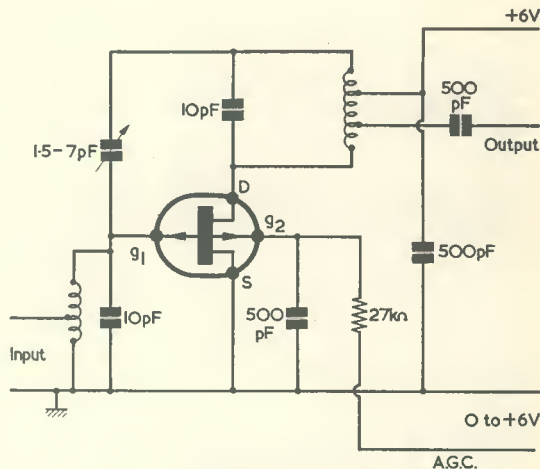


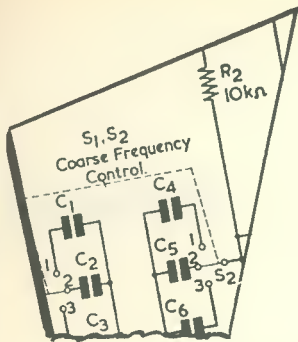
Fig. 2. A 45 Mc/s i.f. amplifier stage

detectable when measured with equipment of resolution 0.1%. If, however, the a.g.c. was applied to the first gate, the resonant frequency altered by about 300 kc/s per volt of a.g.c. bias.

Another circuit designed by the Siliconix Company employs four 3N89's in a high input impedance amplifier which provides a gain of about 60dB for a power consumption of less than one hundred microwatts (actually -3 volts at 28 microamp and +8.5 volts at one microamp). The first three tetrode f.e.t.'s of this amplifier are used as voltage amplifiers, each having a gain exceeding 10, whilst the final stage is a source follower. ○

Further information on the applications of the 3N89 is available from the manufacturers, Messrs. Siliconix Inc., 1140, West Evelyn Avenue, Sunnyvale 14, California, U.S.A.





ONE-MINUTE FLASHER FOR TRANSISTOR EQUIPMENT

SUGGESTED CIRCUIT No. 196

By G. A. FRENCH

IN THE "SUGGESTED CIRCUIT" article published last month the writer pointed out that a shortcoming with battery operated transistor equipment is that it is impracticable to fit a pilot lamp to give continuous warning that the equipment is switched on. This is due to the relatively heavy current which such a lamp would draw from the battery. At the same time, it is quite easy to forget to switch off transistor equipment, whereupon current is, once again, unnecessarily drawn from the battery.

Last month's article then carried on to discuss a simple circuit which caused a pilot lamp to flash immediately after an item of transistor equipment was switched on and to flash again immediately after it was switched off. The necessity for causing the lamp to flash when switching off would then act as a reminder to the user of the equipment. A disadvantage with the circuit was that it could not be employed with transistor equipment having high-value bypass capacitors across the supply lines.

The writer has returned to the same subject this month, but the circuit to be described offers a completely different type of visual indication to show that the equipment is switched on. In this case a pilot lamp is caused to flash approximately once per minute whenever the associated equipment is switched on. Due to the use of a high-value capacitor which discharges into the pilot lamp to cause the flash, the overall battery consumption averages at slightly less than 2.5mA only. The flashing of the lamp gives a positive indication that the equipment is switched on and, despite the fact that the flash occurs only once per minute, this should still catch the attention of the user of the equipment. The flashing circuit is connected across the battery supply lines of the equipment, immediately after the on-off switch.

Some considerable thought was given to evolving a design which would employ the minimum number of components in the flashing circuit, and the writer was eventually able to pare these down to 2 transis-

tors, a diode, 4 resistors and 2 capacitors. Of these, the most bulky component is the high-value capacitor (which would be necessary with any design following the flashing principle employed here), whilst the remaining components are all small and take up little space. Thus, the flasher circuit could be installed in many transistor radios of the larger type.

The present circuit is intended only for equipment using a 9-volt supply, and it is immaterial whether or not high-value capacitors are connected across the supply lines.

Basic Multivibrator Circuit

To enable a lamp to flash at regular intervals some form of periodic switching is required, whereupon it becomes difficult to devise a circuit which does not employ an electro-mechanical component such as a relay or thermal switch, etc., or a relaxation oscillator. Circuits incorporating electro-mechanical devices are obviously unattractive here, if only because of the operating current entailed. So far as relaxation oscillators are concerned, the simplest circuits are provided by the blocking oscillator and the multivibrator. A blocking oscillator requires fewer components than a multivibrator, but one of these components is a transformer, which will take up more space than several resistors or capacitors. So it was finally decided to produce a design based on the multivibrator.

The most important requirement of the flashing circuit is that it should draw a low average current from the battery supplying the equipment in which the circuit is to be fitted. It would be a fairly acceptable approach to have the battery supply the flashing light directly,

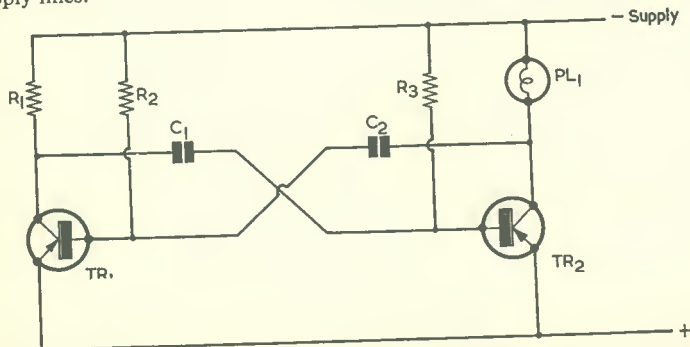


Fig. 1. A simple multivibrator which could possibly function as a flashing circuit

even though a sudden high current would be drawn from the battery during the period of the flash. This method of operation would cause a low average current to be drawn from the battery, even though the peak current was relatively high.

It might in consequence be thought that the multivibrator circuit of Fig. 1 would give the required performance. The operation of the circuit is, at first sight, very simple: the values of the resistors and capacitors can be chosen so that TR₂ is on (i.e. conductive) for a short time and is off (i.e. non-conductive) for a long time. The lamp will then only be illuminated when TR₂ is on, whereupon it would seem that the low average current requirement for a flashing indicator has been met.

Unfortunately, the simple circuit of Fig. 1 presents several design difficulties which have a serious effect on its performance. Since these difficulties have a bearing on the flashing circuit which was finally evolved, they will now be briefly considered.

If, in Fig. 1, TR₂ is to remain off for a long period, C₁ requires a relatively large value. When TR₂ comes on, to cause the bulb to be illuminated, TR₁ becomes non-conductive and C₁ commences to charge by way of R₁ and the base-emitter junction of TR₂. However, one of the requirements of the circuit is that TR₂ be conductive for a short period only. If C₁ is to acquire a reasonable charge during this short period it is therefore necessary to give R₁ a low value. But the consequence of a low value in R₁ is that, when TR₁ becomes conductive after the lamp has extinguished, a relatively high current is drawn through this resistor.

Thus, the circuit of Fig. 1 has the disadvantages, for the present application, that the full lamp current is drawn from the supply when the lamp is illuminated, and that a relatively high current is drawn from the supply when the lamp is extinguished. Although Fig. 1 is practicable, it is not because of these two shortcomings, particularly attractive.

A Practical Circuit

By the addition of another resistor and a small silicon diode, it is possible to overcome these problems to a very great extent. A practical circuit devised by the writer is shown in Fig. 2, and it represents this month's "Suggested Circuit".

As will now be explained, the circuit of Fig. 2 allows the lamp to flash at intervals of about one minute, whilst the average current drawn

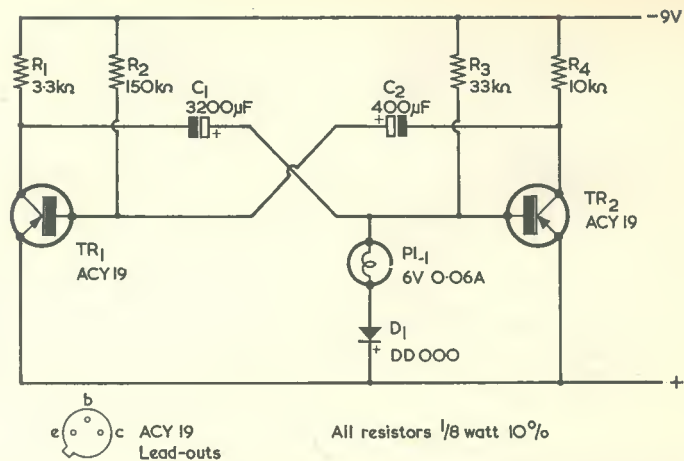


Fig. 2. Shortcomings in the arrangement of Fig. 1 are overcome in the circuit shown here, which causes considerably less current to be consumed from the 9-volt supply

from the supply is slightly less than 2.5mA, with peaks of the order of 3.7mA.

Let us commence an examination of Fig. 2 at a moment in the multivibrator cycle when TR₂ is on and TR₁ is off. In this condition, C₁ is charging via R₁ and the base-emitter junction of TR₂, whilst C₂ is discharging via R₂. The component values in the circuit are such that, when the base of TR₁ commences to go negative of its emitter, capacitor C₁ is charged nearly to the full 9 volts of the supply. As soon as the base of TR₁ goes negative of its emitter, multivibrator action takes place and the two transistors change states, with TR₂ becoming non-conductive and TR₁ becoming conductive. Transistor TR₁ now bottoms, and its collector voltage becomes very slightly negative of the positive supply line. Since C₁ is charged, its positive plate now swings positive of the lower supply line by nearly 9 volts. The pilot lamp is connected to the positive plate of C₁, and the polarity of the voltage supplied to it is such that silicon diode D₁ conducts. In consequence, capacitor C₁ discharges rapidly into the pilot lamp by way of D₁ and the bottomed TR₁, and the lamp flashes.

C₁ discharges rapidly into the lamp until the voltage on its right-hand plate is about 0.3 to 0.4 volts positive of the lower supply line, whereupon D₁, exhibiting the normal forward characteristic of a silicon diode, becomes non-conductive. The rapid discharge of C₁ ceases and it

continues to discharge, much more slowly, into R₃. After a period the discharge is sufficient for TR₂ to become conductive, whereupon multivibrator action causes the two transistors to change over again. TR₂ is now on and TR₁ is off. C₂ commences to discharge via R₂, and C₁ to charge via R₁.

This is the point at which we commenced our examination of the circuit. It will be seen that we have passed through one multivibrator cycle, during which the pilot lamp has been caused to flash once by the discharge of C₁ through it. As may be gathered, the circuit allows C₁ to charge up slowly, with the result that the current drawn from the supply is always relatively low despite the fact that the full filament current passes through the lamp when it flashes. The high value of C₁, 3,200µF, is sufficient to cause the lamp to flash brightly during the discharge.

The resistor values specified in Fig. 1 are of considerable importance for correct circuit operation. This point will be more readily appreciated after an examination of Fig. 3, which shows the voltage across C₁ and the total current drawn from the 9-volt supply during the multivibrator cycle. The curves were plotted from observations of meter readings with the prototype circuit.

In Fig. 3, zero seconds along the time scale is chosen arbitrarily to coincide with a point near the end of the period when TR₂ is on, the first changeover taking place at 10

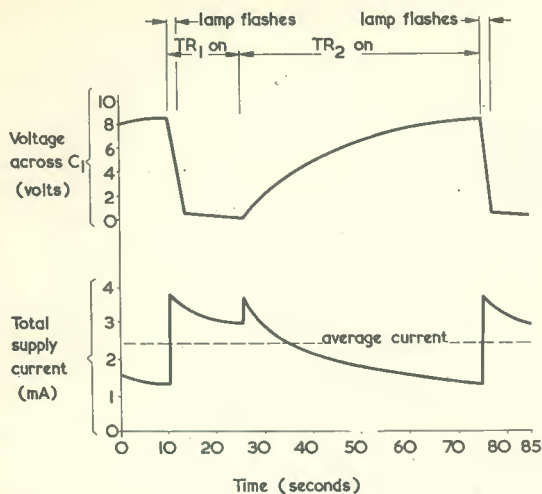


Fig. 3. Voltage and current waveforms given with the circuit of Fig. 2

seconds along the time axis. At this changeover point, TR₁ comes on and TR₂ becomes non-conductive. As shown in the upper curve, there is about 8.5 volts across C₁ and this rapidly drops to about 0.4 volts as the capacitor discharges into the lamp. The period of rapid discharge is indicated as "lamp flashes" in the diagram. Also at the 10 second point, the total current drawn from the supply rises suddenly from about 1.4mA to some 3.7mA. The reason for this increase will be discussed shortly.

At the end of the "lamp flashes" period in Fig. 3, C₁ has a potential of about 0.5 volts. C₁ continues to discharge (into R₃) until, at about 25 seconds along the curve, the voltage across its plates is nearly zero. TR₂ then comes on whilst TR₁ becomes non-conductive, and the second changeover takes place. This fact is indicated in the lower curve by a sudden rise in current from 3 to 3.6mA. With TR₂ on, capacitor C₁ commences to charge, and the voltage across its plates slowly increases until, at about 75 seconds along the curve, it is around 8.5 volts. The increasing voltage across C₁ is accompanied by a decreasing current drawn from the 9-volt supply, as shown in the lower curve. Around 75 seconds the second changeover takes place, and another cycle commences.

The lower curve may now be analysed in detail. There is, at the 10 second point, a rise in current to some 3.7mA. Ignoring leakage currents, this 3.7mA is due to the

current drawn through R₁ via TR₁ (which has now come on), plus the charging current in C₂ via R₄ and the base-emitter junction of TR₁, plus the smaller currents in R₂ and R₃. This total current falls fairly rapidly to about 3mA, the fall being due to the decreasing charging current in C₂. When TR₂ comes on, at about 25 seconds along the time scale, the current again rises to a peak, this time of about 3.6mA. Ignoring leakage currents, the 3.6mA is given by the current drawn through R₄ by TR₂ (which is now on), plus the charging current in C₁ via R₁ and the base-emitter junction of TR₂, plus the smaller currents in R₂ and R₃. As C₁ becomes more fully charged, the total current drawn from the supply drops to about 1.4mA, and consists almost entirely of the currents drawn via R₂, R₃ and R₄.

It may now be appreciated that the values of the resistors are critical if a low average supply current is to be achieved. The most important value is that for R₁, since it must have a value sufficiently low to allow C₁ to become nearly fully charged when TR₁ is off, yet not so low to allow too great a current to be drawn when TR₁ is on. The value quoted, 3.3kΩ, allows slightly less than 3mA to flow when TR₁ is on, and provides the best practical compromise between the two conflicting requirements for a 1-minute cycle.

When TR₁ comes on it must bottom sufficiently hard for C₁ to pass a 60mA pulse of current

through the pilot lamp. Assuming a minimum gain of 80 in TR₁ (the minimum figure specified by Mullard for an ACY19) at least 1/80 of 60mA must, therefore, flow in TR₁ base. R₄ has a value of 10kΩ, which allows a pulse of slightly less than 1mA to flow through the initially discharged C₂ into the base of TR₁ immediately after the changeover, and this is sufficient to allow the lamp to flash. The value of 10kΩ is therefore about the highest which can be given to R₄ if reliable flashing is to be achieved. (Indeed, should a low-gain specimen be used in the TR₁ position, it may be necessary to slightly reduce the value of R₄ to obtain a good flash). This means that, during the period when TR₂ is on, the standing current it draws cannot be reduced below a value which is slightly less than 1mA. An increase in R₄ to reduce this current may prevent reliable flashing from occurring.

The period during the cycle after the termination of the flash, and until TR₂ comes on, detracts from the overall efficiency of the circuit because no useful function is being carried out. All that happens here is that C₁ is discharging slowly into R₃ whilst TR₁ is drawing nearly 3mA because of the necessarily low value in R₁. This period could be decreased by reducing the value of R₃. If, however, R₃ is reduced below the 33kΩ specified, the advantage gained—a reduction of the length of the useless period—is offset by the additional current flowing throughout the cycle in R₃ itself! The value specified for this resistor represents the best compromise between these two opposing effects.

The values of C₂ and R₂ control the time that TR₁ is off and C₁ is allowed to charge up, and it was found that 400μF and 150kΩ worked well in practice. Bearing in mind that leakage currents may be present in both C₂ and TR₁, the value chosen for R₂ is somewhat high. At the same time, if this value is reduced the capacitance in C₂ has to be increased and the latter component becomes more expensive. In practice, the prototype circuit functioned perfectly with a 150kΩ resistor in the R₂ position. To check that a measure of reliability was still "in hand", a 470kΩ resistor was temporarily connected in the R₂ position. The circuit still functioned satisfactorily, although the length of the multivibrator cycle was, of course extended.

Practical Points

The components employed in the circuit require a little comment.

The pilot lamp is a Radiospares

6 volt 0.06 amp m.e.s. bulb, and is available from Home Radio (Mitcham) Ltd. under Cat. No. PL7. It gives a bright flash when the 3,200 μ F capacitor discharges through it, but it has to be remembered that it has a low-consumption filament and that, if it is fitted in a panel mounting, the latter should be provided with a clear "lens" to allow maximum illumination to pass through. The silicon diode, D₁, is a Lucas DD000.

The two transistors should be ACY19, as shown, or equivalents for gain and, in the case of TR₁, current handling.

Capacitor C₁ in the prototype was

a Mullard electrolytic component in the C431BR series, with a value of 3,200 μ F and a maximum voltage rating of 10. Capacitor C₂ may be any electrolytic capacitor of 400 μ F with a working voltage of 9 or more.

When installed in a piece of equipment, the circuit may be connected across the 9-volt supply lines immediately after the on-off switch, whereupon the lamp will flash at intervals of approximately 1 minute at an average current consumption of slightly less than 2.5mA. As may be seen from Fig. 3, the lamp in the prototype flashed at intervals of about 65 seconds.

The prototype circuit was checked with a brand-new PP9 battery having a terminal e.m.f. of 9.5 volts (which, assuming negligible leakage currents, is reflected in the fact that the currents shown in the lower curve of Fig. 3 tend to be marginally above their calculated values). To stimulate the effect of a run-down battery, the circuit was then checked with a 6-volt battery. Although the flash given at this lower voltage was less brilliant than with the 9-volt battery it was still noticeable and capable of attracting attention.



WIDE-BAND PHOTOPHONE, PT. 2

by D. BOLLEN

It should be noted that some of the equipment described in these articles is the subject of a provisional patent application

Installation of the Transmitter

CAREFUL SCREENING IS NECESSARY AT THE TRANSMITTER end if a.c. bias is to be employed, otherwise interference will be caused to nearby long and medium wave receivers. The reflector and tube can be mounted on a pole, or placed in a window, where line of sight transmission of several hundred yards can be arranged.

Switch on the power amplifier and tube power supply, and either dab a finger on the open circuit

power amplifier input or feed a high level signal to the tube. This should cause the tube to fire and the 100mA meter to read. If all is well, a loud hiss should be audible from the receiver, with possibly a low note of around 700 c/s generated by the tube. With gain set at zero, switch on the pre-amplifier and bias, and adjust VC₁ of Fig. 6 (published last month) until the noise at the receiver diminishes. If there is any instability in the system, the background mush will continue or get louder.

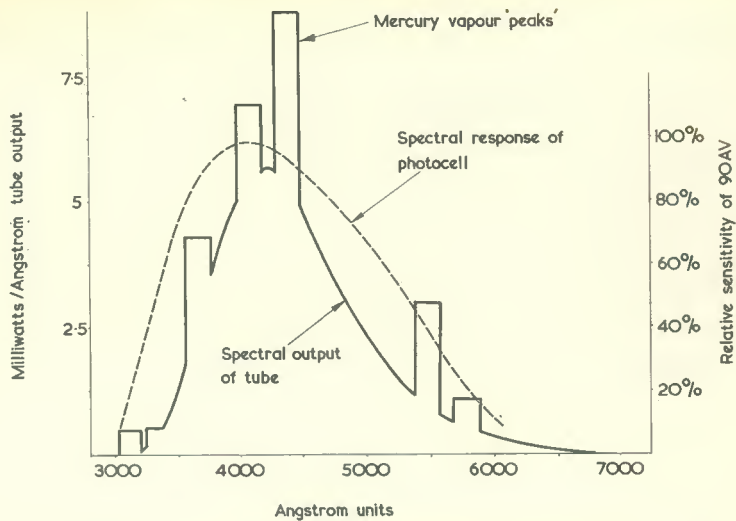


Fig. 9. The spectral output of a Blue fluorescent tube and the relative sensitivity of the 90AV caesium antimony photocell. The 90AV photocell is employed in the receiver described in this article

When these arrangements are complete a signal can be transmitted and range tests begun. At full modulation the needle of the 100mA meter will begin to flicker and, at the point where excessive modulation is applied, the reading will rise appreciably. At distances up to 100 yards very simple receivers will serve for tests. A selenium cell coupled to a three-transistor a.f. amplifier allows intelligible short range working, but the frequency response will be poor. Such an arrangement, however, may be used as a "field strength" meter, prior to more sophisticated reception arrangements. A phototransistor type of receiver will give full frequency response, but tends to be noisy in operation.

For comparison, a 15 watt 240 volt bulb was connected in place of the fluorescent tube, under similar conditions with the exception that the dropper resistor R_1 of Fig. 2 (published last month) was short-circuited so that the lamp glowed brightly. With a wide-band phototransistor receiver, full

treble boost and maximum bass cut from a QUAD pre-amplifier (added to the large degree of treble boost given in the Photophone's compensation network), and taking 200 c/s as the 0db reference point, the signal was down to -10db at 5 kc/s, and -15db at 15 kc/s. At 15 kc/s the bulb was grossly overloaded and threatened to burn out, and this was the highest level of treble pre-emphasis that could be reasonably applied. Under the same circumstances, the fluorescent tube will peak above 30 kc/s and maintain its output to well beyond 60 kc/s. Without a lens to focus its light to a beam, range from the 15 watt bulb was inferior, being scarcely 100 yards.

Overall Efficiency

The limit to range with any system employing modulated light is set by receiver noise, light excursion or degree of modulation of the transmitter lamp, and the matching of the lamp's peak spectral response to the optimum spectral response of the receiving photocell, photodiode, or phototransistor.

With a good design of receiver, the presence of strong ambient light, during the hours of bright sunlight, should have no effect on receiver sensitivity or noise. The resultant effect may be likened to a large d.c. component upon which a small a.c. signal is imposed, daylight being the d.c. component. If a vacuum photocell operated at a low voltage is used, this should contribute virtually nothing to the noise generated by its following amplifier. Unfortunately, phototransistors are not quite so good in this respect, and their noise tends to be significantly high. Light excursion is mainly dependant on the type and power of lamp used. Due to its thermal inertia, a filament bulb is very inefficient at anything approaching high frequencies and the

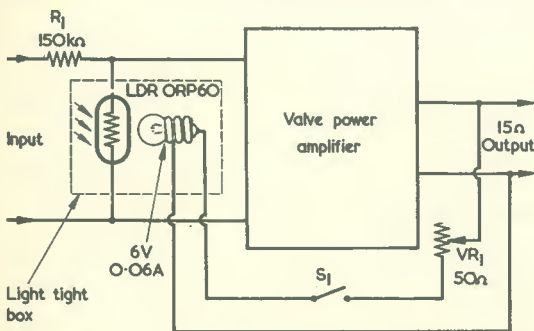


Fig. 10. The simple volume compression circuit employed at the transmitter to obviate overmodulation

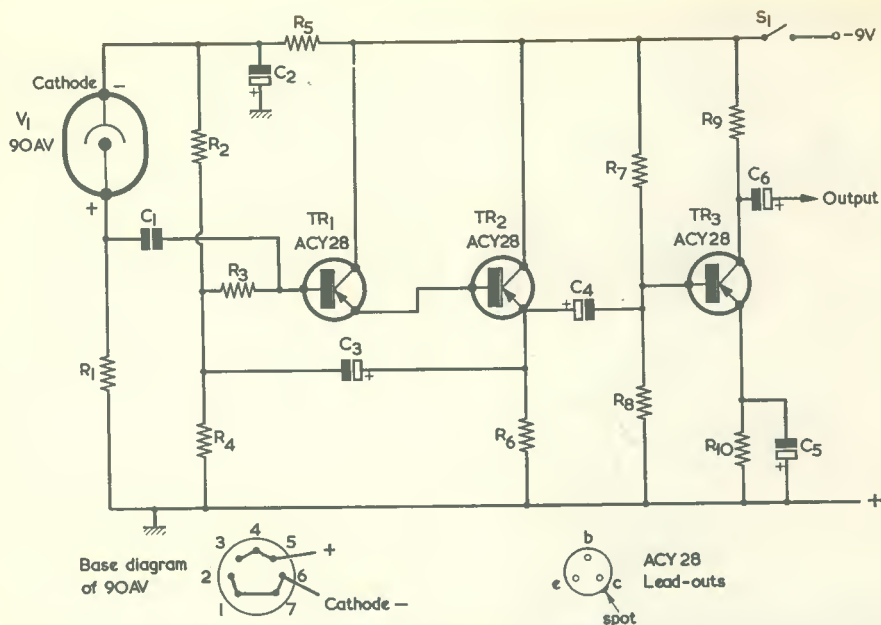


Fig. 11. The circuit of the photocell and pre-amplifier unit. It is recommended by the manufacturers of the photocell that the cathode connection should be made to pins 1, 2, 6 and 7 connected together, and the anode connection to pins 3, 4, and 5 connected together. The output connects to the subsequent amplifier via screened cable with the braiding connected to chassis

COMPONENTS

Resistors

(All resistors $\frac{1}{4}$ watt 10%)

- R₁ 300k Ω
- R₂ 47k Ω
- R₃ 300k Ω
- R₄ 47k Ω
- R₅ 12k Ω
- R₆ 4.7k Ω
- R₇ 27k Ω
- R₈ 4.7k Ω
- R₉ 4.7k Ω
- R₁₀ 1k Ω

Capacitors

- C₁ 0.1 μ F paper
- C₂ 100 μ F electrolytic, 12V wkg.

- C₃ 15 μ F electrolytic, 12V wkg.
- C₄ 15 μ F electrolytic, 12V wkg.
- C₅ 100 μ F electrolytic, 12V wkg.
- C₆ 15 μ F electrolytic, 12V wkg.

Transistors

TR_{1,2,3} ACY28 (or equivalent)

Photocell

V₁ 90AV (Mullard)

Switch

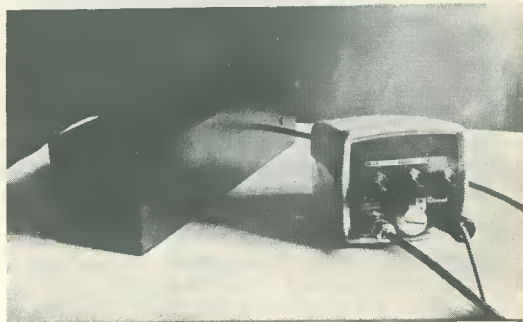
S₁ s.p.s.t., toggle

Valveholder

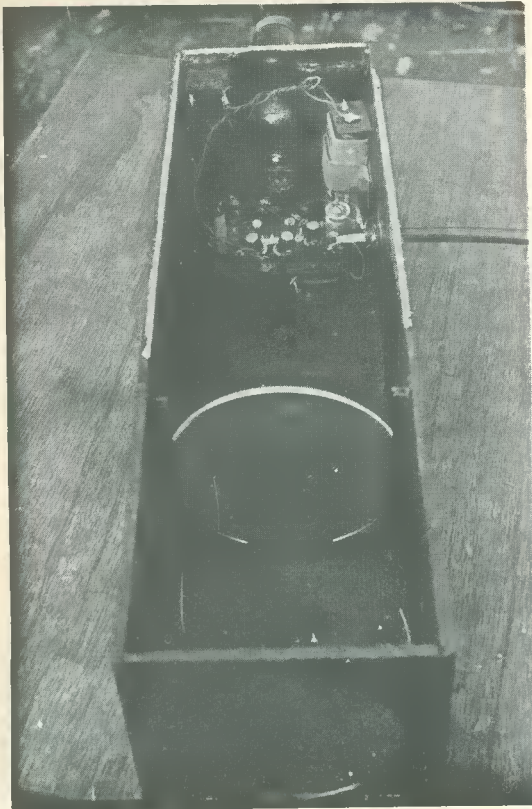
B7G valveholder with centre spigot

more powerful the bulb the worse does this become, thus setting a natural limit to transmitter power. Discharge tubes will permit large modulation powers without loss of treble response but these, too, do not normally respond to very high frequencies because of the ionisation effects of their gas filling. The efficiency of a fluorescent tube in this context is something of a mystery, but the writer attributes it to the ability of its phosphor coating to respond rapidly to changes in the ultra-violet content of the internal discharge. It is this ultra-violet which activates the phosphor and causes it to throw off visible light. The performance of different types of tube has already been discussed in Part 1 of this series.

The light output from filament bulbs peaks



The collimator box coupled by coaxial cable to the portable a.f. amplifier



The inside of the receiver collimator box. The photocell is at the rear with the pre-amplifier before it. The battery is fitted to the right-hand wall of the housing, and the on-off switch to the rear wall at top left

towards the infra-red end of the spectrum, and phototransistors and caesium oxidised silver photocells are sensitive to this. The fluorescent tube's maximum output lies towards the blue and ultra violet and, depending on the phosphor employed, a caesium antimony cell will provide a good spectral "match". The light output from a fluorescent tube is rather peculiar, consisting of sharp peaks of varying heights and widths distributed along the spectrum. With a caesium antimony photocell, the most efficient tubes to use for an optimum match are Ultra Violet, Blue, and Daylight coatings, in that order. The Ultra Violet tube will be approximately four times as efficient as the Daylight tube. In fact, performance figures given here for the Photophone were based upon Warm White tubes, which are even a less efficient match than the Daylight, so there is considerable scope for improvement. Fig. 9 shows the spectral responses of a Blue tube and a caesium antimony photocell.

With the prototype Photophone transmitter at full drive, distortion on peaks was objectionable, particularly when a music signal was being transmitted, and it was difficult to strike a suitable balance between distortion and long range with a

good signal-to-noise ratio. Eventually, the simple volume compressor of Fig. 10 was added to the Photophone power amplifier, and this proved to be most effective. A 6V 0.06A bulb is encased in a small light-tight container, in close contact with a light dependent resistor type ORP60. As the amplifier output approaches the point where the fluorescent tube overloads and clips the peaks the bulb tends to glow brightly, and its light causes the resistance of the l.d.r. to drop and shunt the p.a. input, which in turn reduces p.a. output. VR₁ controls the extent to which compression is applied and S₁ allows compression to be completely removed when desired. With compression, a very high modulation level can be maintained, to enhance signal noise ratio and increase "talk power" without distortion.

Range Tests

For range tests with the original, distances were determined by taking a 6 inch to 1 mile map of the area, marking it with 100 yard circles centred on the transmitter location, and identifying the position of the receiver by means of landmarks. The quality of music signals up to 200 yards was excellent, but beyond that the noise level of the receiver tended to drown the quiet passages unless treble cut was used to diminish receiver hiss. At 400 yards speech signals were completely intelligible and the vocal "character" of the talker came over well. 500 yards seems to represent a "light barrier" to all moderate power light signals. A tone could just be heard at that distance, using a single 20 watt fluorescent tube at the transmitter, but speech was indistinct.

The Receiver Equipment

We next turn to the equipment fitted at the receiver end. As has already been mentioned, the performance of a modulated light receiver will be dependent on self-generated noise, ambient light, and correct matching of the transmitter spectral output to the receiver's photosensitive device. There is also a further factor, this being the optical coupling between transmitter and receiver. Phototransistors were ruled out for the author's receiver because they are noisy and do not provide a good spectral match to a fluorescent tube. On the other hand, a caesium antimony photocell can be operated at a low voltage to give a good match and adequate gain, with suitable transistor circuitry, as well as a very low noise performance.

The author's receiver was constructed in two units, these consisting of a photocell pre-amplifier, and a portable a.f. amplifier with tone controls. The former unit is suitable for feeding directly into a fixed-location valve or transistor amplifier. The portable amplifier was designed to give a good quality output to a pair of Hi-Fi headphones. Alternatively, it can feed into a small 25 to 30Ω speaker at around 400 milliwatts.

Photocell Pre-amplifier

The circuit of Fig. 11 shows how a 90AV caesium

antimony photocell is coupled to a bootstrapped Darlington Pair. This particular circuit offers a very high input impedance and has the advantage that, with TR_1 passing a very small collector current, low noise amplification is possible. Capacitor C_3 feeds the bootstrap in-phase signal to the bias network end of R_3 , and the a.c. signal from the photocell passes, via C_1 , to the base of TR_1 . With this arrangement, tests were conducted using higher photocell voltages, up to 48 volts, but there was no measurable increase in sensitivity and photocell noise became apparent as voltage increased. C_1 effectively blocks the large standing current when the photocell is functioning in bright sunlight, but allows the minute a.c. signal superimposed on the

d.c. to pass freely to the amplifier. R_1 was selected to give the best overall performance. TR_3 is a straightforward common-emitter amplifier, coupled to the emitter of TR_2 . At short ranges, the circuit of Fig. 1 gives headphone strength signals, and the overall response of the pre-amplifier is substantially flat to more than 60 kc/s. The photocell itself has an inter-electrode capacitance of only 0.7pF and appears to be able to respond to frequencies well beyond the capabilities of the pre-amplifier. The output from the pre-amplifier is coupled to the following amplifier by means of a length of coaxial cable.

(to be concluded)



SEMICONDUCTOR

DEVICE CODING SYSTEMS

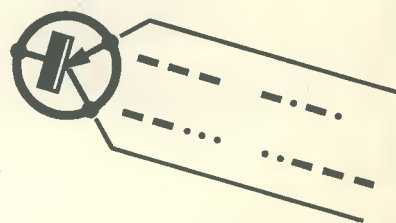
by A. Thomas

Notes on some of the semiconductor coding systems currently encountered in the U.K.

NOW THAT THERE ARE SUCH A VAST NUMBER OF semiconductor devices obtainable, it is thought that a review of some of the coding systems in use in the U.K. would be helpful for the amateur enthusiast.

Semiconductor manufacturers have slipped up in not realising the vast potential of their devices and the need for a significant coding system right from the start. This has been corrected somewhat in Europe and Great Britain by the Pro-Electron system but, even now, some manufacturers stick to their own individual systems.

In America, right from the early days, a serial coding has been in use. The system is purely a serial number which has no device significance. This system is known as the JEDEC code, and any manufacturer may submit a complete set of device specifications to a registering body, and a code is issued; any other manufacturer may then use the code if their device follows the specification entirely. The writer has seen many alternative significant systems proposed in the American technical press, but none have been universally adopted.



Pro-Electron Code as Applied to Semiconductors

This system is controlled by the organisation Pro-Electron in Luxembourg, and enjoys a wide use on the Continent and in this country.

For industrial devices the code consists of three letters and two digits. The first letter specifies the device material, whilst the second letter specifies the device function as shown in Table I. The remaining letter and two digits constitute a serial number.

For entertainment devices the code is modified to give immediate indication that it is an entertainment device. The first two letters carry the same significance as above and this is followed by three figures which constitute the serial number.

JEDEC Code

Under this coding system, the devices are registered with the Joint Electron Device Engineering Council, and as previously stated, there is no device significance. Transistors to this code are obtainable from a great many manufacturers in this country, whether the company is of American parentage or not. The code starts with 1N, 2N, or 3N, and may have a prefix, USA (army), USN (navy), JAN (joint army navy), if for military (MIL) use.

Newmarket House Code

Newmarket Transistors Ltd. have a significant coding system, and this is set out in Tables II and III.

CV Specifications

CV specifications apply to transistors, diodes and

valves, and to other electronic devices manufactured for use by Military, Post Office and other Government departments. The code consists of CV followed by a four figure serial number which has no device significance. There are several books of equivalents available, but these are not entirely up to date.

Mullard Coding

This company codes the majority of its devices to the Pro-Electron system, some being available to JEDEC codes.

Some exceptions are:

Germanium diodes and rectifiers, GEX, GEM, OA (2 figures)

Silicon diodes and rectifiers, OA (3 figures)

Germanium transistors, GET, OC (up to 199)

Silicon transistors, OC (200 up)

Phototransistors, OCP

Photodiodes, OAP

Photo cells, ORP.

TABLE I
Pro-Electron Code as Applied to
Semiconductors

First Letter. Material.

- A. Germanium.
- B. Silicon.
- C. Intermetallic Compound.
- R. Materials for radiation detectors, Hall Effect generators, etc.

Second Letter. Function.

- A. Diodes, (apart from other diodes listed below), including voltage dependent capacitors.
- C. Transistors for a.f. applications, low power.
- D. Transistors for a.f. applications, high power.
- E. Tunnel diodes.
- F. Transistors for r.f. applications, low power.
- G. Miscellaneous.
- H. Field Probe.
- K. Hall Generator.
- L. Transistors for r.f. applications, high power.
- M. Hall Modulators and multipliers.
- P. Photo semiconductors, including radiation devices.
- R. Breakdown devices, low power, including controlled switches, etc.
- S. Transistors for switching applications, low power.
- T. Breakdown devices, high power, controlled rectifiers, p.n.p.n. diodes.

- U. Transistors for switching applications, high power.
- Y. Diodes, high power.
- Z. Zener diodes.

TABLE II
Newmarket House Code—Silicon Transistors
NKT. 1st digit. 2nd digit. 3rd digit. 4th digit.
5th digit.

1st Digit. Fabrication.

- 1. Epitaxial n.p.n.
- 2. Epitaxial p.n.p.
- 3. Diffused n.p.n.
- 4. Diffused p.n.p.

2nd Digit. Frequency (f_T).

- 0. Less than 100 Mc/s.
- 1. 100 Mc/s min.
- 2. 200 Mc/s min.
- 3. 300 Mc/s min.
- etc. up to
- 9. 900 Mc/s min.

3rd Digit. Gain (h_{FE}).

- 0. Less than 10.
- 1. 10 min.
- 2. 20 min.
- 3. 40 min.
- 4. 80 min.
- 5. 160 min.
- 6. 320 min.
- etc.

4th Digit. (V_{CBO}).

- 0. Less than 15V.
- 1. 15V min.
- 2. 30V min.
- 3. 45V min.
- 4. 60V min.
- 5. 90V min.
- 6. 120V min.

5th Digit.

Device Registration Number.


TABLE III
Newmarket House Code—Germanium
Transistors

NKT. 1st digit. 2nd digit. 3rd digit.

1st Digit. Type.

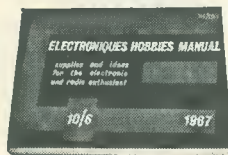
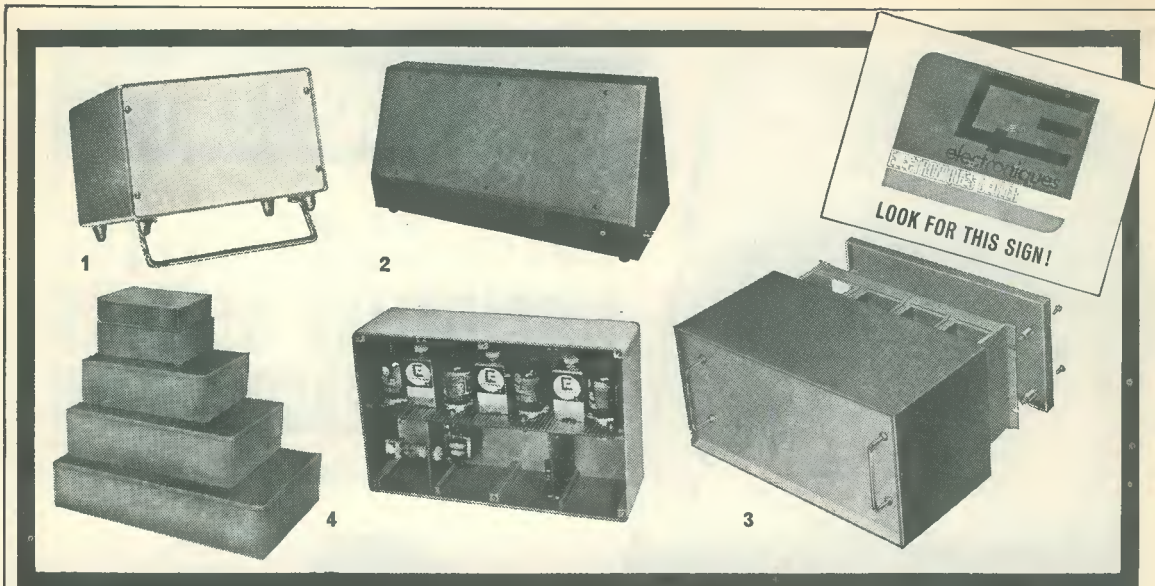
- 1. R.F. p.n.p.
- 2. A.F. p.n.p.
- 3. Intermediate Power p.n.p.
- 4. Power p.n.p.
- 6. V.H.F. p.n.p.
- 7. A.F. n.p.n.

2nd and 3rd Digits.

Device Registration Number. 

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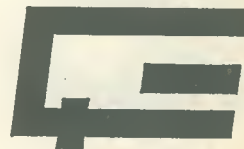
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Transistor Direct-Reading Frequency Meter

by

C. Crosbie

A reliable design which employs circuit principles that are simple to understand, and which offers a direct indication of frequency up to 100 kc/s.

SEVERAL MODELS OF THE DIRECT-READING FREQUENCY meter described in this article have been built by the author and have given reliable results without introducing any difficulties. The principle of operation is easy to follow, and the method employed for final setting up dispenses with close-tolerance component.

The specification is as follows;

- Frequency range—10 c/s to 100 kc/s
- Minimum operating signal—0.5 volt peak
- Input impedance—1k Ω approx.
- Power supply—10mA at 6V.

Circuit Operation

The circuit of the meter appears in Fig. 1. The frequency to be measured is applied to TR₁ via capacitor C₁, the input voltage being 0.5 volt or more. TR₁ is a conventional amplifier and is overloaded by this input signal, producing a squared wave at its collector. The collector voltage can only swing by about ± 1.5 volts from its no-signal condition. The waveform shown in Fig. 2 (a) represents an input sine wave at the base of TR₁, whilst that at Fig. 2 (b) shows the squared wave at its collector.

It is desired to obtain one pulse per input cycle, and so the square wave is next applied to the differentiator given by C₃ to C₆ (as selected by S_{1(a)}) and R₅. S_{1(a)}, which is a section of the range switch, selects suitable capacitance values for the range required. The differentiated signal applied to diode D₁ is illustrated in Fig. 2 (c), where it may be seen that positive and negative "spikes", or pulses, are produced.

Diode D₁ performs two functions. Firstly, it selects only positive pulses and, secondly, only the top parts of such pulses. This latter function is necessary due to the imperfect squaring given by TR₁, which allows small spurious pulses to appear at the foot of the main pulse. Thus, one positive pulse for each input cycle is available at the cathode of D₁, as shown in Fig. 2 (d).

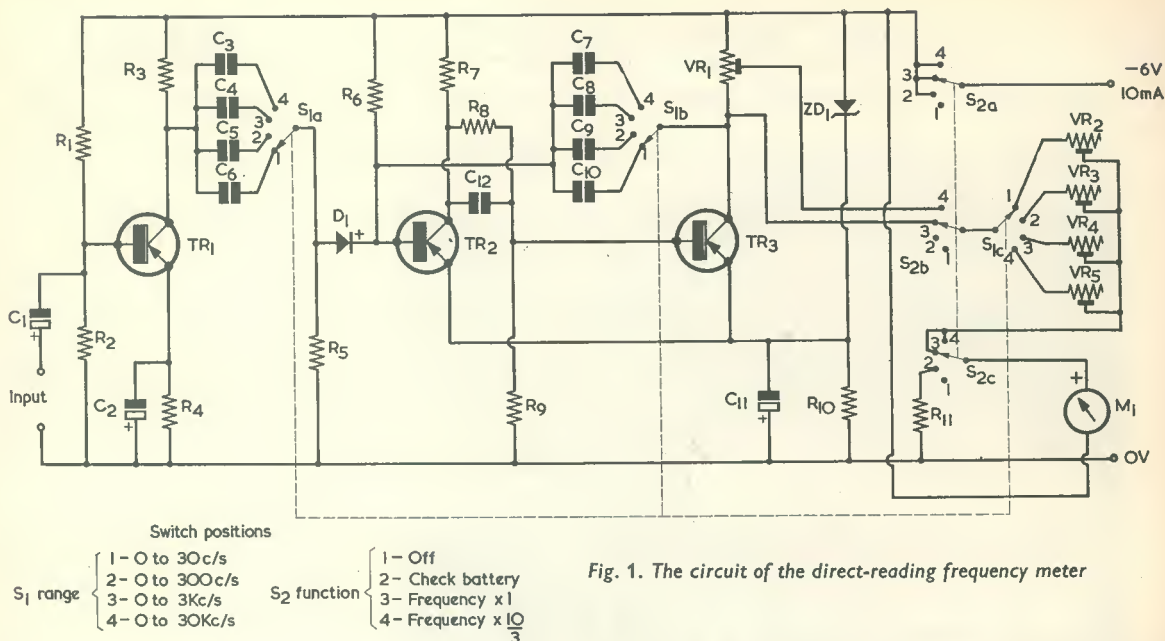


Fig. 1. The circuit of the direct-reading frequency meter

COMPONENTS

(All fixed values $\frac{1}{4}$ watt 10%)

- R₁ 22k Ω
- R₂ 10k Ω
- R₃ 1.5k Ω
- R₄ 1k Ω
- R₅ 1.5k Ω
- R₆ 10k Ω
- R₇ 1k Ω
- R₈ 4.7k Ω
- R₉ 6.8k Ω
- R₁₀ 150 Ω
- R₁₁ See text
- VR₁ 1k Ω preset potentiometer
- VR₂-VR₅ preset potentiometers (see text)

Capacitors

- C₁ 25 μ F electrolytic, 6V wkg.
- C₂ 250 μ F electrolytic, 3V wkg.
- C₃ 100pF ceramic or silver-mica, 20%

- C₄ 0.005 μ F plastic foil, 20%
- C₅ 0.05 μ F plastic foil, 20%
- C₆ 0.47 μ F plastic foil, 20%
- C₇ 100pF ceramic or silver-mica, 20%
- C₈ 0.005 μ F plastic foil, 20%
- C₉ 0.05 μ F plastic foil, 20%
- C₁₀ 0.47 μ F plastic foil, 20%
- C₁₁ 250 μ F electrolytic, 3V wkg.

C₁₂ Value between 330 and 470p (see p 632)

Semiconductors

- TR₁, TR₂, TR₃ OC45 or OC44
- D₁ Silicon diode, OA200 or OA202
- ZD₁ 4.7 volt zener diode, OAZ200 or OAZ240

Switches

- S₁, S₂ 3-pole 4-way

Meter

- M₁ Moving coil meter (see text)

Battery

- 6-volt battery

Pulse-Shaper

The pulse-shaper consists of a conventional monostable multivibrator, of which a simplified diagram is shown in Fig. 3. Fig. 3 illustrates how, due to the constant voltage (of the order of 4.7 volts) appearing across zener diode ZD₁, the circuit may be looked upon as having a main supply of 4.7 volts negative together with a bias supply of 1.3 volts positive.

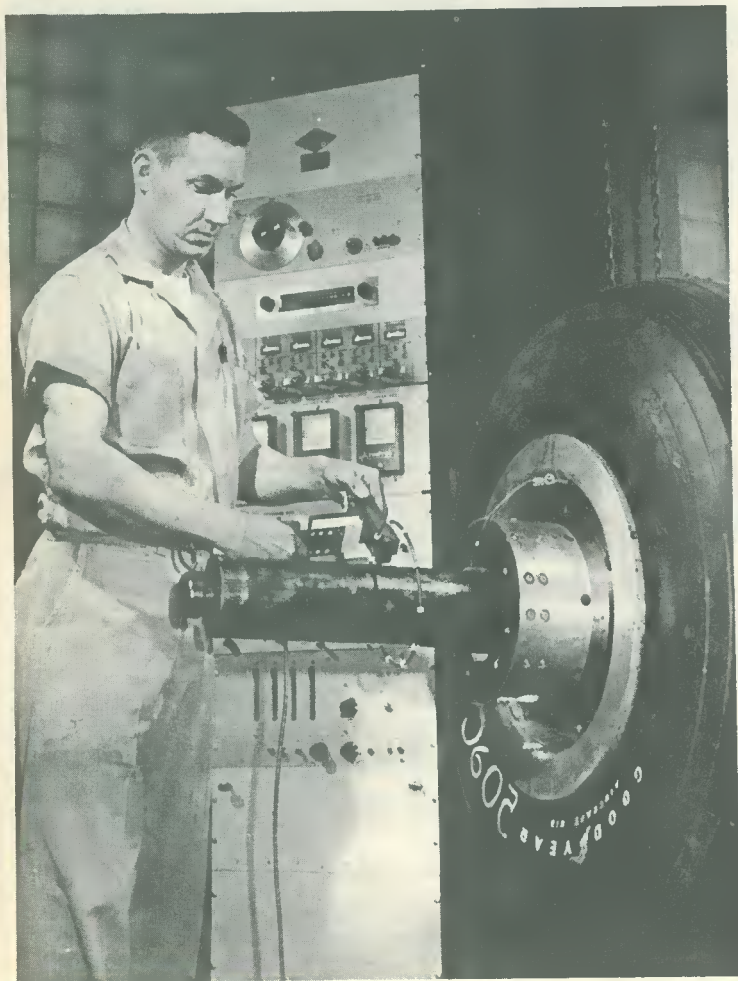
In the absence of signal, the positive bias voltage on TR₃ base holds this transistor off, transistor

TR₂ being fully on. A positive pulse applied to the base of TR₂ causes this transistor to become non-conductive, whereupon multivibrator action brings TR₃ on. TR₃ remains on until the capacitor switched in by S_{1(b)} has discharged sufficiently into R₆ for TR₂ to become conductive again. Thus, a positive-going pulse appears at the collector of TR₃ whose length depends upon the values of C₇ to C₁₀ and R₆. Switch S_{1(b)} ensures that the capacitance value selected provides a pulse that

(continued on page 476)

NEWS . . . AND .

Talking Tyre Reports on its Performance by Radio



American engineers can now fix a tiny radio transmitter to a tyre and pick up information about its performance.

The miniature radio transmits extremely accurate measurements of the temperature and inflation pressure of the tyre while it is being accelerated and decelerated on a wheel, and subjected to other realistic operating conditions.

Information on pressure comes from a tiny thermistor attached to the valve. Temperatures are measured by four thermistors, each about the size of a pinhead. Three are embedded in different parts of the tyre. The other is suspended in the air inside.

These devices convert their measurement to electrical signals which are then broadcast by the transmitter over five channels simultaneously and picked up by an FM receiver attached to machines which record the data as graphs.

The system, developed by Goodyear, is enabling engineers to test new materials and designs for both car and airplane tyres.

Picture on left shows a technician adjusting the miniature radio transmitter which broadcasts information about a tyre's pressure and temperature as the wheel is speeded up and slowed down

New Thames-Side Venue for VHF Convention

The Radio Society of Great Britain is taking its annual VHF Convention out of London this year for the first time, and is to hold it at the Winning Post Hotel at Whitton, near Twickenham. The date is 13th May.

The Convention—this year's will be the thirteenth in the series,

always has an international flavour, and attracts many v.h.f. transmitting men from the nearer European countries.

As in past years, a lecture session will occupy the afternoon proceedings. It will be followed by a banquet in the evening at which several distinguished personalities from the

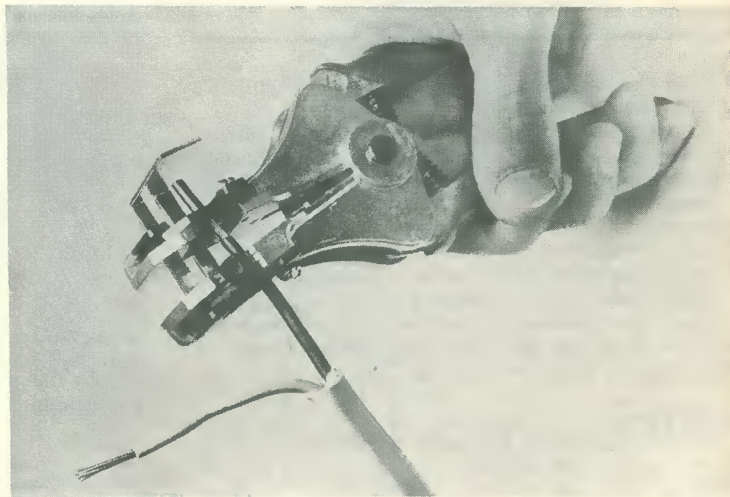
electronics profession will be present.

A comprehensive trade show by v.h.f. equipment specialist suppliers and a home construction competition will once again play their part in making the Convention one of the RSGB's most successful technical and social events.

COMMENT

Precision Wire Strippers for Miniature Assemblies

Plansel automatic wire and cable-stripping pliers, a product of Creators Ltd., are now obtainable from Henri Picard & Frère, Ltd., as a new addition to their range of tools for miniature electrical assembly work. The Plansel is a precision tool, with interchangeable blades capable of accurately stripping conductors from $\frac{1}{32}$ in to 0.005in and from 4mm to 0.4mm. The action is synchronised with a single pressure on the handles, which causes one jaw to grip the cable, while the other strips the insulation gently sideways with an automatic release to avoid damage to fine stranded cables. One set of blades is supplied with each tool and others are available from stock.



Portable Electric Tools Safe and Unsafe

With the increasing number of do-it-yourself enthusiasts there has been a startling rise in the number of accidents due to the improper use of portable electric tools.

Aware of this problem, Wolf Electric Tools Limited, who manufacture what are probably the safest "double insulated" portable electric tools on the market, together with the British Safety Council, have produced a useful "at-a-glance" ten point safety code in the form of a double-crown poster for use in factories, mines, building sites, garages, technical colleges and even in the home workshop.

Says Council Controller James Tye: "We are concerned in the British Safety Council at the number of accidents involving electric tools when the elementary rules are ignored. When observed, the ten points in this safety code would prevent 95% of the accidents involving portable electric tools."

The contents of the code are set out in the adjoining paragraph.

The above remarks are concerned with safety in the use of tools, unfortunately from time to time one learns of tools and equipment which are themselves unsafe. One such case concerns some electrical soldering irons imported from Hong Kong, in respect of which the Home Office has issued a warning because the insulation is inadequate and there is no provision for earthing.

The irons have black plastic pistol-grip handles 4 or 5in long. Sold in polythene bags, the tools bear one of two labels. One states: "Power Soldering Iron 40 watt 200-240 volt. Pistol Grip. Empire Made." The other states: "Soldering Iron 40 watt 240 volt. Made in Hong Kong."

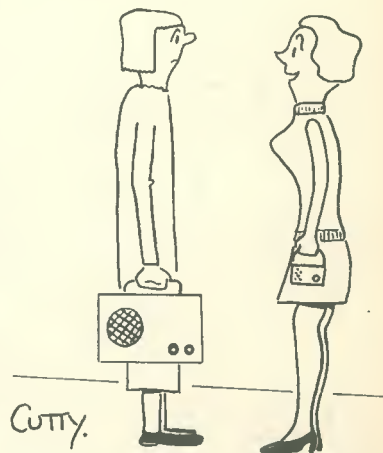
The Hong Kong government reported that the soldering irons were produced in 1964 to the specifications of a British buyer who submitted a sample to the factory. A total of 3,500 of the tools were made. The factory closed down in 1965.

No further consignment of this particular type of tool was exported by the factory after this sole transaction. The Colony's Commerce and Industry Department was seriously concerned about the matter, and carried out a full investigation.

This particular case does not mean that other electrical goods imported from Hong Kong are unsafe.

Safety Code:

1. Check that voltage on tool nameplate agrees with supply voltage.
2. Tools with 3 core cables must be efficiently earthed.
3. Double-insulated tools (2 core supply) must bear the B.S.I. "Kitemark".
4. Connect leads firmly to the correct terminals.
5. Always treat cables with care.
6. Ensure that tools are properly plugged into supply sockets.
7. Remove plug from supply socket before making any adjustment to tool.
8. Never tamper with machines.
9. Report defects immediately.
10. Wear goggles when grinding or sanding.



"Get with it Gladys, you are about as modern as a valve mains-portable"

Transistor Direct-Reading Frequency Meter

(continued from page 473)

is always shorter than a cycle of any frequency in the range selected.

It is necessary to include the pulse-shaper offered by TR₂ and TR₃ in the frequency meter circuit because the pulses obtained after D₁ are not constant in amplitude. They depend on the rise time of the squared output from TR₁ which, in turn, depends on both the amplitude and frequency of the input signal. The pulses at TR₃ collector remain constant in shape and length irrespective of input amplitude (above the minimum specified) and frequency (within the range selected). The output pulses at TR₃ collector are shown in Fig. 2 (e).

Meter Circuit

Since equal pulses are produced at the collector of TR₃ at the rate of one per input cycle, it follows that the average voltage across VR₁ becomes directly related to input frequency. As input frequency increases, so does pulse frequency and, hence, the average voltage across VR₁. This average

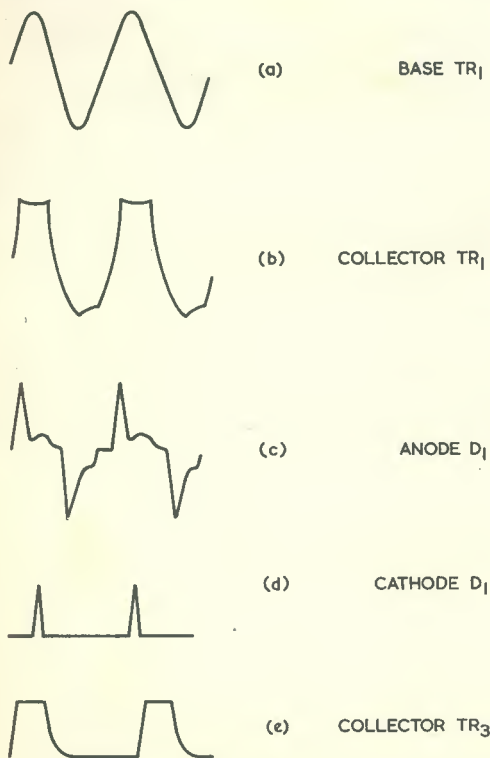


Fig. 2. The waveforms appearing at the circuit points indicated

voltage is measured by the voltmeter circuit given by VR₂ to VR₅ and meter M₁.

When function switch S₂ is in position 3, the lower end of VR₁ is applied to whichever variable resistor—VR₂ to VR₅—has been selected by S_{1(c)}, and hence to meter M₁. The negative terminal of M₁ connects to the top end of VR₁ whereupon it becomes capable of reading the average voltage across VR₁. VR₂ to VR₅ are preset variable resistors which are employed to take up variations, within tolerance, in capacitors C₇ to C₁₀. These resistors are adjusted so that, when S₁ is set to position 1, 2, 3 and 4 respectively, full-scale readings are given in the meter at frequencies of 30 c/s, 300 c/s, 3 kc/s and 30 kc/s. In consequence, with S₂ in position 3 (corresponding to "Frequency × 1") range switch S₁ selects 0-30 c/s, 0-300 c/s, 0-3 kc/s and 0-30 kc/s.

Setting S₂ to position 4 ("Frequency × $\frac{10}{3}$ ") causes the positive input of the voltmeter section to be applied to the slider of VR₁. This potentiometer is adjusted to tap off $\frac{3}{10}$ of the average voltage across VR₁. If, to take an example, S₁ is in position 1 an input frequency of 100 c/s will then cause the meter to give full-scale deflection. The consequence is that, with S₂ in position 4, the range selected by S₁ become 0-100 c/s, 0-1 kc/s, 0-10 kc/s and 0-100 kc/s. The advantage of the multiplying facility offered by VR₁ is that a total of 8 ranges become available with only four preset resistors in the multivibrator. Many commercial measuring instruments use the sequence 1, 3, 10, 30, etc., in their ranges, since this conveniently obviates the difficulty of resolving readings over the first 30% of the scale, and it is possible for the home-constructor to obtain meter movements scaled for both 3 and 10. For instance, the Radiospares Panel Meters type MR26 and MR31 are both scaled 0-3 and 0-10 in volts. These consist of basic movements having an f.s.d. of 100μA.

When function switch S₂ is set to position 2, the positive terminal of the meter connects to the positive supply line via R₁₁. This resistor has a value which enables the meter to read 10 volts f.s.d., whereupon it provides an indication of battery voltage.

Any meter with an f.s.d. of 250μA or less may be used in the M₁ position. The highest average voltage likely to appear across VR₁ is of the order of 3 volts, and this will correspond to 100 c/s on Range 1, and so on. The preset resistors VR₂ to VR₅ should therefore have values which cause an f.s.d. reading in the meter when about 1 volt is applied from S_{2(b)} (either from the lower end of VR₁ for a frequency of, say, 30 c/s at range 1 with S₂ in position 3, or from the tap in VR₁ for a frequency of 100 c/s with S₂ in position 4). For a 250μA meter the final *adjusted* value in VR₂ to VR₅ will be of the order of 4kΩ; for a 200μA meter, 5kΩ; for a 100μA meter, 10kΩ; and for a 50μA

meter, 20k Ω . The actual values chosen for these components should be about twice those just mentioned. The corresponding values in R₁₁ are 40k Ω , 50k Ω , 100k Ω and 200k Ω respectively.

Calibration

The 30 c/s, 300 c/s, 3 kc/s and 30 kc/s ranges are calibrated first, with S₂ in position 3. S₂ is then set to position 4 with, say, a 300 c/s input on range 2. VR₁ is next adjusted for 30% deflection of the meter. This setting will then hold for all the other ranges.

The preset resistors VR₂ to VR₅ are set up with known input frequencies obtained, preferably, from a calibrated signal generator. If this is not available, a non-smoothed full-wave rectified mains signal makes a fair 100 c/s reference. This, in combination with an uncalibrated variable frequency oscillator, could be used for calibration, carrying on calibration points from one range to the next. However, greatest accuracy will be given with the calibrated generator.

Construction

The frequency meter may be built up in any convenient form, the major layout requirement being that short connecting leads are used in the signal and multivibrator circuits and that the stages

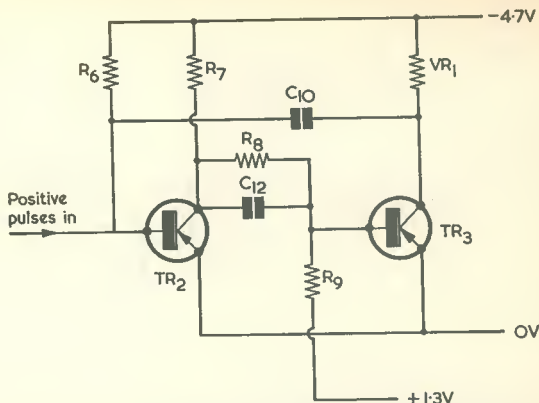


Fig. 3. Simplified version of the monostable multivibrator. This is shown with C₁₀ between the collector of TR₃ and the base of TR₂, with which capacitor the circuit gives pulses of 3 to 4 milliseconds duration

proceed logically in the same order as in the circuit diagram.

Switches S₁ and S₂ may be 3-pole 4-way Yaxley switches of the "wave-change" type. The other parts are specified in the Components List. *



Building Transmitter V.F.O.'s

by

F. G. RAYER G3OGR

Drawing from a wealth of experience in the design and construction of transmitters, our contributor describes a number of v.f.o. circuits which offer excellent practical results. The amateur bands covered are 160 metres (1.8-2.0 Mc/s), 80 metres (3.5-3.8 Mc/s), 40 metres (7.0-7.1 Mc/s), 20 metres (14.0-14.35 Mc/s) and 15 metres (21.0-21.45 Mc/s).

A VARIABLE FREQUENCY OSCILLATOR is often used with a transmitter so that operating is not limited to those frequencies for which crystals are to hand. The v.f.o. circuits described in this article can be built into a home-constructed transmitter, or may be used as separate units with home- or commercially-made transmitters.

If the coil and other information is followed, there should be no difficulty in securing suitable band coverage. Some of the circuits can

employ easily obtained ready-made coils, and suitable tuning ranges can then be expected without the nuisance of winding or modifying home produced coils.

Hartley Oscillator

The Hartley oscillator is much less used than in earlier years, but is still worth considering. Fig. 1 gives a suitable circuit for 3.5-3.8 Mc/s. Coupling from the oscillator tuned coil L₁ is by way of the electron stream in the valve, with

further isolation given by the screen grid. A large output is available. With a single buffer/multiplier stage before the p.a., sufficient drive is available on the 3.5, 7, 14 and 21 Mc/s bands.

L₁ is a Wearite PHF5 coil, which is air-cored and of fixed inductance. The unwanted primary is removed. A lead for the cathode tap is soldered to the tuned winding 3¼ turns from the earthy end, the enamel on the winding wire being scraped away at the appropriate point to

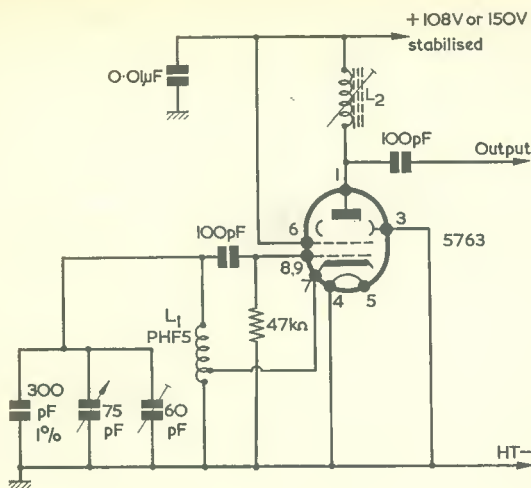


Fig. 1. An electron-coupled Hartley oscillator. In this and subsequent diagrams, close-tolerance capacitors (1 or 2%) should be silver-mica

enable the solder joint to be made. Set the 75pF capacitor with its vanes nearly closed, then adjust the 60pF trimmer until the v.f.o. is heard at 3.5 Mc/s on a receiver. Coverage is then 3.5–3.8 Mc/s with a little to spare.

L_2 depends on the purpose in view. For 3.5–21 Mc/s working, a small core-tuned coil broadly resonant at about 7.1 Mc/s is suitable. This functions as a choke for 3.5 Mc/s and, for higher bands, allows an output to be taken at about 7.1 Mc/s. The core is adjusted for maximum p.a. grid current near the middle of the 14 Mc/s or 21 Mc/s band. (Details of a suitable coil for 7.1 Mc/s operation are given when the circuit of Fig. 3 is discussed.)

A similar circuit can be used for 160 metres. L_1 can then be 35 turns of 26 s.w.g. enamelled wire, the turns being wound side by side on a 1in diameter former, with the cathode tap 8 turns from the earthy end. The fixed capacitor changes to 200pF and the tuning capacitor to 100pF. L_2 can be broadly resonant at about 1.9 Mc/s.

V.F.O. and Operating Frequency

Pulling or frequency modulation effects are less likely when the transmitter output is a multiple of the v.f.o. frequency. For 3.5–3.8 Mc/s, the v.f.o. can tune 1.75–1.9 Mc/s. A 7 Mc/s v.f.o. is often used for the 14, 21 and 28 Mc/s bands, to avoid the need for much frequency multiplication.

Results can still, nevertheless, be

satisfactory with the transmitter output on the v.f.o. frequency, whereupon a 1.75–2.0 Mc/s v.f.o. allows coverage of the 1.8–2.0 Mc/s and 3.5–3.8 Mc/s bands. Should 1.8–2.0 Mc/s be unwanted, a 3.5 Mc/s v.f.o. can be used for operating straight through on the 80 metre band, as well as higher frequency bands. When no frequency multiplication is wanted, the anode circuit, such as L_2 in Fig. 1, should be untuned, or at least not resonant at the working frequency.

If a v.f.o. is added to a crystal controlled transmitter, it is convenient to use the original crystal oscillator stage as a doubler. Instability is then very unlikely. Input to the crystal oscillator grid circuit is thus at half the frequency required at the anode.

Clapp Oscillator

The Clapp oscillator is popular and is capable of high frequency stability. Fig. 2 shows a suitable triode circuit. The large capacitors from grid, cathode and anode circuits swamp changes in capacitance in the valve and valveholder.

For 3.5–3.8 Mc/s, L_1 can be 22 turns of 32 s.w.g. enamelled wire wound side by side on a ½in diameter cored former. Nearly close the variable capacitor vanes, and adjust the core until the v.f.o. is heard on 3.5 Mc/s. The 3.5–3.8 Mc/s band will then be covered.

For 1.8–2.0 Mc/s, L_1 can be 55 turns of 32 s.w.g. enamelled wire, wound side by side on a ½in cored former. The fixed capacitor from coil to chassis is changed to 250pF.

If there is no objection to further adjustments, the total capacitance of 270pF (or 250pF for 1.8–2.0 Mc/s) can be made up from a 200pF fixed component with 100pF trimmer in parallel. It will then be found that adjustments to L_1 core and this trimmer allow some variation in the total band coverage secured by full rotation of the tuning capacitor.

Output from the cathode is generally sufficient to drive an

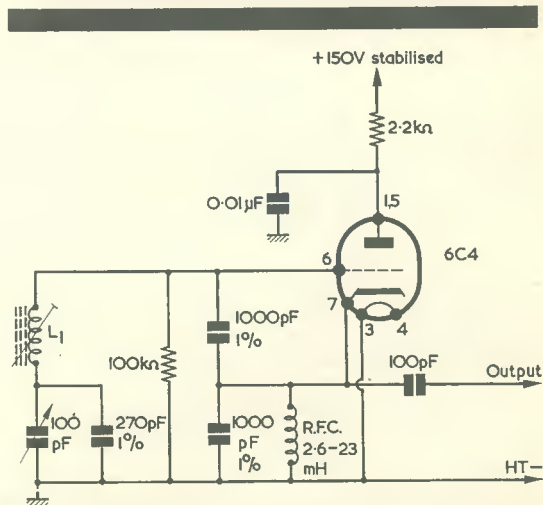


Fig. 2. A triode Clapp oscillator. Satisfactory results have been obtained using either a 2.6mH short-wave choke or a 23mH all-wave choke in the r.f.c. position

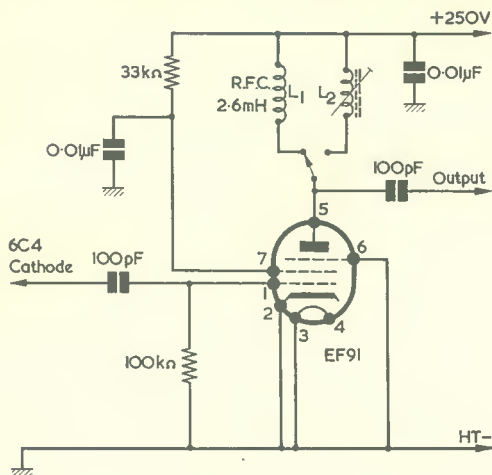


Fig. 3. A buffer-multiplier stage, to follow the v.f.o.

original crystal oscillator stage at fundamental frequency or double fundamental frequency. For higher frequency multiplication, an extra buffer/multiplier will generally be needed.

Fig. 3 shows a buffer/multiplier circuit suitable for use with Fig. 2 and other v.f.o. circuits. L_1 is an r.f. choke, and is selected when the amplifier is working straight through at the same frequency. L_2 is tuned to about 7.1 Mc/s, and is in circuit when the 3.5 Mc/s v.f.o. is driving the p.a. for 14 Mc/s and higher frequency bands. L_2 can have 34 turns of 32 s.w.g. enamelled wire, wound side by side on a $\frac{1}{2}$ in diameter cored former. The core is adjusted for maximum p.a. grid current at about the centre of one of the h.f. bands (say 14.2 Mc/s).

Double-Triode V.F.O.

The circuit in Fig. 4 uses a 12AT7 as Clapp v.f.o., cathode coupled to a buffer-amplifier-doubler. This gives excellent isolation of the v.f.o. from following stages, and a relatively large output.

For 1.8–2.0 Mc/s, L_1 can be 95 turns of 34 s.w.g. enamelled wire wound side by side on a $\frac{1}{2}$ in diameter cored former. Set the 75pF tuning capacitor vanes half closed, and adjust L_1 core until the v.f.o. is heard on 1.9 Mc/s.

For working straight through without multiplication, the r.f. choke shown is satisfactory. But if a higher frequency band is to be worked (e.g. 3.5 Mc/s) a broadly tuned coil, resonant at twice the

frequency of L_1 , may be substituted. This will increase output while multiplying. The cathode bias of the second section is essential for full output.

As with the other circuits, the h.t. voltage can be varied to some extent, according to the supply available, or output wanted. The v.f.o. should generally be operated at a fairly low voltage. If maximum output is wanted, the second triode section can receive 250V.

Band Coverage

Exact capacitor values are quoted in the circuits given here so that

the wanted band can be found and covered immediately. In all cases there is a little unused rotation of the tuning capacitor towards the maximum and minimum settings. This allows calibration at 100 kc/s points—for example, at 3.5, 3.6, 3.7 and 3.8 Mc/s—and filling in between these marks with equal divisions when using a capacitor with semi-circular vanes. It is preferred that a band completely fill the 180-degree rotation of the variable capacitor, then a vane or so should be taken off at the rear. Check coverage at both ends of the band before removing a further vane.

Probably the simplest, and most accurate method of checking coverage and calibrating the v.f.o. dial is to use a 100 kc/s crystal marker in conjunction with a receiver. This will supply 0.1 Mc/s calibration points. Should 50 kc/s (0.05 Mc/s) calibration points be wanted, calibrate the 100 kc/s points first, then tune the receiver to the second harmonic of the v.f.o. The 50 kc/s points can then be found against the 100 kc/s pips of the crystal marker.

Coils

Buffer and multiplier anode coils can be of normal type, as no particular frequency stability is required here. The actual v.f.o. tuning coil should not, however, be subject to random changes in inductance. Some commercially made variable frequency oscillators use small coils resembling those found in receivers; others have baked and impregnated coils.

(continued on page 487)

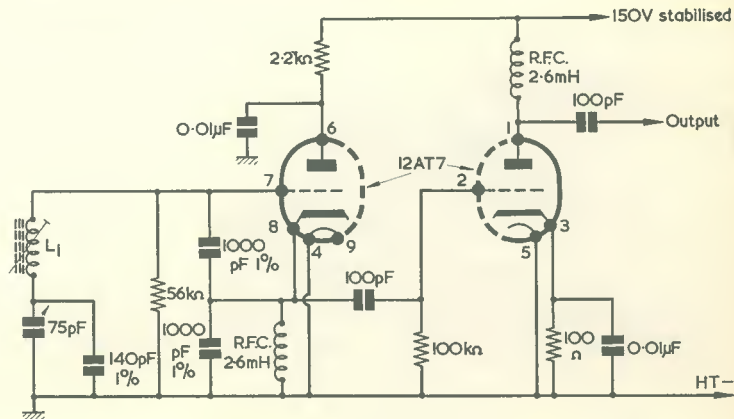
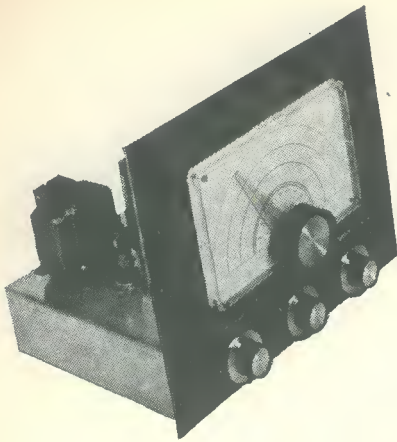


Fig. 4. A double-triode employed as a v.f.o./buffer-amplifier



THE "8-2 SP

1.67



ECF82 ★ ECC

The "8-2 Special" short-wave receiver has been designed specifically for the beginner who wishes to start on the right track—with t.r.f. construction—before crossing the line to superhet building with its inherent tracking problems. As a comparatively simple design, it signals a clear way ahead for ease of construction and the author expresses the view that the receiver brings home the goods with regard to first class results, particularly in respect of Top Band Dx operation

THE "8-2 SPECIAL" SHORT-WAVE RECEIVER HAS been specifically designed with the beginner in mind and it may be constructed in the first instance as a 2-valve design by omitting the r.f. stage given by V_1 , adding this valve later as time and available cash permit. The design itself is simple with no frills or "gimmick" circuits and is very easy to get working. The completed 3-valve receiver performs very well for it is, as may be seen from the circuit diagram of Fig. 1, really a 6-valve design, each glass envelope consisting of (in effect) two valves. On Top Band (1.8 to 2.0 Mc/s) c.w. or Morse transmissions have been logged from all over the British Isles—GM, GW—and several from

the Continent. On all the other bands, both amateur and broadcast, the results have been equally successful. With the coils specified, the overall coverage is 1.67 Mc/s to slightly higher than 31.5 Mc/s (180 to 9.5 metres).

The foregoing results were obtained with an outside inverted "L" long wire aerial, about 30 feet high, oriented roughly 12 degrees North (North at the end remote from the receiver) of a line running directly from East to West. If possible, the constructor should use a similar outside aerial. An important factor for the beginner to bear in mind is the receiver earth connection. This should preferably be taken to earth itself. Connection to a water pipe often gives inferior results, and it has also to be remembered that in modern houses the piping may be of the plastic type. The earth wire should be of a thicker gauge than the aerial wire and should be as short and direct as possible, terminating in a metal rod (copper is preferable) driven as far into the earth as possible. A good earth connection is all-important with a receiver of this type.

Circuit

From Fig. 1 it will be seen that, as was just mentioned above, each of the three valves performs a double function. Also each has a type coding with the suffix 82; hence the title of this article—the "8-2 Special". The majority of t.r.f. designs tend to be of the three valve type but in this case it was decided to make each of the valves a "double type". The result is a six stage design, bringing the final performance close to that of a communications class receiver.

"SPECIAL" RECEIVER



Cover Feature

—32 Mc/s



ECF82 ★ ECL82

With the three valve circuit shown in Fig. 1, the main troubles associated with t.r.f. designs—radiation interference by an oscillating detector stage, difficult reaction or “dead spots”, and lack of selectivity have been very largely overcome. Provided the r.f. valve V_1 , is included, radiation interference by an oscillating detector stage is eliminated. In addition, reaction is nearly constant (i.e. the oscillation point remains almost always at the same setting of the reaction control C_7) and dead spots are completely absent, whilst selectivity is adequate bearing in mind the fact that this is a t.r.f. design.

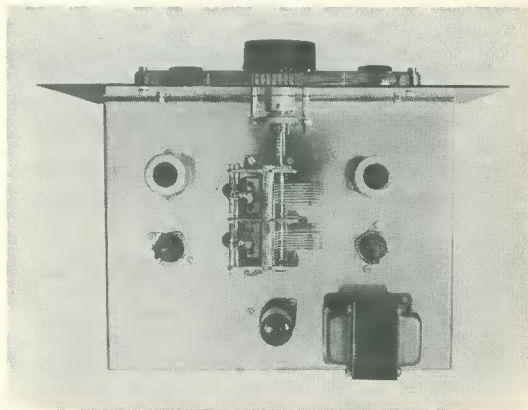
The r.f. stage employs an ECF82 triode-pentode frequency changer, this being a type for use in television receivers. It performs here as firstly ($V_{1(a)}$) a grounded grid triode presenting a low impedance input to the aerial. This circuit isolates the tuned r.f. pentode, $V_{1(b)}$, from the aerial, thereby removing all damping effects from the tuned grid circuit of the pentode and providing a high degree of selectivity with respect to the r.f. stage as a whole.

The aerial input is, via the capacitor C_1 to the cathode (pin 8) of $V_{1(a)}$, the resistor R_1 providing a small amount of bias for this stage. The grid is grounded direct to chassis and the output from this isolating stage is taken to the top of the coil L_1 via the capacitor C_2 . The winding L_1 is, of course, that which would normally be used in conventional r.f. stages for aerial coupling.

The winding L_2 is tuned by $C_{4(a)}$ this being one half of a two-gang variable capacitor, the other half of which ($C_{4(b)}$) tunes the grid of the detector stage. The cathode of the pentode section obtains a bias

voltage via the components R_2, R_3 and C_3 , it being made variable by means of the potentiometer R_3 . This is the r.f. gain control, and it will be found capable of taking the stage from minimum to near maximum gain without unwanted oscillation taking place. Screen-grid voltage (pin 3) is supplied via the resistor R_6 , the screen-grid being bypassed to chassis by C_5 . The anode h.t. supply is via R_5 and RFC_2 . The r.f. signal tuned by the combination $C_{4(a)}, L_2$, is presented to the pentode grid (pin 2), amplified and passed via the coupling capacitor C_6 to the top of winding L_3 .

Winding L_3 is that normally associated with the aerial input of a two-valve design and it is to this



Above—chassis view showing layout of main components

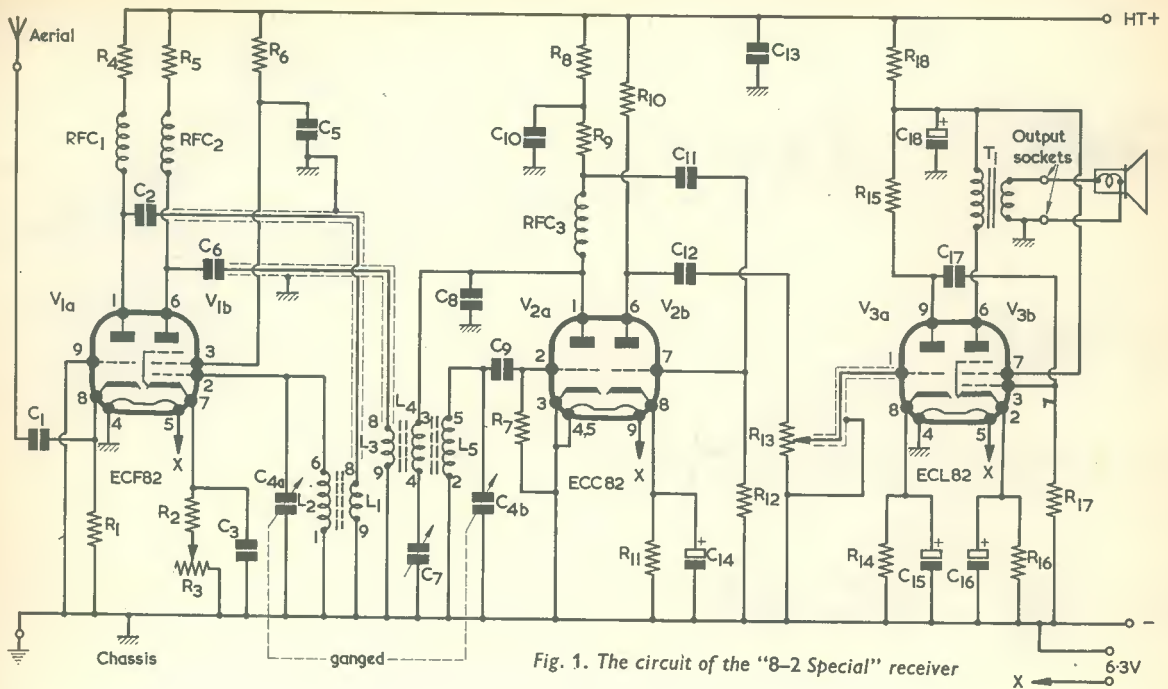


Fig. 1. The circuit of the "8-2 Special" receiver

X ← 6.3V

point, should V_1 be initially omitted from the design, that the aerial should be connected if a two-valve design is preferred in the first instance. Alternatively, the beginner could leave the connection from C_6 to tag 8 of L_3 unsoldered in the first instance so that, when testing the receiver, the aerial could be connected to this point. This would ensure that the circuit around V_2 and V_3 is working correctly before inserting V_1 into its valveholder.

It will be noted that, in this design, the pentode couples to the detector tuned circuit by way of winding L_3 , and that this ensures that maximum selectivity is obtained. In some designs, connection from the r.f. stage is made via a small-value capacitor direct to the grid circuit of the detector stage. Similarly, in some designs, a capacitive coupling might be used to the grid tuned circuit of $V_1(b)$. These methods were not considered advisable for the present receiver, where one of the main aims is to obtain better selectivity than is given by the more usual class of t.r.f. receiver.

The amplified r.f. signal applied to L_3 is coupled into winding L_5 , this being tuned by $C_4(b)$. The signal is then applied to the grid of the detector, $V_2(a)$.

The detector and first audio amplifying stages are constructed around an ECC82 double-triode, $V_2(a)$ being the detector and $V_2(b)$ the audio stage. The components C_9 and R_7 are the grid capacitor and grid leak respectively and these, together with the other components in the detector stage, ensure that reaction is smooth and positive with no "backlash" or overlap. It is necessary for this stage to oscillate

when receiving c.w. signals. When reaction is just short of the oscillation point, sensitivity and selectivity are at a maximum. The feedback winding necessary to obtain reaction is given by L_4 , the variable capacitor C_7 being the reaction control. The capacitor C_8 , connected from the anode of $V_2(a)$ to chassis, ensures that reaction is "silky smooth" with no trace of "ploppy" results or howl. The anode decoupling components are R_8 and C_{10} , the h.t. being applied via R_9 and RFC_3 . The detected signal is fed to the first audio stage via the coupling capacitor C_{11} . The components R_{11} and C_{14} supply cathode bias to this stage.

The audio signal from $V_2(b)$ is tapped off by the slider of the a.f. gain control, R_{13} , and is applied to the grid of the following stage $V_3(a)$, this being the second audio amplifier.

The second audio stage and the power amplifier stage use an ECC82, $V_3(a)$ being the triode amplifier and $V_3(b)$ the pentode power output stage.

The amplified audio signal from the anode of $V_3(a)$ is passed via the coupling capacitor C_{17} to the grid of $V_3(b)$, R_{14} and C_{15} providing cathode bias for the triode and R_{16} , C_{16} for the pentode section. The whole of this stage is decoupled from the h.t. line via the 5 watt resistor R_{18} , R_{15} supplying h.t. voltage to the anode of the triode, and the speaker transformer primary winding supplying h.t. voltage to the anode of the pentode section. The electrolytic capacitor C_{18} is the h.t. bypass component to chassis.

It will be seen that the receiver has a considerable amount of a.f. gain. In most instances it may be

operated with the a.f. gain control slider set fairly well back at the minimum volume end of its track. The high a.f. gain ensures that there is plenty of amplification in reserve for the reception of very weak signals. Due to the amplification provided it is possible that some ECC82's may exhibit a small level of microphony, this giving rise to a "ping" in the speaker when the valve, or the chassis adjacent to the valve, is tapped. This does not represent a fault symptom and will not prevent operation of the receiver in the normal manner. It would be preferable not to attempt to mount the speaker in the same cabinet as the receiver chassis, since acoustic feedback may be troublesome if a particularly

microphonic ECC82 were used in the V_2 position. In the circuit as shown, no provision has been made for headphone operation by the inclusion of a headphone jack. If the constructor particularly desires to use headphones, a jack socket may be fitted to the receiver at any convenient point away from the V_1 and V_2 stages, this coupling to the anode of $V_{3(a)}$ (pin 9) by way of an $0.05\mu\text{F}$ capacitor and screened cable, as illustrated in Fig. 2. The audio gain control should normally be kept at a low setting when using headphones because of the high a.f. amplification in the receiver, and because an uncomfortably high output may otherwise be passed to the phones when tuning through a powerful

COMPONENTS

Resistors

(All fixed values $\frac{1}{2}$ watt 10% unless otherwise stated)

R ₁	330 Ω
R ₂	100 Ω
R ₃	5k Ω pot. 1in.
R ₄	33k Ω
R ₅	5.6k Ω
R ₆	100k Ω
R ₇	1.5M Ω
R ₈	100k Ω
R ₉	10k Ω
R ₁₀	33k Ω
R ₁₁	1k Ω
R ₁₂	470k Ω
R ₁₃	500k Ω pot. log
R ₁₄	2.2k Ω
R ₁₅	220k Ω
R ₁₆	470 Ω
R ₁₇	680k Ω
R ₁₈	1.5k Ω 5 watt

Capacitors

C ₁	0.001 μF silver mica or ceramic
C ₂	100pF silver mica
C ₃	0.01 μF tubular, (Mullard)
C _{4a,b}	310+310pF, variable, 2-gang, type E (Jackson Bros Ltd)
C ₅	0.01 μF tubular, (Mullard)
C ₆	100pF silver mica
C ₇	100pF variable, type C804 (Jackson Bros Ltd)
C ₈	200pF silver mica
C ₉	100pF silver mica
C ₁₀	0.1 μF tubular, (Mullard)
C ₁₁	0.01 μF tubular, (Mullard)
C ₁₂	0.01 μF tubular, (Mullard)
C ₁₃	0.1 μF tubular, (Mullard)
C ₁₄	12 μF electrolytic, 6V wkg.
C ₁₅	12 μF electrolytic, 6V wkg.
C ₁₆	25 μF electrolytic, 25V wkg.
C ₁₇	0.01 μF tubular, (Mullard)
C ₁₈	8 μF , electrolytic, 350V wkg.

Output Transformer

T₁ Type 117E (H. L. Smith & Co Ltd)

Speaker

7 × 4in elliptical, 3 Ω

R.F. Chokes

RFC_{1,2,3} 2.5mH, type CH1
(H. L. Smith & Co, Ltd)

Dial & Drive

Electroniques (STC Ltd), type SMD2.
(Home Radio Cat. No: DL6)

Valves

V ₁	ECF82 (Mullard)
V ₂	ECC82 (Mullard)
V ₃	ECL82 (Mullard)

Valveholders

5 off B9A with centre spigots (2 with skirt & can for V₁ and V₂)

Tagstrips

7-way (3 tags earthed)
5-way (2 tags earthed)

Knobs

3 off (H. L. Smith & Co. Ltd)

Coils

Miniature Dual-Purpose,
Blue (L₁, L₂) and
Green (L₃, L₄, L₅).
Ranges 3, 4 and 5. (Denco Ltd.)

Chassis

9 × 7 × 2in (H. L. Smith & Co. Ltd)

Panel

10 × 8in (H. L. Smith & Co, Ltd)

Miscellaneous

2BA and 4BA nuts and bolts; aerial/earth and speaker output socket strips each fitted with plugs; p.v.c. wire; rubber grommets (1 off $\frac{1}{2}$ in, 4 off $\frac{1}{4}$ in) coaxial cable, solder, etc.

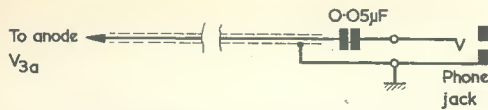
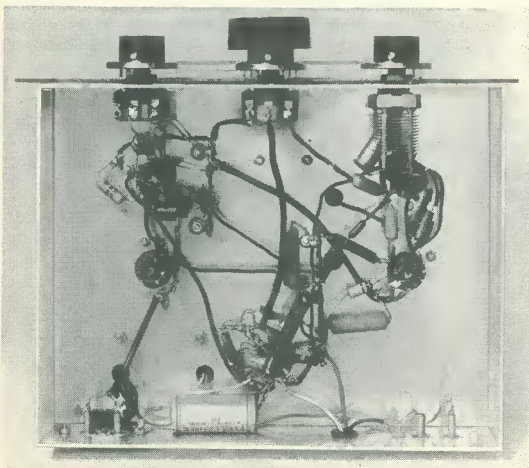


Fig. 2. If it is desired to use headphones, a phonejack may be added, as shown here

signal. An important point to note is that the receiver should not be switched on without a suitable load (such as the speaker) connected across the output sockets. Without such a load, high a.f. voltages may be developed across the primary of the output transformer with a possible risk of damage. If it is intended to disconnect the speaker when the receiver is switched on (so that, for instance, headphones may be used on their own) a 10Ω 1 watt resistor should be permanently connected across the output sockets inside the chassis. This will provide a permanent load of sufficiently low value, and will not noticeably affect speaker matching.

Power Supply

The power supply for this receiver is given by the power pack described in the December 1966 issue of this magazine and the writer has been informed that back numbers of that issue are still available for those who have not seen a copy. However, for those who wish to incorporate a power supply on the same chassis as the receiver, a suitable circuit is shown in Fig. 5. With the prototype the power supply is, of course, a separate unit, the writer favouring this method for the reasons that it considerably assists with the reduction of mains hum and that the separate power supply may be used for other items of equipment if so desired. The power supply featured in the December 1966 issue is fitted with a voltmeter and a milliammeter and this may not be required by all beginners—hence the circuit shown here. Any power supply, provided it is fully isolated from the mains by a mains transformer, and incor-



Below—chassis wiring of the "8-2 Special"

porates a full-wave rectifier and adequate smoothing components, may be used. The ratings should be 250 volts at 50mA and 6.3 volts at 2A. When the power supply described in the previous issue is employed, ensure that the 6.3 volt wiring is such that both receiver and power unit chassis are at the same potential.

Construction

Fig. 3 (a) shows the plan view of the chassis deck, the drilling positions for V_1 , V_2 , V_3 , L_1 , L_2 and L_3 , L_4 , L_5 being clearly set out alongside the respective dimensions. Measurements should be made on the chassis deck and the drilling points marked with a pencilled cross such that, if a mistake has been made, it may be removed by rubbing lightly with a finger. Having ensured that the correct markings have been made, drill the holes firstly with a $\frac{3}{16}$ in. drill, enlarge with a $\frac{1}{8}$ in. drill inserted into a handbrace and then cut the valveholders for the valves and coils with a B9A chassis cutter. It is an advantage in this type of work—"chassis bashing" as it is termed—to have at hand a self-tapping centre punch with which to mark the drilling points prior to the actual drilling operation. In this manner the drill will not "wander" when applied to the chassis deck.

Having cut the five holes, position V_1 valveholder such that pins 1 and 9 are nearest the centre of the chassis; V_2 , pins 1 and 9 nearest the front chassis apron; V_3 , pins 1 and 9 nearest the centre of the chassis; the valveholder for $L_{1,2}$, pins 1 and 9 nearest the rear chassis apron; and the valveholder for $L_{3,4,5}$, pins 1 and 9 nearest the centre of the chassis. Mark and drill the mounting holes with a $\frac{3}{16}$ in. drill. Fit all five valveholders with an earth tag each, under one of the securing nuts.

Two tagstrips are used with this receiver and they should be positioned as shown in the below-chassis illustration. The 7-way tagstrip lies midway between V_3 and $L_{3,4,5}$. This tagstrip has 3 tags earthed to chassis, and the nut and bolt securing the tag nearest the front of the chassis also secures the rear fixing lug of the two-gang tuning capacitor. After the tuning capacitor has been fitted (as will be described shortly) the 7-way tagstrip is mounted in position, using 4BA nuts and bolts at each of the remaining earth tags.

The 5-way tagstrip is secured to the chassis by means of 4BA nuts and bolts, both earth tags being bolted into position. This tagstrip is shown alongside V_1 valveholder.

For ease of construction, the rear chassis apron should next be dealt with, and this is shown in Fig. 3 (c). A $\frac{1}{2}$ in. rubber grommet is fitted to this apron for the power input leads. Drill and fit the grommet. To the left of the grommet fit the speaker output Paxolin socket strip—no dimensions are given here because these components are often of differing sizes. Fit also the socket strip for the aerial and earth connections, as shown in the photograph. These strips are secured to the chassis rear apron by means of two 6BA nuts and bolts, an earth tag being fitted under the nut of each strip nearest the centre of the apron. The sockets proper of each tagstrip

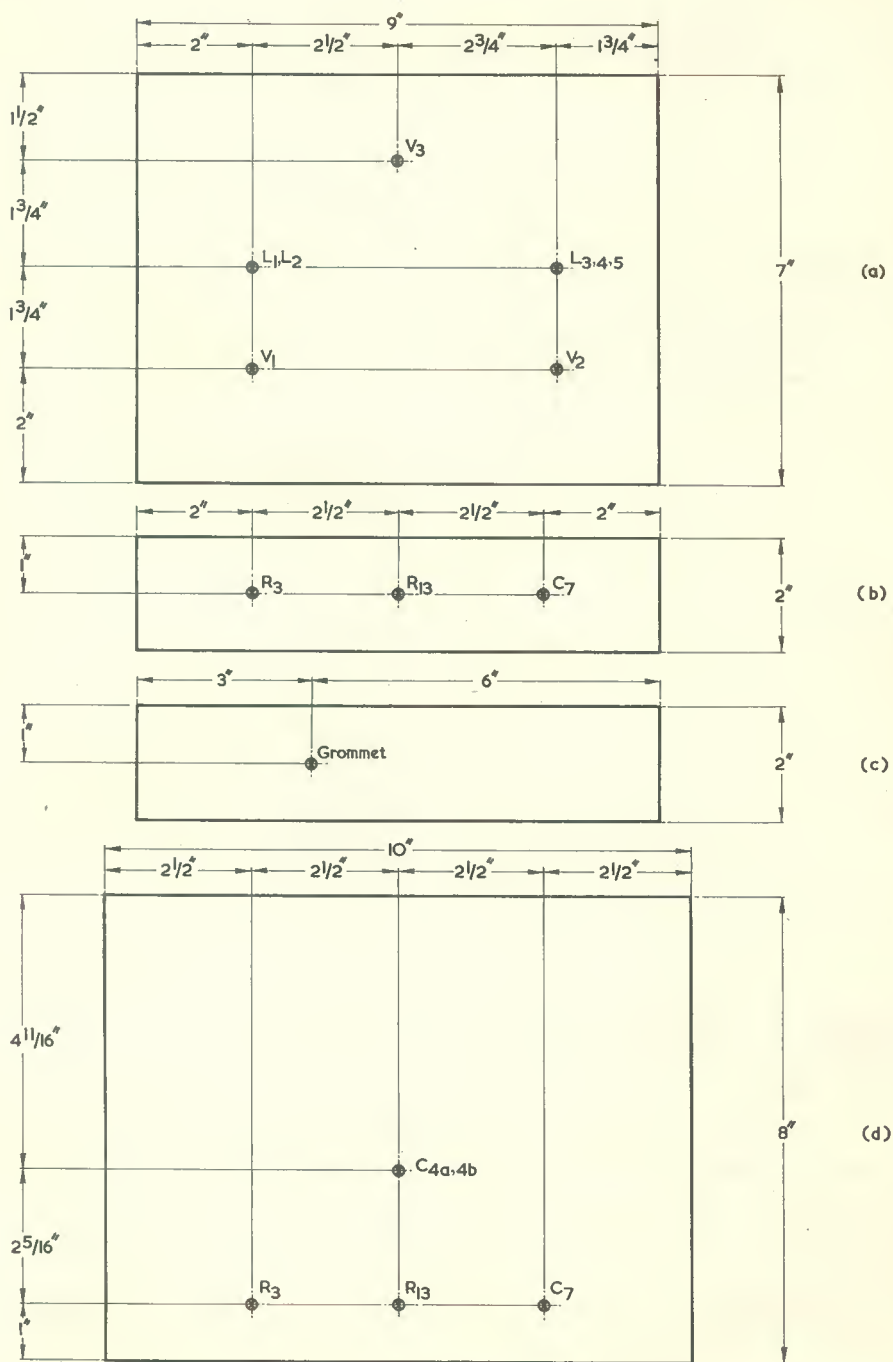


Fig. 3. Drilling measurements for (a) the chassis deck, (b) chassis front apron, (c) chassis rear apron (see text), and (d) the front panel. The hole for C_{4(a)(b)} in (d) is marked out with the aid of the tuning capacitor itself

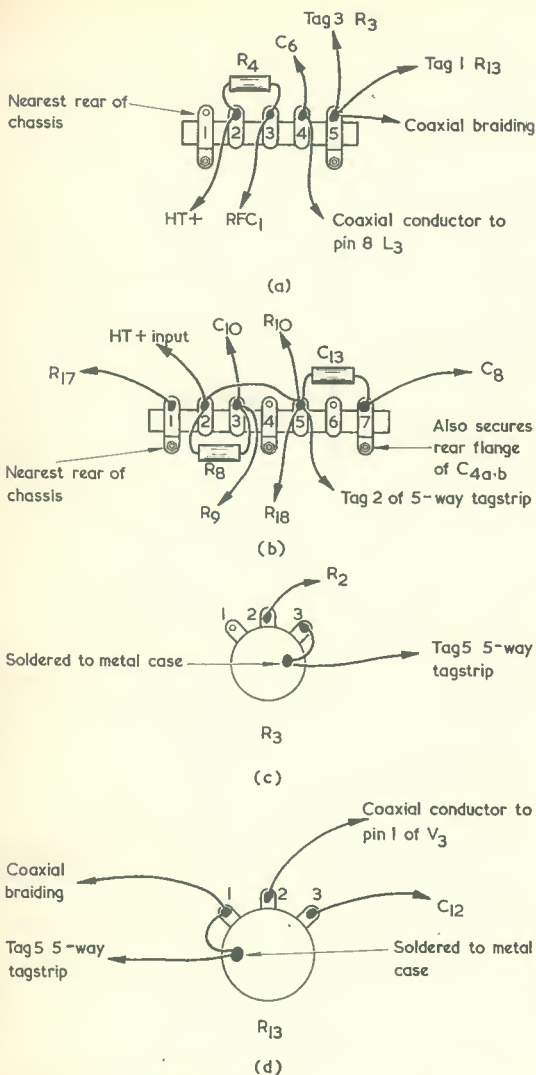


Fig. 4. Wiring-up details of (a) the 5-way tagstrip, (b) the 7-way tagstrip, (c) the r.f. gain control R_3 , and (d) the a.f. control R_{13} . Both (c) and (d) are rear views

should protrude through $\frac{3}{8}$ in. holes, and therefore a total of four such holes must be drilled. Take care to ensure that none of the sockets are short-circuiting to chassis at the $\frac{3}{8}$ in. holes.

Before proceeding with the last of the chassis

drilling work, obtain the front panel and mark and drill the three lower holes as shown in Fig. 3 (d). Drill the holes firstly with a $\frac{1}{8}$ in. drill and enlarge with a $\frac{3}{8}$ in. drill as previously described.

The next task is to hold or clamp the panel to the front apron of the chassis in the position that it will finally occupy, in which $\frac{1}{2}$ in. of panel extends on each side of the chassis. Carefully mark the centres of the three holes in the front apron and drill to $\frac{3}{8}$ in. diameter. The panel and chassis holes should now coincide when R_3 , R_{13} and C_7 are fitted and secured to the assembly. Note that, with C_7 , both nuts should be removed, only one being replaced to secure this component to the panel and chassis.

Before finally fitting these three components, however, the tuning capacitor, $C_{4(a),(b)}$, should be placed on the chassis deck such that the spindle end is against the rear of the front panel when the latter is temporarily in place. Mark the spindle position on the rear of the panel and drill to $\frac{3}{8}$ in. diameter, following this by cutting a hole with the B9A chassis cutter. Fit the panel with the dial drive mechanism as shown in the manufacturer's literature. Secure the panel to the chassis by fitting the three lower control R_3 , R_{13} and C_7 into position. Next, place the tuning capacitor on the chassis deck and mark and drill ($\frac{1}{8}$ in) its three fixing holes after the capacitor spindle has been inserted into the drive mechanism and temporarily secured. Using the rear bolt also for securing the appropriate earth tag (tag 7—see Fig. 4 (b)) of the 7-way tagstrip, drill the other two holes for this tagstrip and secure both this and the capacitor into position. Referring to the literature supplied with the dial, use the template provided and drill the holes required for fitting the actual dial and cover to the front panel.

Alongside each tag, of the tuning capacitor nearest the chassis, drill a $\frac{1}{4}$ in hole and fit rubber grommets to these holes. These will carry the fixed vane connections from the tuning capacitor through the chassis deck.

Place the output transformer T_1 on the chassis deck in the position shown in the above-chassis illustration and drill the four holes required to secure this component into position ($\frac{3}{8}$ in) by using the transformer lugs as a template. Mark and drill a further two holes ($\frac{1}{4}$ in), and fit each with a rubber grommet. Position T_1 such that the white and orange wires protrude through the chassis nearest V_3 . Secure T_1 into position by means of 2BA nuts and bolts.

The mechanical assembly is now complete and wiring-up may commence.

(to be concluded)

NEW CATALOGUE

Arthur Sallis Radio Control Ltd., of 93 North Road, Brighton, Sussex, have just issued their new illustrated Catalogue No. 17. A large selection of new miniature components, semiconductors, Government and manufacturers surplus electronic equipment are listed and the price of this informative publication is 3s. post free from the above address.

Building Transmitter V.F.O's

(continued from page 479)

The coil should be wound tightly, with turns cemented in place. It is best not to fit it in a closely fitting screen, nor unnecessarily near a valve or other source of heat.

If the coil core is adjustable, this should be sealed afterwards. Should an adjustable inductance be felt troublesome, a coil of fixed inductance can usually be substituted. A pre-set capacitor is then required. For example, L_1 in Fig. 2 may be a Wearite PHF6 coil with the primary removed, and 31 turns unwound from the grid end of the secondary. In this case the variable capacitor is 80pF, and the parallel capacitance made up from a 200pF 1% fixed capacitor and a 30pF trimmer in parallel. Coverage is then 3.5–3.8 Mc/s, with a little to spare at band-edges.

For an air cored coil, a ceramic former, wound with wire under tension, is considered helpful. But with ordinary coils of the type described, frequency drift is likely to be less than can be observed by visual dial readings.

RF Chokes

Suitable choke inductances are indicated in the circuits, but others are often satisfactory in practice. Current in v.f.o. and buffer stages is small, so that receiver type chokes are suitable. The choke should be efficient at the required frequency. In the circuits described, sectionalised short wave 2.6mH chokes, and even 23mH all-wave chokes, have been used. In some equipment where space was limited, miniature 5mH and similar chokes (as often fitted in transistor receivers) have been employed with no apparent change in results. A 2.5mH choke can of course be used where a 2.6mH component is referred to, and typical of the components which may be employed are the Electroniques CCC7 (2.5mH) the Denco RFC.5 (2.6mH) and the Osmor QC1 (23mH).

Temperature Compensation

Temperature compensation may be attempted by a selection of positive and negative temperature coefficient capacitors. The temperatures found in different parts of the v.f.o. are however, likely to vary in a dissimilar manner after switching on,

with the result that anything approaching complete temperature compensation is very difficult to achieve.

Drift can be heard by beating the v.f.o. with a crystal oscillator or marker output on a receiver. It is likely to be relatively fast when the v.f.o. is switched on, settling down after fifteen or twenty minutes. Commercially manufactured equipment usually shows a similar drift.

If the v.f.o. is soundly built, with the valve and other sources of heat clear of coil and capacitors, a circuit having no temperature compensation can be satisfactory for the bands from 1.8 Mc/s to 28 Mc/s.

Stability

A common fault with a v.f.o. is lack of stability, this showing up as frequency modulation. Several causes may be responsible. The trouble is probable if the v.f.o. is operating on the transmitter output frequency, and has insufficient screening and isolation. To avoid this, screen the v.f.o. completely. A good method of construction is to use a small box or four-sided chassis to contain *all* the v.f.o. components except the valve, whose holder is mounted on the box top. A screening is fitted to the valve-holder. Simply improving v.f.o. screening, with no other change whatever, may remove frequency modulation troubles.

At least one buffer or buffer-amplifier-multiplier should be present between the v.f.o. and p.a. The h.t. supply to the v.f.o. should also be regulated, or obtained from a separate supply, so that the voltage does not fluctuate with modulation.

Freedom from unwanted back-coupling effects to the v.f.o. can readily be checked by operating the transmitter into an artificial load (such as a lamp) and switching on the receiver b.f.o. Tune the receiver to zero beat with the transmitter output. Adjustment of the grid or anode tuning of the p.a., or aerial loading capacitor, should not cause shifts in frequency, with consequent heterodynes in the receiver.

Frequency modulation can usually be traced to one of the following causes: inadequate v.f.o. screening, poor regulation of the v.f.o. h.t. supply, having the v.f.o. on the same frequency as the p.a. stage, or not employing at least one buffer between the v.f.o. and the p.a.

Multi-Band Coverage

When the v.f.o. has a single range, coverage of the h.f. bands grows cramped. This fact does not reduce efficiency, but it is worthwhile opening out the scale if the h.f. bands are used frequently.

If the v.f.o. tunes 3.5–3.8 Mc/s, the 3.5–3.6 Mc/s sector will cover 7–7.2 Mc/s, 14–14.4 Mc/s, and 21–21.6 Mc/s when followed by buffer-multipliers. So if the v.f.o. can tune 3.5–3.8 Mc/s for 80 metres, and 3.5–3.6 Mc/s for the 40, 20 and 15 metre bands, this gives a more convenient coverage.

Fig. 5 shows two ways of altering v.f.o. range. Each employs a 2-way switch. In Fig. 5 (a), with the switch

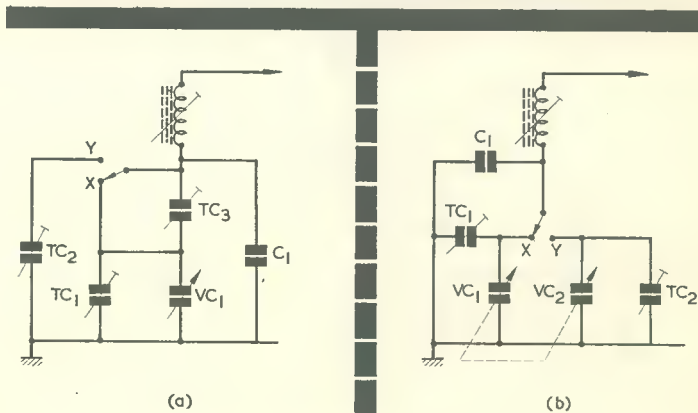


Fig. 5. V.F.O. tuned circuits with switched coverage

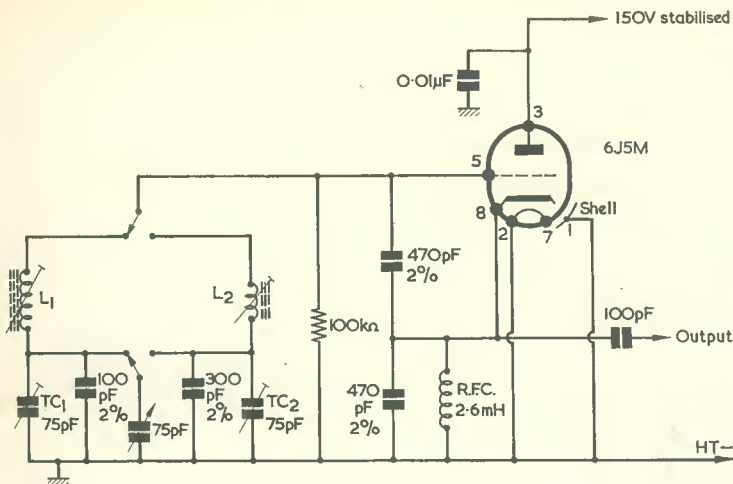


Fig. 6. A 2-band v.f.o. intended for a transmitter operating on the 3.5, 7, 14 and 21 Mc/s bands. The two switches shown form a single d.p.d.t. switch

at X, C_1 in conjunction with VC_1 tunes 3.5–3.8 Mc/s, TC_1 being the trimmer. Capacitor values can be as mentioned earlier. When the switch is at Y, the series capacitor TC_3 reduces the effective value of VC_1 , while TC_2 compensates for the reduction in effective capacitance of TC_1 and VC_1 . Adjust TC_2 and TC_3 for 3.5–3.6 Mc/s coverage.

The circuit in Fig. 5 (b) employs a 2-gang capacitor, and gives very

satisfactory results. C_1 is present on both switch positions. With the switch at X, VC_1 with TC_1 covers 3.5–3.8 Mc/s. Values can be as for a single range v.f.o. When the switch is at Y, VC_2 and TC_2 are in use. VC_2 has nearly two-thirds of its vanes removed, and TC_2 is larger so that 3.5 Mc/s is reached with VC_2 vanes closed, and 3.6 Mc/s with VC_2 vanes open.*

With these circuits, each band is

marked on the v.f.o. scale. The 3.5–3.8 Mc/s band is individually calibrated. Frequencies on the 7, 14 and 21 Mc/s bands are multiples of those on the 3.5–3.6 Mc/s band. There is good bandsread, with easy calibration and very few additional components in the v.f.o. circuit.

Two-Band V.F.O.

Constructing the v.f.o. to give an output on two bands, offers several advantages. A useful circuit of this kind is shown in Fig. 6. A 2-pole 2-way switch selects L_1 or L_2 , with trimmers TC_1 or TC_2 . L_1 tunes 1.75–1.9 Mc/s, and the output may be doubled for operating in the 3.5–3.8 Mc/s band.

When L_2 is in use, coverage is from 7–7.2 Mc/s. This allows ample drive without much frequency multiplication for 14 and 21 Mc/s bands, and straight through working for the 7 Mc/s band.

By adjusting L_1 inductance with TC_1 , and L_2 with TC_2 , band coverage can be modified to achieve a

* In Fig. 5 (a) the values of TC_2 and TC_3 depend upon the values of TC_1 , VC_1 and C_1 , and may be found experimentally. In Fig. 5 (b), TC_2 should have a maximum value equal to or greater than TC_1 plus two-thirds of VC_1 . Initially adjust TC_2 so that VC_2 , without vanes removed, tunes to 3.6 Mc/s near the minimum capacitance end of its range. Then remove vanes to achieve the 3.5–3.6 Mc/s range required. This approach obviates the risk of removing too many vanes from VC_2 .—EDITOR.

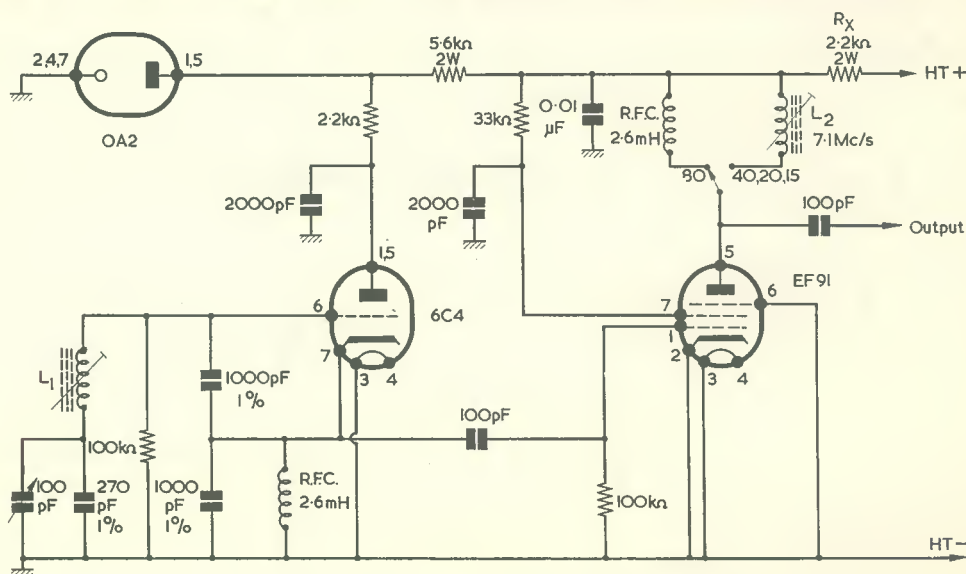


Fig. 7. A v.f.o. with regulator and buffer-amplifier

good spread of frequencies on both scales. Each band is individually calibrated, after both have been set up. The 7 Mc/s and higher frequency bands are on the same scale.

In Fig. 6, L_1 consists of 55 turns of 32 s.w.g. enamelled wire occupying $\frac{1}{4}$ in on a $\frac{1}{2}$ in cored former. L_2 consists of 12 turns of 24 s.w.g. enamelled wire wound side by side on a $\frac{1}{2}$ in cored former.

2-Stage V.F.O.

Fig. 7 shows a circuit with voltage regulator, and buffer-multiplier-amplifier stage. L_1 consists of 22 turns of 32 s.w.g. enamelled wire wound side by side on a $\frac{1}{2}$ in diameter cored former, or of any other coil giving 3.5–3.8 Mc/s coverage. For this band the r.f. choke forms the anode circuit of the EF91. With some equipment, this untuned choke may also prove to be more suitable for the 7 Mc/s band. For increased

output on the higher frequencies L_2 is switched in. It can have about 34 turns, wound with same wire and on the same size former as L_1 . Its core is adjusted by observing p.a. grid current.

The h.t. line resistors are for a 275–300V supply. Resistor R_X may be reduced in value or omitted for a 250V or similar supply.

Drive Obtained

A guide to the results obtained with the circuits described may be useful. If the circuit of Fig. 1 is followed by a 5763 buffer-multiplier, sufficient drive for an 807 or 6146 can be obtained on 3.5–21 Mc/s bands. In these circumstances, Fig. 2 provides enough drive on the fundamental and second harmonic and, with the additional buffer-amplifier in Fig. 7, sufficient drive for 3.5–14 Mc/s bands.

Fig. 4, followed by a buffer-multiplier, will allow three bands

to be worked. For a simple type top band transmitter, Fig. 4 can drive the p.a. directly. Fig. 6 gives increased 7 Mc/s output, and may be used for all bands when followed by two buffer-amplifiers.

Maximum drive is obtained when the v.f.o. is built into the transmitter. If it is separate, a screened co-axial cable can take the v.f.o. output to the original crystal controlled stage, but this cable reduces the drive obtained.

Editor's Note

The Wearite PHF5 and PHF6 coils referred to in this article are r.f. coupling coils having tuned winding inductances of 5.5 and 37.5 μ H respectively. They are available from Home Radio (Mitcham) Ltd. under Cat No. CO84/5 for the PHF5 coil and CO84/6 for the PHF6 coil.



Transistorised 2-Tone Alarm Generator

(continued from page 460)

board panel, thereby eliminating the need for etching processes while retaining all of the advantages of printed circuit construction.

Start construction by cutting the Veroboard panel to size, as shown in Fig. 2 (a), and break the copper strips with the aid of a small drill or the special cutting tool that is available, as indicated in the diagram.

Now assemble TR_1 , TR_2 , R_1 , R_2 , C_1 , C_2 , R_3 , R_4 , RV_1 , and the battery supply leads, as shown in Fig. 2 (b), taking care to use heat shunts when soldering the transistors in place. All components are mounted vertically on the panel. The mounting legs of RV_1 should be carefully reduced in width with the aid of a small file before attempting to fit them in the small holes in the Veroboard. When this section of the circuit has been wired up, it can be given a functional check by connecting a suitable supply and connecting a high impedance earphone between TR_2 collector and the positive supply line via a 1μ F blocking capacitor. If the circuit is operating correctly a series of "clicks" should be heard at a repetition rate of the order of 2 to 3 c/s.

If satisfactory, disconnect the supply and wire up the rest of the circuit as indicated. The unit is now complete and ready for use. If required, the keying frequency of the generator can be increased by reducing the values of C_1 and C_2 , or it can be decreased by increasing the values of these two capacitors. The tone frequency can be altered by changing the value of C_4 .

In use, the output of the generator should be fed into an amplifier having an input impedance of greater than $1k\Omega$. If the output is to be fed into a lower impedance than this, wire a $1k\Omega$ resistor in series with the output lead.

The generator is turned on and off by a switch wired in series with the negative supply lead, this being shown as S_1 in the circuit diagram. If required, the unit can be switched from several alternative points by simply wiring suitable switches in parallel. These switches can also connect the supply to the power amplifier or alternative equipment with which the generator is to be used.



HIGH SENSITIVITY TRANSISTOR VHF PORTABLE

Part 2 of this article, a double superhet receiver for reception of the B.B.C. f.m. broadcasts, will be published next month and this concluding instalment will be devoted to constructional points together with details of optional a.f.c. and tuning indicator circuits.

AVAILABILITY . . .

When items are specified in our various Components Lists, we take the utmost care to ensure that such components are readily available to readers. This entails not only the careful editing of original copy but also, in many cases, correspondence with the author concerned ("where did you obtain this component—junk box or the Ark?") and telephone calls to both manufacturers and retailers to ascertain current availability and/or the supply of suitable alternatives.

IN LAST MONTH'S ARTICLE IN THIS SERIES WE GAVE some final notes on the three-stage long, medium and short wave receiver which forms the second constructional project in the "Understanding Radio" series, after which we discussed r.f. amplification and showed how the added screen-grid in a tetrode valve enabled r.f. amplification to be obtained without instability. The tetrode valve is the fore-runner of the pentode valve, which will be described in this month's contribution.

In the next issue we return to constructional matters, and details will be given for the addition of an r.f. amplifier to the three-stage long, medium and short wave receiver.


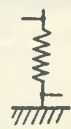

Negative Resistance

When we discussed the tetrode valve last month we saw that, when electrons from the cathode strike the anode, further electrons (known as secondary electrons) are emitted from that anode. These secondary electrons are attracted to the screen-grid if the anode voltage is more negative than the screen-grid voltage. This secondary emission effect results in the tetrode having $I_a V_a$ curves of the

1. The grid is given a fixed bias, which may, conveniently, be 1 or 2 volts negative of the cathode. The screen-grid is bypassed to the cathode by a capacitor having a low reactance at the frequencies it is intended to deal with. For simplicity, batteries are shown in Fig. 2, but any other means of supplying h.t. and grid bias voltages can, of course, be employed.

Oscillation is obtained in Fig. 2 because the negative resistance offered by the tetrode anode cancels out the positive resistance given by the losses in the tuned circuit. To maintain oscillation the negative resistance must be greater than the positive resistance due to tuned circuit losses, and the amplitude of the oscillation is limited between the points where the anode voltage changes from the negative to the positive resistance condition. In Fig. 1, this corresponds to the points where the downward-sloping section of the curve meets the upward-sloping sections.

The circuit of Fig. 2 is known as a *Dynatron oscillator*. It is not normally encountered in receiver circuits, but it has applications in experimental or test equipment work. The resonant frequency of an unknown tuned circuit can, for instance, be

<h1>UNDERSTANDING RADIO</h1>	<h2>The R. F. Pentode</h2>	$f = \frac{1}{2\pi\sqrt{LC}}$	
			

by W. G. Morley

type shown in Fig. 1. (Fig. 1 is Fig. 8 (b) from last month, reproduced again for reference.)

We also saw, in the last article, that the curves between points A and B in Fig. 1 exhibit *negative resistance* because, when anode voltage increases, anode current decreases. Such an effect is, of course, completely opposite to the positive resistance offered by a normal resistor with which, when voltage increases, so also does current.

It is possible to make use of the negative resistance exhibited at the anode of a tetrode by assembling the oscillator circuit shown in Fig. 2. This oscillator has the interesting feature that only two connections are made to the parallel tuned circuit, there being no coupling windings or other means of obtaining positive feedback as has occurred in the single-valve oscillators we have previously examined. The supply voltages applied to the anode and screen-grid are such that the anode is more negative than the screen-grid. Best results are given when the applied anode supply voltage is in the centre of the negative resistance section between points A and B in Fig.

found by simply connecting it into a Dynatron oscillator circuit. Provided its losses are not too high the tuned circuit will then oscillate at its resonant frequency, which can be measured on other equipment. The measuring equipment could be a receiver with a calibrated tuning scale which is adjusted to receive the oscillatory frequency.

RF Amplification

In receiver work we are more concerned with using a tetrode valve as an r.f. amplifier than with its performance in the Dynatron oscillator circuit. When it is used as an r.f. amplifier, the tetrode valve will appear in a circuit of the type shown in Fig. 3, in which tuned circuits appear in both grid and anode circuits. In a circuit of this nature we set up the anode and screen-grid supply voltages so that the anode voltage is always on the more linear section (i.e. the section more closely resembling a straight line) of the $I_a V_a$ characteristic, as shown by the curves to the right of point C in Fig. 1. This requirement may be assured in

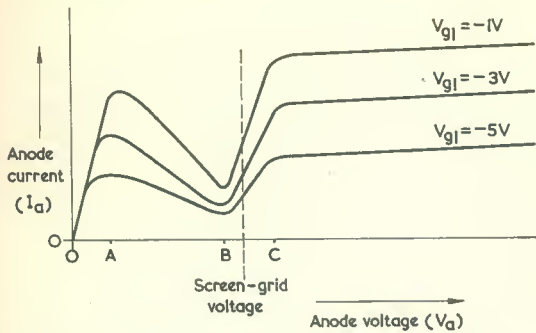


Fig. 1. Representative I_aV_a curves for a tetrode valve

Fig. 3 by giving the screen-grid dropping resistor a value which allows the screen-grid to be well negative of the anode. We obviously do not want the anode voltage to approach the negative resistance section between points A and B of Fig. 1, because the anode tuned circuit may then tend towards oscillation due to Dynatron effect. Even with an anode supply voltage considerably higher than the screen-grid voltage, we may still encounter difficulties if the amplified r.f. signal at the anode has a very high amplitude, since negative half-cycles may cause the anode voltage to swing into the negative resistance section.

Anode A.C. Resistance

Let us now turn to another aspect of the curves shown in Fig. 1. In this instance we shall assume that the tetrode has supply voltages, and handles signal amplitudes, which are such that the anode is always on the more linear section of the character-

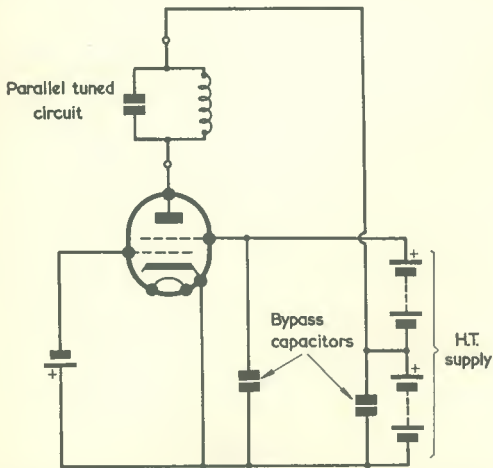


Fig. 2. The Dynatron oscillator. Anode and screen-grid bypass capacitors are shown to emphasise that a low impedance at these points in the circuit is necessary. The capacitors should have values which give a low reactance at the frequencies to be handled

istic to the right of point C in Fig. 1. Let us next examine what occurs under these conditions.

The most evident feature of the more linear section is that the I_aV_a curves are very nearly horizontal. We have already seen¹ that the anode a.c. resistance (r_a) of a valve may be represented by the slope of its I_aV_a curve. Fig. 4 shows two imaginary I_aV_a curves, the curve at Fig. 4 (a) having a greater slope than that at Fig. 4 (b). The lower slope at Fig. 4 (b) infers that unit change in anode voltage results in a smaller change in anode current than in Fig. 4 (a), and that the curve of Fig. 4 (b) represents a higher anode a.c. resistance in consequence. Inspection of the tetrode I_aV_a curves

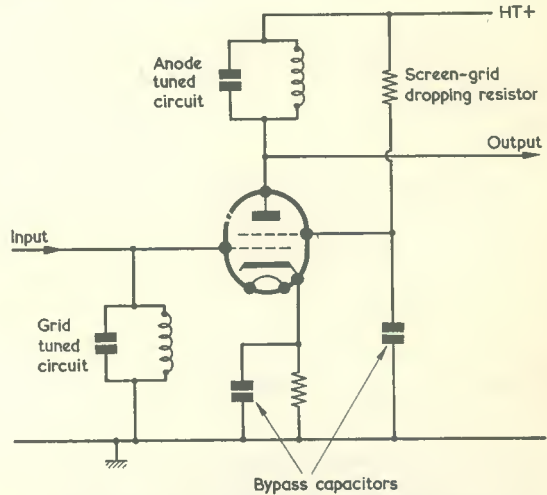


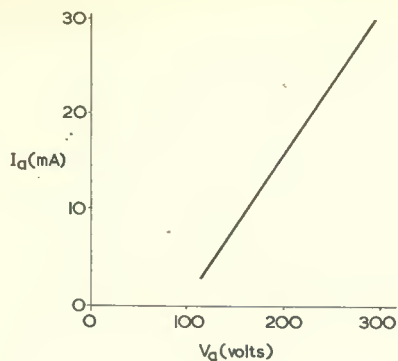
Fig. 3. As an r.f. amplifier, the tetrode will appear in the basic circuit shown here (in which cathode bias is assumed). If the anode goes negative of the screen-grid during part of the r.f. cycle, the I_aV_a characteristic may enter the negative resistance section of Fig. 1.

to the right of point C in Fig. 1 shows that these have very little slope and are, indeed, nearly horizontal. In consequence, a change in anode voltage results in an extremely small change in anode current, and it follows from this that the anode a.c. resistance must be very high.² This is in fact true and, in practice, tetrodes can exhibit anode a.c. resistances of $1M\Omega$ or even more.

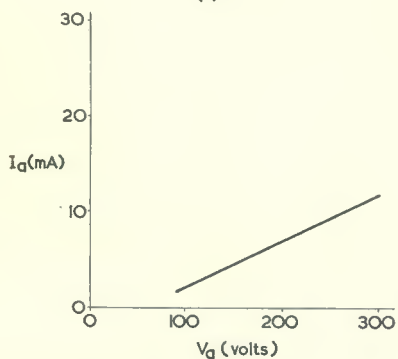
Another valve characteristic previously dealt with is mutual conductance (g_m), which defines the change in anode current for a unit change in

¹ The valve characteristics which are next referred to were described in "Understanding Radio" in the issues for July, November and December, 1965.

² Readers who are unfamiliar with the anode a.c. resistance concept and who do not have the appropriate earlier issues available, may remember that we use the equation $R = \frac{E}{I}$ with reference to normal resistance. We find the anode a.c. resistance by measuring the alternating current which flows for a small alternating voltage applied to the anode, and we similarly employ $R = \frac{E}{I}$ for these two alternating quantities. If an alternating voltage is applied to the nearly horizontal curves of Fig. 1 the resulting alternating current will be very small, hence the effective anode a.c. resistance will be very high.



(a)



(b)

Fig. 4. Two imaginary $I_a V_a$ curves which illustrate the fact that anode a.c. resistance depends on the slope of the $I_a V_a$ characteristic. Since, in (b), unit change in anode voltage corresponds to a lower change in anode current, this curve depicts a higher anode a.c. resistance

grid voltage, and which is expressed in terms of "mA per volt". Examination of the curves to the right of point C in Fig. 1 shows that we get a very noticeable change in anode current when we change the grid voltage from $-1V$ to $-3V$ and, again, when we change it from $-3V$ to $-5V$. These changes in anode current for change in grid voltage are of the same order as are given by the voltage amplifier triodes we have examined previously. In consequence, we can look upon a tetrode as having about the same value of mutual conductance as we find in a voltage amplifier triode.

A final point arising from the earlier articles is that the amplification factor (μ) of a valve is equal to the product of the anode a.c. resistance and the mutual conductance. That is:

$$\mu = g_m \cdot r_a$$

Since we have a mutual conductance in the tetrode which is of the same order of magnitude as that in a voltage amplifier triode, and at the same time have an anode a.c. resistance which is considerably higher, this equation tells us that the

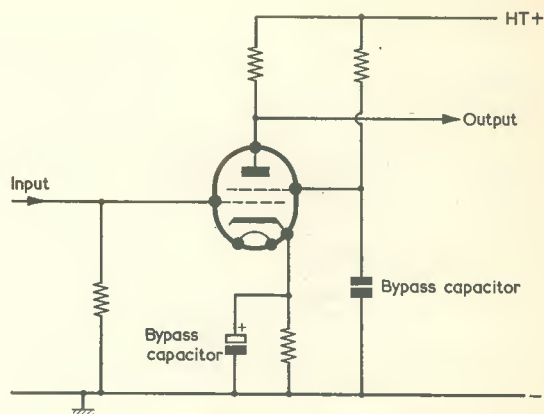


Fig. 5. Using a tetrode valve as an a.f. amplifier. Again, cathode bias is assumed and the cathode bypass capacitor would in practice be an electrolytic component having a value around $10\mu F$ or more. A limitation of the circuit is that anode voltage must remain on the more linear section of the $I_a V_a$ characteristic throughout the a.f. cycle

amplification factor must be similarly much higher. Such an assumption is correct, and the amplification factor of a tetrode can, in practice, be of the order of 1,000 or even more.

The fact that a tetrode offers a high amplification factor makes it possible to use the valve as a high-gain a.f. voltage amplifier in a circuit of the type shown in Fig. 5. With the requisite component values, the tetrode valve will offer a greater amplification than a triode. In Fig. 5 the screen-grid bypass capacitor should have a low reactance at audio frequencies and could, typically, have a value of $0.1\mu F$. The anode voltage must not be allowed to fall below the more linear section of the $I_a V_a$ characteristic. A signal at the anode with sufficiently large amplitude could cause the anode to enter the non-linear section of the $I_a V_a$ characteristic on the peaks of negative half-cycles, with the result that the anode signal would become seriously distorted. This factor limits the amplitude of the amplified signal which may appear at the anode of the amplifier in Fig. 5.

The Pentode

The original intention behind adding the screen-



Fig. 6. In the pentode valve an additional grid is interposed between screen-grid and anode. The additional grid is connected to cathode

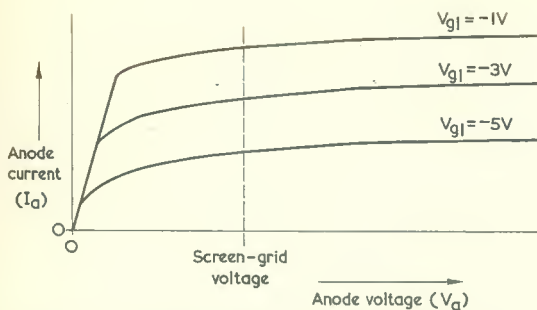


Fig. 7. A representative set of I_aV_a curves for a pentode. These are similar to those in Fig. 1 for the tetrode, with the exception that the non-linear section has now been eliminated

grid to the triode (and thereby forming the tetrode) was merely to reduce the internal capacitance between anode and grid to a negligible value, thereby enabling the valve to function as an r.f. amplifier without instability. However, the presence of the screen-grid causes considerable changes to the operation of the valve. First, if the anode voltage is below screen-grid voltage the I_aV_a characteristic can exhibit negative resistance. Second, if the anode voltage is higher than screen-grid voltage the valve exhibits a much higher amplification factor than does a triode. It would be desirable to eradicate the negative resistance effect in the tetrode whilst retaining the high amplification factor. This requirement is achieved with the pentode valve, which has superseded the tetrode valve in receiver applications. It should be added that, although the tetrode valve is now outmoded, it is necessary to understand how it functions if pentode operation is to be fully appreciated.

In the pentode a third grid is interposed between the screen-grid and the anode, and it is connected to the cathode as shown in Fig. 6. Since this third grid is negative of the anode it repels the secondary electrons which the anode emits, and these are forced to return to that electrode. The third grid consists of wires having the same basic construction as the control grid, but with wider spacing between wires. Because of the wide spacing, this additional grid does not prevent electrons from the cathode travelling to the anode in the same way as occurred in the tetrode valve.

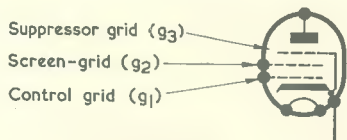


Fig. 8. The circuit symbol for a pentode, with the grids identified. In this instance, an internal connection inside the valve envelope connects the suppressor grid to the cathode. The suppressor grid may, alternatively be brought out to a separate pin, as in Fig. 6, whereupon the connection to cathode is made externally

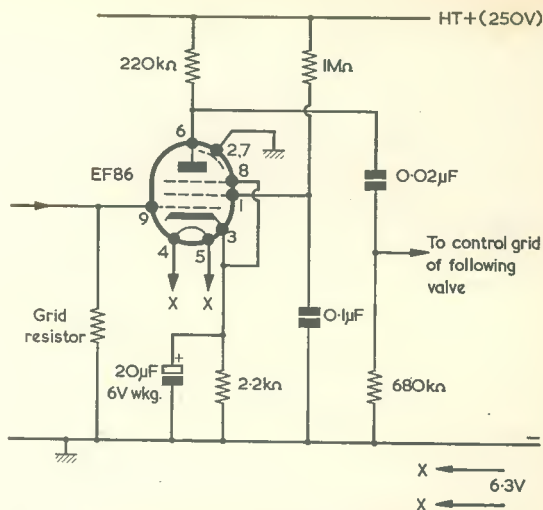
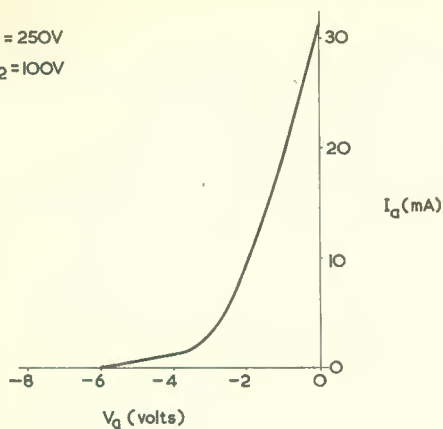


Fig. 9. A practical pentode a.f. voltage amplifier circuit. The grid resistor may have a value between $470k\Omega$ and $1M\Omega$, whilst the $680k\Omega$ resistor at the right is the grid resistor of the following valve. The EF86 has an internal screen around the electrode assembly; this is brought out to pins 2 and 7 and connects to chassis

The fact that the third grid prevents secondary electrons from the anode reaching the screen-grid causes the non-linearity and negative resistance exhibited by the tetrode to be completely eliminated. A set of typical I_aV_a curves for a pentode appears in Fig. 7, and it will be seen that these are similar to those for the tetrode in Fig. 1, with the exception that the non-linear section has now disappeared entirely. The pentode I_aV_a curves remain very nearly straight down to a value of anode voltage which is much lower than screen-grid voltage, and it follows from this that the signal at the anode can have a large amplitude without introducing serious distortion. In other respects, the pentode resembles the tetrode when the latter operates over the more linear section of its I_aV_a curve. The pentode can operate as a stable r.f. amplifier, and it has a high anode a.c. resistance and a high amplification factor.

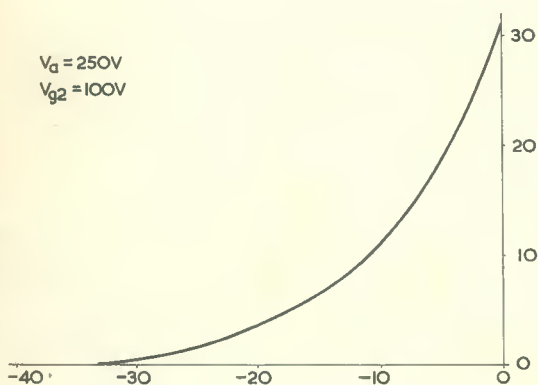
The additional grid introduced into the pentode is known as the *suppressor grid*, because it suppresses the effect of secondary emission from the anode. When the grids in the valve are referred to by the letter *g* and subscript numbers, the suppressor grid is designated as g_3 (the control grid being g_1 and the screen-grid g_2). In some valves the suppressor grid connection is brought out to a separate pin, whereas in others it is internally connected to the cathode. When the suppressor grid connection is brought out separately it may be connected externally to the cathode at the valveholder (or to chassis, which will usually be close to cathode potential). The circuit symbol for the pentode is shown in Fig. 8. The term "pentode" derives from the fact

$V_a = 250V$
 $V_{g2} = 100V$



(a)

$V_a = 250V$
 $V_{g2} = 100V$



(b)

that the valve has five electrodes (the cathode, the three grids and the anode).

The type numbers for pentodes follow the usual coding practices. For instance, the 6BW7, an r.f. pentode on the B9A base, follows the American coding in which the figure 6 signifies a 6.3 volt heater and the final 7 signifies the number of "useful elements", which in this case consist of the heater, the cathode, the three grids, the anode and an internal screen around the electrode assembly. The letters BW have no significance to the user apart from identification of the valve.

A typical pentode coded with the European system is the EF91. In this code the letter E signifies a 6.3 volt heater and the letter F a voltage amplifier pentode.³ The first figure, 9, indicates a B7G base, whilst the second figure, 1, is merely a number indicating the design or development.

Because of the high amplification offered by the pentode, it is often used as an a.f. amplifier. A representative circuit is shown in Fig. 9 in which an EF86 pentode is employed. The EF86 is specifically designed for this class of work and, with the component values shown in the diagram, the stage gain (anode signal voltage divided by grid signal voltage) is very nearly 200. The letters EF in the coding for the EF86 signify a 6.3 volt heater and a voltage amplifier pentode respectively, whilst the first figure, 8, indicates a B9A base. The second figure, 6, is the design or development number.

Variable-Mu Pentodes

For a number of r.f. amplifier functions in receivers it is desirable to be able to control the gain of a pentode stage by varying its grid voltage. The required effect is that the stage gain should reduce as the control grid goes negative.

Fig. 10 (a) shows an $I_a V_g$ curve for a pentode which is not specifically intended for this class of operation. As may be seen, the anode current falls to zero at a grid voltage of $-6V$, this being the *cut-off voltage* for the valve at the anode and screen-grid voltages indicated. By suitably spacing the wires in the control grid it is possible to obtain the $I_a V_g$ characteristic illustrated in Fig. 10 (b). This characteristic shows a continual curvature throughout its length and the cut-off voltage is reached at $-33V$. The curve of Fig. 10 (b) is representative of a *variable-mu* (or *variable- μ*) r.f. pentode, and it demonstrates the fact that the mutual conductance of the associated pentode may be adjusted "smoothly" by varying the control grid voltage. With the curve of Fig. 10 (a), changes in mutual conductance are liable to be more abrupt as cut-off point is approached.

The variable-mu effect is obtained by variations in the spacing between the control grid wires along the length of the control grid assembly. In a typical assembly the wires at each end may be closely spaced, the spacing increasing towards the centre so that the centre wires are proportionately widely spaced apart. When an alternating signal voltage is applied to the grid, the consequent control over electrons reaching the anode decreases as the grid wire spacing increases. Thus, with the grid assembly just mentioned, greatest control at low negative grid voltages is exerted by the closely spaced wires at the end. At high negative grid voltages the electrons from the cathode will not pass through the closely spaced wires at the end of the grid structure, and will only pass through the widely spaced wires at the centre, where less control is exerted.

Although the type of valve under discussion is known as a variable- μ valve, it is more helpful to think in terms of the changes in mutual conductance which occur when the control grid voltage is varied.

³ As we shall see later, some pentodes are designed to drive loud-speakers or carry out similar functions involving a relatively high power in the anode circuit. These are known as "output pentodes" and are designated by the letter L.

In Your Workshop



DO YOU REMEMBER," ASKED DICK, "what we were doing this time last year?"

Smithy took a prodigious draught from his battered tin mug, then placed it reflectively down on his bench alongside the debris remaining from his recently consumed lunch. A bleak February wind howled outside the window behind him, but inside the Workshop all was warm and snug.

"I can't," replied Smithy eventually, "say that I do remember."

"Cast your mind back."

Obligingly, Smithy allowed his thoughts to wander once more to the February of 1966.

"All that I can recall," he remarked eventually, "is that life twelve months ago followed exactly the same pattern as it's following today. In other words, I was repairing sets whilst you were wrecking them!"

"That February," said Dick impatiently, "we were talking about computers."

A sudden glimmer in Smithy's eye betokened recollection.

"Why, that's right," he exclaimed. "I remember now! I gave you some gen on simple AND and OR gates, together with the Boolean algebra that goes with them. If I'm not mistaken, I also mentioned NAND and NOR gates, too."

Binary Numbers

"You did," agreed Dick. "You explained what truth tables were, too. Dead fascinating it was!"

"As I remarked at the time," said Smithy, "you must bear in mind that I was only just brushing over the surface of the subject."

"Perhaps so," conceded Dick. "but you *did* give me an insight into principles which, until then, were completely unknown to me. Also, with computers being used as much as they are these days, a keen young working lad such as myself *has* to pick up at least a smattering of knowledge on the subject. As it happens, there's something else to do with computers I don't know anything about."

"What's that?"

"This binary number business. Computers work with binary numbers, don't they?"

"Digital computers do," said Smithy. "But don't forget that there are analogue computers as well. Analogue computers are used to solve problems by replacing each item in an equation representing the problem with an analogous electrical quantity such as capacitance, inductance, voltage, current or amplification. A circuit is set up in which

This month Smithy the Serviceman, at the instigation of his able assistant, Dick, once more enters the fascinating world of the digital computer. On this occasion he embarks on an explanation of the simpler facets of binary notation.

the terms of the equation are replaced by electrical quantities, with the result that the overall effect on a system of varying one or more quantities can be found very quickly. Analogue computers are very useful for working out complicated problems concerning aircraft performance and things like that. However, the answer they give to any problem is an approximate one, because it depends on the tolerances in the physical components which represent the terms in the problem. Digital computers, on the other hand, deal directly with figures and can give answers which are completely accurate. Most of the large computers you read about these days are digital computers, and these are the ones which work with binary numbers."

"Ah," said Dick, gratified. "I knew I was on the right track. How does this binary number business work, then?"

"It's dead simple," replied Smithy carelessly. "You just have two numbers. One of these is zero and the other is one."

With the manner of one who has answered a complicated question to the complete satisfaction of the enquirer, Smithy sipped at his tea and lapsed into silence. Dick glared at him irritably.

"Thank you very much," he remarked sarcastically. "After having picked up that priceless piece of information, I am now fully qualified to go out and service my first computer!"

Resignedly, Smithy drained his mug. It was obvious that yet another lunch-break was going to be sacrificed to Dick's relentless pursuit of knowledge.

"Fill that up," he ordered, holding out the mug, "then come back over here, and I'll see if I can introduce you to the subject of binary numbers."

DECIMAL	ROMAN	BINARY
9	IX	1001

Fig. 1. A quantity may be expressed accurately in any numerical system. Here we see the figure 9 in its decimal form, with its equivalent in Roman presentation and in binary notation

Delighted, Dick rushed to the Workshop teapot with Smithy's mug, returning almost immediately with this disgraceful utensil fully charged with the life-saving fluid.

"Right," said Smithy, briskly. "Let's get down to business! Now the first thing you have to bear in mind is that the calculations in a digital computer are carried out with the aid of circuits which can exist in one of two conditions. Transistors represent a typical example of the components employed, and they appear in circuits where their state represents one of the numbers handled by the computer in its calculations. Now, the only reliable way of using a transistor in this manner is to allow it to have two states only, these being either fully conductive or fully cut-off. Other circuits in the computer, including those used for storing numbers for later use in a calculation, are similarly restricted to two-state working if they are to operate reliably. Since the individual computer circuits can only work in one of two states, this fact has to be catered for in the arithmetic system that they use. That's where the binary notation comes in, because in the binary system there are only two digits, those being zero and one.

A transistor can then represent zero by being in the fully conductive state, and one by being in the fully cut-off state. Or vice-versa, according to the requirements of the particular circuit into which the transistor is connected."

"In other words," said Dick, "the whole reason for using binary numbers with computers is that these are the only numbers which the computer circuits can reliably handle."

"Exactly," confirmed Smithy. "The problem can be presented to the computer in terms of ordinary decimal numbers, and the answer can similarly be printed out by the computer in decimal numbers. But the actual calculations carried out in the works of the computer will be done in binary."

Number Systems

"I see what you mean," said Dick. "Is it possible, then, to represent a decimal figure in binary numbers?"

"Of course it is," replied Smithy. "All numbering systems correspond to accepted conventions. If, for instance, we count nine apples we can use the decimal digit, 9, to denote the quantity. The old Romans would have used the symbol IX.

In the binary system we can present the quantity of apples with the number one, zero, zero, one."

Smithy pulled his notepad towards him and scribbled the three numbers down. (Fig. 1).

"Now, it so happens," he continued, "that mankind has fallen, almost by accident, into the custom of using the decimal system of numbers for counting and calculation requirements. Both the figure 9, and the Roman IX appear in this system. It's not beyond the realms of possibility that mankind could, however, have adopted another system for his counting. This could, for example, have been an octal system in which there is no single digit higher than 7."

Smithy scribbled on his pad again. (Fig. 2).

"Now, in the decimal system," he went on, "we count in single digits up to 9, which represents the highest single digit in the system. After 9 we put a 0 and add a 1 to the left, giving us 10. The numbers then continue as 11, 12, 13 and so on. With an octal system we do the changeover to two digits after 7, because this is the highest single digit in the system, whereupon we get the sequence 5, 6, 7, 10, 11, 12 and so on. If we used the octal system we could say that there were 11 apples instead of, as with decimal, saying that there were 9 apples. The number of apples remains unaltered, and is expressed by 9 in a decimal system and by 11 in an octal system."

"I think I understand what you're getting at," said Dick thoughtfully. "You have a fixed quantity and it may be denoted with complete accuracy by digits in two quite different systems. The symbols appearing in the systems will, however, be different if the fixed quantity is higher than the highest single digit in either one or both of them."

"That's the idea," confirmed Smithy.

"We are all familiar with the decimal system because it's what we learnt at school and what we use for our everyday calculations. But it exists only by virtue of a single arbitrary rule which states that there shall be no single digit higher than 9. The octal system of numbering is unfamiliar to us but it could be just as useful for counting and calculations. The octal system has the arbitrary rule that there shall be no single digit higher than 7."

"What other numbering systems could you have?"

"Pretty near as many as you like," replied Smithy. "All you have to do

DECIMAL	OCTAL
1	1
2	2
3	3
4	4
5	5
6	6
7	7
8	10
9	11
10	12
11	13
12	14
13	15
14	16

Fig. 2. In the decimal system the highest single digit is 9, whilst in the octal system it is 7. This table shows decimal and corresponding octal figures alongside each other up to decimal 14

is to select the number which represents the highest single digit in the system and then work on from there. The highest single digit could be any number within reason, although some numbers would make arithmetical processes very difficult. Quite a convenient system, from the calculation point of view, would be a duodecimal system in which the changeover to two digits occurs at 12. For this you'd have to invent two symbols to act as single digits for 10 and 11, after which 12 would be represented as one, zero."

"I see," said Dick. "I hadn't realised that you could play around with numbers quite as easily as all that."

"You can do all sorts of things with them," replied Smithy cheerfully, "provided you use a bit of common-sense when you do calculations with them. However, the main point I wanted to make is that a quantity doesn't have to be expressed in terms of the decimal system. It can be expressed just as accurately in any other numerical system. I mentioned the octal system because you will encounter references to it in text-books or articles dealing with computers. The duodecimal system doesn't appear to have many applications in computer work, although it is sometimes referred to in general mathematics. With computers, you will also find references to the hexadecimal system, in which the changeover to two digits is made at decimal 16. Six new single-digit symbols have to be introduced to replace the decimal numbers 10 to 15 inclusive, and that's all I need to say about it at the moment."

"Good show," said Dick. "Let's get back to the binary system!"

"Fair enough," grinned Smithy, scribbling out a further series of figures on a new page of his pad (Fig. 3). "I wondered how long you'd be able to restrain your impatience! Now, with the binary system the arbitrary rule is that there shall be no single digit higher than 1. Whereupon the binary numbers increase in the manner shown in this table I'm making out. The binary equivalent of decimal 0 is zero, and the binary equivalent of decimal 1 is one. So there's no change there. When we reach decimal 2, however, we bump into the arbitrary rule in binary which says that there shall be no single digit higher than 1. So we put a zero and add a one to the left of it, giving us binary one, zero. Decimal 3 is easy, and the right hand digit goes up to one, giving us binary

DECIMAL	BINARY
0	0
1	1
2	10
3	11
4	100
5	101
6	110
7	111
8	1000
9	1001
10	1010
11	1011
12	1100
13	1101
14	1110
15	1111
16	10000
17	10001
18	10010
19	10011
20	10100

Fig. 3. The highest single digit in the binary system is 1. This table shows binary numbers corresponding to decimal numbers up to 20.

one, one. At decimal 4 the arbitrary rule comes into action again and so we put two zeros and add a one to the left, giving us one, zero, zero. And so the numbers continue. Decimal 5 is binary one, zero, one; decimal 6 is binary one, one, zero; decimal 7 is binary one, one, one; and decimal 8 is binary one, zero, zero, zero."

"The next one," remarked Dick, "is decimal 9. Which gives us the binary one, zero, zero, one number you put down initially when you were talking about those 9 apples. You know, Smithy, you must be pretty smart to remember binary equivalents just like that!"

"I don't bother to remember

them," remarked Smithy. "With small numbers it's quite easy to convert decimal to binary, or binary to decimal, by simple mental arithmetic."

Binary To Decimal

"I knew," complained Dick, "there'd be a snag somewhere!"

Surprised, the Serviceman looked up at his assistant.

"What's biting you?" he asked. "I thought you'd been following me without any trouble at all up to now."

"So I have been," said Dick unhappily. "But we've now come to the point where I'm stymied. It's all right for you to talk lightly about

10,000	10^4
1,000	10^3
100	10^2
10	10^1
1	10^0

Fig. 4. The decimal numbers corresponding to various powers of 10

7	5	6	0	2
(7×10^4)	$+ (5 \times 10^3)$	$+ (6 \times 10^2)$	$+ (0 \times 10^1)$	$+ (2 \times 10^0)$
$(70,000)$	$+ (5,000)$	$+ (600)$	$+ (0)$	$+ (2)$

Fig. 5. In a decimal number each digit, according to its position, represents a power of 10 multiplied by the digit

mental arithmetic, but I have got to rely on the old paper and pencil, mate!"

"If that's all you're worried about," commented Smithy, "you can forget it. There's no *necessity* to use mental arithmetic in converting from one system to another, and you can do it all on paper if you like."

"Well," that's a relief," said Dick thankfully. "My mind always goes blank when anyone mentions 'mental arithmetic'. Like the fly, I stick to paper!"

"As you like," commented Smithy. "You'll probably find, though, that you'll tend to convert the smaller numbers mentally after you've had a bit of experience with them. Incidentally, I'm glad to see you've picked up the 'one, zero, zero, one' patter when referring in speech to binary numbers. It makes things easier when you're tackling them for the first time."

Smithy returned to his pad.

"Let's press on to the next step," he went on. "Which consists of making a comparison between decimal and binary numbers. I'll start off by writing down a series of decimal figures to form the left hand column of a table, these being 10,000, 1,000, 100, 10 and 1. After that I'll put their equivalents in terms of powers of 10 in the right hand column."

Dick's face took on an expression of agonised concentration as Smithy wrote down the numbers on his pad. (Fig. 4).

"What," asked Smithy, "is 100?"

"10 squared?"

"10 squared," confirmed Smithy, writing the expression on the right of his pad. "And 1,000?"

"10 cubed," said Dick confidently.

"I see what you're driving at now, with this power of 10 business. This bit's a piece of cake."

"10,000?"

"10," replied Dick promptly, "to the power of 4."

"Fine," said Smithy, "now let's go below 100. What, in our table, is 10?"

"Well," replied Dick, "100, just above it, is 10 to the power of 2, so 10 will be 10 to the power of 1."

"And 1?"

"Would that be 10 to the power of nought?"

"It would indeed," said Smithy, entering the expression in his table. "And that's our table finished with for the time being. Next, I'm going to choose a decimal number at random. Let's say that it's 75,602. I'll write this down, then split it up into powers of ten starting from the right. (Fig. 5). Now, the 2 at the right is actually 2 times 10 to the power of nought. The zero next to it is zero times 10 to the power of 1, whilst the 6 is 6 times 10 squared. The 5 is 5 times 10 cubed and the 7 at the furthest left is 7 times 10 to the power of 4. If I put plus signs between each of these individual products, you can see that the figure 75,602 is equal to all of them added up together."

"Well, I'm dashed," exclaimed Dick. "So it is! Do you know, I'd never looked upon decimal numbers as being made up in that way before. Blimey, Smithy, you aren't half making me think of new ideas today!"

"We're getting to the really interesting part now," replied Smithy. "Let's next do the same thing with a binary number chosen at random. This could, for instance, be 10101. I'll write it down."

Smithy once more applied his pen to the pad. (Fig. 6).

"With the previous decimal number," said Smithy, "the base, or radix, of the system was the number 10, and we broke the number down to powers of 10 according to the position of each digit along the number. With the binary system, the base is 2, so we can repeat

the operation using 2 instead of 10. Starting from the right, that gives us 1 times 2 to the power of nought, zero times 2 to the power of one, 1 times 2 squared, zero times 2 cubed and, at the furthest left hand digit, 1 times 2 to the power of 4. Perhaps, Dick, you could work out these individual products for me."

"O.K.," said Dick, pulling Smithy's pad towards him and taking out his pen to enter the numbers. "Let's start at an easy one, which is 1 times 2 squared in the middle. That's obviously 1 times 4. The next one is zero times 2 cubed or zero times 8, so the answer here is zero. The end one at the left is one times 2 to the power of 4, so that will be one times 16. Now I'm getting to what, with me, is the dicey bit. If 10 to the power of one is 10, is 2 to the power of one 2?"

"It is."

"Not that it matters," said Dick, "because the zero makes the product nothing anyway! Also, would 2 to the power of nought be 1?"

Smithy nodded gravely.

"Then that's it," announced Dick triumphantly. "Binary 10101 is equal to 1 plus 4 plus 16. And that's 21."

"Which is," confirmed Smithy, "exactly correct. You've now carried out your very first conversion of a binary number to the decimal system. You use the same process for converting any other binary number to decimal, and all you have to remember is that the series of 2 with increasing powers is the same as the series 1, 2, 4, 8, 16, 32, 64, and so on. You then use this progression to find the decimal equivalent."

"This is a lot simpler than I thought it would be," said Dick excitedly. "Try me out with another one!"

"O.K.," said Smithy obligingly. "Let's think up one that's a bit longer. Write down the decimal equivalent of the binary 1100111." (Fig. 7).

1	0	1	0	1
(1×2^4)	$+ (0 \times 2^3)$	$+ (1 \times 2^2)$	$+ (0 \times 2^1)$	$+ (1 \times 2^0)$
(1×16)	$+ (0 \times 8)$	$+ (1 \times 4)$	$+ (0 \times 2)$	$+ (1 \times 1)$
(16)	$+ (0)$	$+ (4)$	$+ (0)$	$+ (1)$
$= 21$				

Fig. 6. The digits of a binary number qualify 2 to a power which depends upon the position of the digit, in a similar manner to Fig. 5. In this diagram we see that binary 10101 is equal to decimal 21

1	1	0	0	1	1	1
(1x64)+	(1x32)+	(0x16)+	(0x8)+	(1x4)+	(1x2)+	(1x1)
(64) +	(32) +	(0) +	(0) +	(4) +	(2) +	(1)
						=103

Fig. 7. Another exercise in converting a binary number to its decimal equivalent

"Right," said Dick confidently. "Well, working from the right, we get 1 times 1, which is 1. We then have 1 times 2, which is 2; 1 times 4, which is 4; zero times 8, which is zero; zero times 16, which is zero; 1 times 32, which is 32; and 1 times 64, which is 64. So the whole number is equal to 1 plus 2 plus 4 plus 32 plus 64. Half a mo!"

There were several moments of anguished calculation.

"That's it!" exulted Dick. "The answer's 103."

"Excellent," commended Smithy warmly. "You've got the whole binary to decimal business completely buttoned up, Dick. We're now going on to a further step which involves the demonstration of a very intriguing little effect. I'm going to add a zero to the right hand end of that binary number you've just tackled and then see what the decimal result is."

Smithy retrieved his pad and wrote down the new number. (Fig. 8).

"I'll start from the right," he remarked. "Whereupon I get no 1, one 2, one 4, one 8, no 16, no 32, one 64 and one 128. And that little lot adds up to 206."

"Blimey," said Dick, impressed. "It's exactly twice the previous number."

"I know," replied Smithy, "and this little exercise confirms the fact that, if you want to multiply a binary number by 2, you add a zero at the right hand end. Similarly, if you want to divide a binary number by 2 you just chop off the digit at the right hand end. If this is a zero then you've achieved an exact division by 2. If the right hand digit is a one, you've achieved a division by 2 with a remainder of one. Which also goes to show that all even numbers in binary end in zero and all odd numbers end in one."

Decimal To Binary

"Blow me," ejaculated Dick, "you don't half bump into some crafty effects with this binary bit."

"It's funny you should say that."

"Why?"

"Because the word 'bit' is, actually,

short for 'binary digit,'" replied Smithy, draining his mug for the second time that afternoon. "A 'bit', or 'binary digit', may have a value of one or zero only. You'll also encounter the term 'bit' in information theory, in which it defines the smallest possible piece of information which can be transmitted or stored. Now, how about filling up my mug for me?"

But there was no reply from Smithy's assistant and, glancing round, the Serviceman saw that Dick's face wore the withdrawn expression which indicated one of his periodic visitations from the Muse. Dick had the disconcerting habit of being able to compose instant doggerel to suit almost any occasion. Smithy sighed, and hoped it wouldn't be too bad this time.

Suddenly, Dick's face cleared.

"Here we are, Smithy," he said proudly. "Just listen to this little effort I've made up."

"Said the One to the Nought, Please admit it. I'm your better in learning—I'm with it."

"Said the Nought, What bombast! My exams are all passed."

"I'm an O-level binary digit!"

"I suppose," commented Smithy, "that I've heard worse."

"I'm glad you liked it," said Dick, pleased, as he picked up Smithy's proffered mug and took it over to the Workshop sink for replenishment. "Dead swinging I am on the ode scene, these days!"

"I wouldn't argue," replied Smithy carefully. "Anyway, I haven't yet told you how to convert decimal into binary."

"Ah, that sounds interesting," said Dick, returning with Smithy's mug. "Is it as easy as converting the other way?"

"Pretty well," replied Smithy

1	1	0	0	1	1	1	0
(1x128)+	(1x64)+	(0x32)+	(0x16)+	(1x8)+	(1x4)+	(1x2)+	(0x1)
(128) +	(64) +	(0) +	(0) +	(8) +	(4) +	(2) +	(0)
							=206

Fig. 8. Demonstrating that the addition of a zero at the right hand end of the binary number of Fig. 7 causes its value to be doubled. (Inspection shows that each of the useful products in the second line is moved up by a factor of 2)

laconically. "I'll demonstrate the idea by choosing a decimal number at random. I'll take a fairly high number, say 181. What I next do is to write out the series 1, 2, 4, 8, and so on, until I get to the highest number which is below the decimal number I want to convert. I write the series down from right to left." (Fig. 9).

Once more Smithy's pen clattered over his pad.

"In this case," he continued, "I'll stop the progression at 128 because the next number in the series is 256 and this is higher than 181. I put a one under 128 and subtract it from 181 whereupon I've got 53 left. 64 is higher than 53 and I put a zero under 64. At the same time, 32 is lower than 53 and so I put a one under it and subtract, giving me 21. 16 is lower than 21 and I put a one under this, too, and subtract again, whereupon I have 5 left. 8 is higher than 5 and a nought goes under it. 4 is lower than 5, so a one goes under it and, after subtraction, I've got 1 left. So a nought goes under the 2 and a one goes under the 1, and there's no further remainder. And there you are! The binary number is read from the bottom line and, in this case, it's 10110101."

"Stap me," breathed Dick. "Now that is something. Can you do this with any decimal number?"

"Any number you like," replied Smithy. "If you look at the numbers in that table I made up, you'll see that all the binary equivalents can be produced by following the procedure I've just described. You just keep working along the progression until there's no remainder left. If the remainder becomes zero before the end you just continue with noughts along the bottom line as appropriate. For instance, with decimal 8 the remainder goes to zero after 8 in the progression, whereupon the binary equivalent is 1000."

Addition And Subtraction

"What," asked Dick, "about addition of binary numbers?"

"That's pretty simple too," said

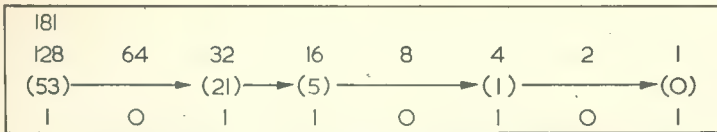


Fig. 9. Showing the steps involved in converting a decimal number to the binary system

Smithy. "All you have to remember is that there's no 2 in binary, with the result that 1 plus 1 comes to 10. After that, it's just the same as a decimal addition. Let's work out a little addition problem using binary numbers chosen at random."

Smithy jotted down the figures on his pad. (Fig. 10).

"Here we are," he said briskly. "Let's work this one out, starting at the right-hand column of digits. Here we have 1 plus 1, giving us 0 for the bottom line with 1 to carry to the left. The second column from the right is 0 plus 0 so, with the 1 we've carried over, we put down 1 in the bottom line. The third column from the right is 0 plus 0. There are no numbers carried from the previous column so we just put 0 here. The fourth column from the right is 0 plus 1 with no numbers carried, so we put down 1. The fifth column is 1 plus 1, giving us 0 to put down and to carry. The sixth column is 1 plus 1 plus the 1 carried over, and the answer here is 11. We put down 1, and carry 1 to the seventh column from the right where the result comes to 10."

Smithy scribbled on his pad for a few moments.

"I'm just working out the decimal equivalents of those numbers," he explained. "Here we are! The top line comes to 113 in decimal, the middle line to 57 in decimal and the bottom line to 170 in decimal. So the answer we got in binary is proved correct in decimal."

"That seems easy enough," commented Dick. "What about subtraction?"

"That's fairly easy in principle," replied Smithy, "but it tends to get a little involved in practice, as I'll show you in a moment. You first of all have to remember three obvious facts. These are that 1 minus 0 is 1, 1 minus 1 is 0, and

that 10 minus 1 is 1. You then carry out the subtraction using the same basic principles as in decimal. Let's try out an example." (Fig. 11).

"Off you go, then," said Dick. "I'm right behind you!"

"Good," said Smithy. "Well, we'll start at the right hand column again. This is 1 minus 1, giving us 0. No difficulty there! The next column is 0 minus 1, so we say we're subtracting 1 from 10, whereupon we put the answer 1 in the bottom line and we have to borrow 1 from the next left-hand column. We carry out the same borrowing process, of course, when we subtract in decimal. The third column from the left is 1 minus 0, but we've

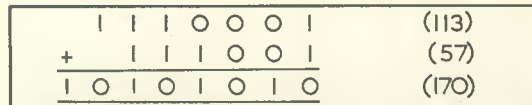


Fig. 10. An addition with binary numbers. The numbers in brackets at the right are the decimal equivalents

already borrowed a 1, so it becomes 0 minus 0, which equals 0. The fourth column from the right is 0 minus 1, so we once again borrow a 1 from the next left top digit and put down 1 on the bottom line. But the next left digit in the top line is also a 0, and so the borrowing action causes this to change to 1 and a 1 to be borrowed from the next digit along. The fifth column from the right is equivalent therefore to 1 minus 1, giving us zero. When we look at the sixth column, from the top digit of which we've already borrowed a 1, we're faced with yet another 0! So this reverts to 1 also and we borrow again from the seventh left column which, at long last, is a 1. The sixth column from the right is therefore 1 minus nothing, which gives 1. The 1 in the seventh column has already been borrowed by the sixth column

so this is now 0, and there is no figure to put down in the bottom line. Well, that's the subtraction done. Hang on a minute while I work out the decimal equivalents. Here they are! The top line is equal to 69, the middle line to 27 and the bottom line to 42. Which proves that we have worked the problem out correctly."

Not So Easy

"That borrowing business," said Dick critically, "is a little complicated, isn't it?"

"It is rather," agreed Smithy. "And it's easy to make mistakes when you're doing it, particularly if the borrowing carries on for several columns. The computer boys have a very crafty alternative method of subtracting in binary which consists of adding the complement of the number to be subtracted and which gets over all these borrowing problems. But the principles by which it operates are slightly more advanced than the very simple ones we've been looking at today so, for the time being at any rate, I won't

do more than just mention the fact that this alternative approach exists."

Smithy glanced at the Workshop clock and started.

"Come on Dick," he said, getting up hurriedly. "We should have got back to work all of twenty minutes ago! You'll get me the sack one of these days, keeping me nattering on like this after lunch-break."

"Sorry, Smithy," said Dick contritely. "I just hadn't noticed the way the time was going."

A thought occurred to him and he dug into his trouser pocket and handed a coin to Smithy.

"Whilst I remember it," he remarked. "Here's that 10 shillings I borrowed from you last week."

"This isn't 10 bob," replied Smithy, suspiciously. "It's a florin."

"I know," said Dick. "But I've suddenly discovered an easy way of clearing up all my outstanding debts. From now on I'm going to assume that they're all in binary notation!"

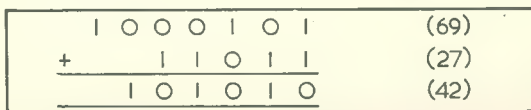


Fig. 11. Subtracting one binary number from another. Again, decimal equivalents are given, in brackets, at the right



MOST READERS WILL BE FAMILIAR WITH COLD cathode gas filled stabiliser tubes; they enable a stabilised potential in the range of about 75 to 150 volts to be obtained at currents of the order of 10mA. Various types of electronic equipment such as Geiger tubes, photomultiplier tubes and cathode ray tubes require a supply potential in the range 400 to 2,000 volts or more, but the current taken is usually less than 1mA. Although the normal type of gas filled glow discharge stabiliser tube is unsuitable for this application (unless used as a part of a complex stabiliser), a type of tube known as a corona voltage stabiliser has been developed during the last decade specifically for this type of application.

The Corona Discharge

A corona discharge can take place from an electrode which is shaped so that a high electric field strength is present at some parts of its surface. All television servicemen are familiar with the corona discharge which sometimes occurs from any point in the e.h.t. circuit of a television receiver and which is sometimes suppressed by means of a small piece of wax. Such corona discharges do not normally occur on smooth surfaces, but readily occur from a pointed part of the surface. The end of a stray wire is a common source of such an unwanted corona discharge.

Any stray electrons (formed by cosmic rays, stray radioactive atoms, etc.) will be accelerated by the electric field until they have enough energy to form ions in the air. The electrons thus formed are in turn accelerated until they have sufficient energy to form more ions. Thus enough ions are formed to carry a small current. Some of the air molecules gain energy from the electrons in the discharge and give out light as they return to the normal state. In a darkened room one can therefore see the well-known weak glow in the region of a corona discharge; it surrounds the area of high electric field strength.

Structure

A corona stabiliser tube consists of a central wire surrounded by a cylindrical cathode, the central wire being made positive in order to render the discharge as stable as possible. Some types of

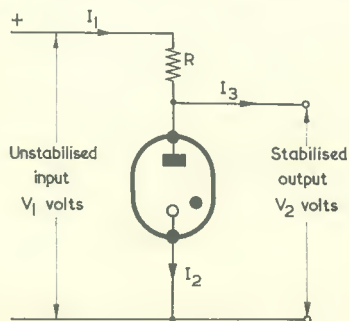


Fig. 1. Basic circuit for a corona stabiliser

CORONA VOLTAGE STABILISER TUBES

J. B. DANCE M.Sc.

corona tube have a B7G or B9A base with a top cap anode, whilst other types have a cap at each end or flying leads. Owing to the high potentials used few (if any) single-ended corona tubes have been manufactured.

Almost the whole of the potential applied to the tube appears in the region of the central anode wire where the ionisation occurs. The fact that this potential is virtually independent of the current passing through the discharge enables the corona tube to be employed for stabilisation purposes. A relatively small voltage is present between the outside of the ionisation region and the cathode. This voltage is, however, approximately proportional to the current flowing through the tube and therefore limits the voltage stability of the tube.¹

Circuit

The basic circuit in which a corona stabiliser circuit is used is shown in Fig. 1. This type of circuit is the same as that for the common glow discharge stabiliser tube. In the case of corona stabiliser tubes, however, the applied potential, V_1 , is much greater than in the case of glow discharge tubes, and the currents I_1 and I_2 are much smaller than with the glow discharge tube circuits.

Operating Conditions

As in the case of the glow discharge stabiliser tubes, there is a minimum current which must be passed through the tube to ensure satisfactory operation. This minimum current is usually in the range of 1 to 20 μ A. At smaller currents the discharge may be intermittent. The maximum current which should be passed through a corona stabiliser is usually in the range of 300 μ A to 1mA. Higher currents may cause the voltage across the tube to

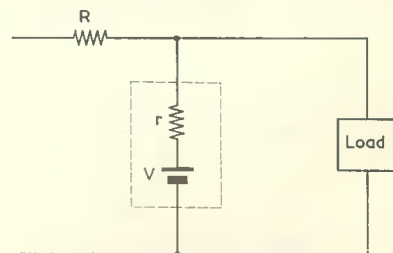


Fig. 2. Equivalent circuit of Fig. 1

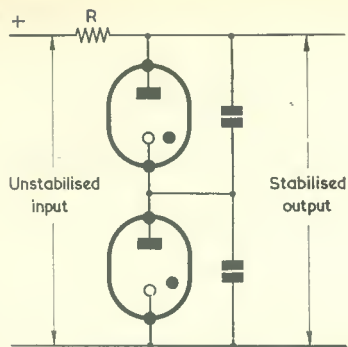


Fig. 3. Two corona stabilisers in series

fall as the discharge takes the form of a "streamer".

The operating voltage of a corona tube is usually in the range of 350 to 7,000 volts, although tubes with higher operating voltages can be manufactured. Two or more tubes can be connected in series if a very high voltage is to be stabilised. This will also enable stabilised voltages of less than the full voltage present across the series connected tubes to be obtained.

Corona stabiliser tubes are normally filled with pure hydrogen, since the highly mobile positive ions formed in this gas enable the rate of change of operating voltage with change of current to be kept to a minimum. In addition the use of hydrogen results in a higher maximum operating current than would be possible with other gas fillings. The operating voltage of a tube is approximately proportional to the gas pressure²; it may also be raised by increasing the diameter of the electrodes, but this will result in an increased change of operating voltage for a given change of the tube anode current.

A small amount of thorium oxide is included in some of the lower voltage corona tubes. The thorium is very weakly radioactive and provides ions which initiate the discharge. This ensures that the tubes strike promptly when a suitable potential is applied. The higher voltage tubes will strike promptly without this material.

Circuit Design

In the circuit of Fig. 1 the resistor R is usually more than 1MΩ. Its value can be found by applying Ohm's Law to the circuit. The current I₂ is usually fixed at a value about half-way between the maximum and minimum tube operating currents. The

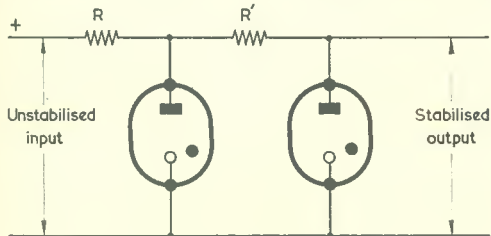


Fig. 4. A cascade circuit

current flowing through R is equal to the sum of I₂ and the load current I₃. The unbalanced input voltage, V₁, is generally about 1.2 to 1.5 times the operating potential of the tube, V₂. The potential across the resistor R is equal to (V₁-V₂). The value of R is calculated from the voltage across it and the current passing through it.

When the circuit has been designed using these principles, it should be checked to ensure that any variations in the input voltage and in the load can be accommodated without the operating conditions of the tube passing outside the recommended limits. For example, if the input voltage is liable to increase by up to 10%, a check should be made to ensure that this increase will not result in the current through the tube exceeding the maximum permissible value. Similarly if V₁ is liable to fall by 10%, a check should be made to ensure that a current at least equal to the minimum recommended tube current will continue to pass through the stabiliser tube. In addition, the unbalanced input voltage should always be kept at least 1.1 times the operating voltage of the tube or there may be a delay in striking.

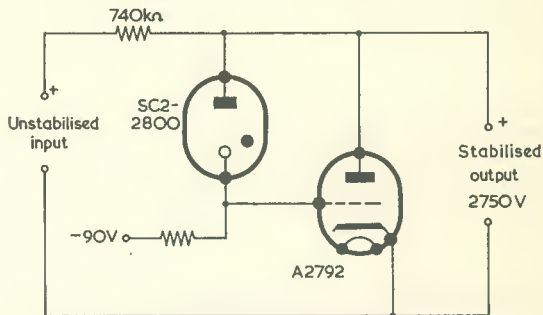


Fig. 5. Combining a stabiliser with a triode valve provides a high voltage supply having a low output impedance. The output is approximately 2,750 volts

Equivalent Circuit

The equivalent circuit of Fig. 1 is shown in Fig. 2. The components inside the dotted lines represent the tube itself. It can be seen that the tube behaves as a steady source of potential, v, in series with the resistor r. This resistor r represents the impedance of the tube. It is typically of the order of 100kΩ; the lower the value of this imaginary resistance, the better the stabilisation. The value of r varies somewhat with the current passing through the tube owing to gas heating effects.

Corona stabiliser tubes (like glow discharge tubes) cannot be connected in parallel for higher current operation. This is because one of the tubes will have a slightly higher operating potential than the other and will probably pass a negligible current.

Further Circuits

When corona stabilisers are used in series, capacitors should be connected in parallel with each tube as shown in Fig. 3. The values of these capaci-

tors may be a few thousand picofarads. If they are omitted, relaxation oscillations may occur.

In the circuit of Fig. 4 two corona stabilisers are employed in cascade. The variation of the stabilised output will be much less than in the simple circuit of Fig. 1 for changes in the supply voltage. However, the output impedance of the circuit is similar to that of the circuit of Fig. 1 and therefore the change in the output voltage with change in the load will also be similar. The operating potential of the first tube should be at least about 15% above that of the second tube.

Occasionally a corona stabiliser may be used to control the grid potential of a valve. In the circuit of Fig. 5³ V₁ is a corona tube. As the input voltage varies between 3,720 and 4,280, the anode current of the triode varies from 0.29 to 1.05mA at full load and 1.09 to 1.85mA at no load, but the variation

in the output voltage is too small to measure.

The use of the corona stabiliser in Fig. 5 enables almost 100% of negative feedback to be available (even at zero frequency) so that the output impedance is about $1/g_m$ (where g_m is the mutual conductance of the triode). The output impedance is typically about 500Ω—which is extremely low for an e.h.t. supply unit providing some 2,750 volts. The current passing through the stabiliser tube is about 90μA.

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1. R. O. Jenkins. "Corona Discharge finds a useful application in Voltage Stabilisers", *Electrical Review*, 14th June 1963
2. E. Cohen and R. O. Jenkins. "The Characteristics and Applications of Corona Stabiliser Tubes", *Electronic Engineering*, Vol. 32, No. 383, Jan. 1960
3. J. P. Holland. "Stabilised High-Voltage Supplies", *Industrial Electronics*, August 1963.



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RADIO TOPICS

by Recorder

IN TECHNICAL CIRCLES THESE DAYS the with-it word seems to be "fluidics". Cambridge saw the the second Cranfield Fluidics Conference during 3rd to 5th January, and I note that the Ministry of Technology is currently encouraging development and experimental work in fluidics at the British Hydro-mechanics Research Laboratory and the National Engineering Laboratory.

The term fluidics describes a new method of obtaining automatic logic functions, and it is based on a very old observation. If you apply your thumb to the bathroom tap you can change the direction of the stream of water which emerges over a considerable angle, and with negligible physical effort. In fluidics, the direction of a stream of fluid, which can be gas or liquid, is altered under the influence of another stream having a much lower pressure. Thus, a jet of fluid can be transposed from one output tube to another. As can be imagined, this gives the first step towards switching and amplification, as we know these factors in electronics. In fluidics, however, the flow of current is replaced by a flow of fluid.

Fluidic Devices

At the moment, fluidic devices are being used as though they were analogues of electronic devices. Fluidic devices are referred to as flip-flops, AND gates, OR gates, NOR gates, Schmitt triggers and amplifiers, and they are coupled together by tubing to form basic operating circuits in just the same way as the corresponding electronic devices are coupled together by copper conductors. The fluidic operation in each device is obtained by carefully designing the shape of the channels, jets and chambers through which the fluids pass. There are even fluidic resistors, whose function is to provide a linear pressure-flow restriction. All these devices have nozzles for the application of interconnecting tubing, in the same way as electronic devices have terminals or lead-outs for interconnecting wiring.

The speed of operation of fluidic devices is in the order of milliseconds. This is a very much slower speed than is given by electronic components, and so it is unlikely that fluidics will ever offer any significant competition in the computer field. But fluidics have an important advantage in the sphere of machine

control and automated production line processes where a cycle of relatively simple operations has to be controlled. Pneumatic and hydraulic machine control equipment has been in use for a long time, and the addition of fluidic logic devices enables such equipment to offer more sophisticated control sequences. However, the main attraction of fluidic systems in the production field is that the components are extremely robust and are not affected by high temperatures or vibration levels. Because of this, some production engineers tend to look upon them with greater favour than on electronic systems, which seem to have gained an undeserved reputation for esoteric unreliability. It is a little ironic to think that the logic philosophy engendered by electronics has been the starting point from which the new fluidic systems—which will replace some electronic systems—have been able to develop.

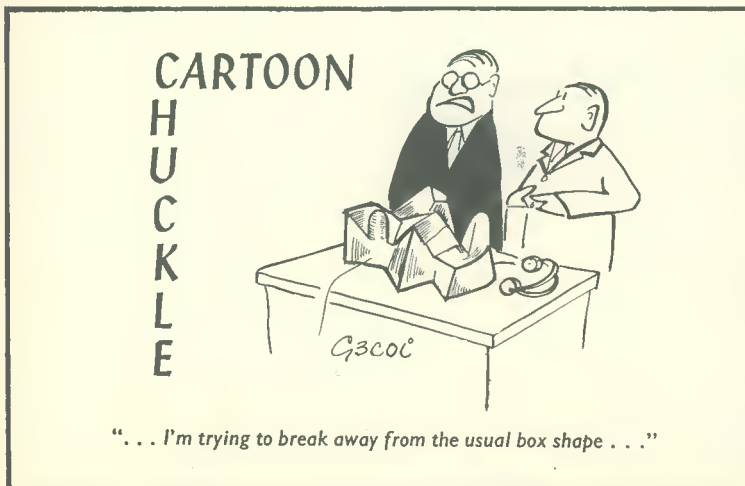
Much of the commercial development work in current fluidic devices has been carried out by Corning Glass Works in America, whose English representatives are ElectroSil Ltd., Sunderland, Co. Durham. There are a number of wholly British concerns engaged in commercial applications, notable amongst these being Techne (Cambridge) Ltd., who specialise in machine control.

Whatever other problems they encounter, students of fluidics will not, at any event, have any difficulties with regard to the direction in which the actuating fluid runs, as opposed to students of electronics who have to deal with a "conventional" flow of electric current which is dead opposite to the flow of the electrons which constitute the current!

I'm All Right, Jack

Turning to matters closer to the home-constructor, I note that a little matter of terminology seems to confuse some newcomers to our hobby.

This matter has to do with jacks and jack-plugs. The jack-plug is, of



course, the plug having a sleeve and tip to which earphones are normally connected. The jack is the socket into which the jack-plug is inserted.

For some reason there is a fairly widespread belief that the word "jack" applies to the plug, but this is quite incorrect and the term actually applies to the socket.

In most of the articles appearing in *The Radio Constructor* we skate over the problem by referring to the "jack plug" and the "jack socket". The component is thereby identified without any risk of ambiguity, even though the addition of "socket" is superfluous.

Some newcomers are also a little bewildered by "h_{fe}", with respect to transistors. This defines the small signal gain (actually "the small signal forward current transfer ratio with the output short-circuited to alternating current") of a transistor when it is connected in the common emitter mode, and is equivalent to the earlier terms α' (alpha "dash") or β (beta). All three mean the same thing and may be encountered in general electronic literature. The term h_{fe} is that which is most commonly used these days.

Arising from a conversation with a friend, there is a little further confusion over the American word "transconductance", as applied to valv.s. To make things a wee bit more unfamiliar, this transconductance is quoted in micromhos. In this particular context, transconductance (it should really be "grid-anode transconductance") is the Transatlantic version of our own mutual conductance (gm) which, you will remember, defines the change in anode current for change in grid voltage, the anode voltage being held constant. We express this as "mA per volt" and use the equation:

$$gm = \frac{dI_a}{dV_g} \quad (V_a \text{ constant}).$$

Now, from Ohm's Law we say that resistance is voltage divided by current whereas mutual conductance, from the equation, is current divided by voltage. A mho, which is the unit of conductance, is the reciprocal of an ohm (i.e. it is current divided by voltage) and so we could if we wished refer to mutual conductance in mhos. However, the figures turn out easier if we use micromhos, as the Americans do, whereupon a transconductance of 1,000 micromhos is the same as a mutual conductance of 1mA per volt. Just divide the transconductance micromhos figure by 1,000 and you have the mutual conductance in mA per volt.

Wide-Band Photophone

The "Wide-Band Photophone" articles by D. Bollen have represented an important step forward in the highly interesting field of communication by light. I now pass on a little further information from the author which arrived too late for publication in the articles.

Imported versions of the 90AV photocell specified have gone up in price recently and now cost the same (46s.) as the British-made type. It may be possible to locate a cheaper imported version, but Mr. Bollen has not been able to trace any at the moment. If constructors have difficulty in obtaining the 90AV, a 92AV may be employed instead. The 92AV supersedes the 90AV, has the same pin connections, and has very nearly the same characteristics.

New Photo-devices

And, finally, some notes on two new photo-devices which have been recently developed by American concerns. These are the General Electric "LASCS" and the Raytheon "Raysistor".

The word LASCS is an acronym for "Light Activated Silicon Controlled Switch". This device, like the silicon controlled rectifier, has a PNP structure, but can be switched to the conducting state by a weak beam of light as well as by an electrical pulse. Separate external connections are made to each of the four layers for maximum flexibility of circuit design. These planar passivated devices can switch currents of up to 175mA, the maximum voltage rating being 40 volts.

LASCS may be used in industrial equipment for switching, machine tool positioning, for computer tape and card reading, in calculators, in electronic cash registers, optical encoders, etc. Further details may be obtained from the manufacturers, General Electric Semiconductor Products Department, Electronics Park, Syracuse, New York, U.S.A.

The "Raysistor" consists of a small lamp and a photoconductive cell in a single encapsulation. The lamp may be either a gas discharge source (designated by the letter 'N') or an incandescent filament type (designated by the letter 'I'). The Raysistor provides complete isolation between the input and output. These four-terminal devices can be used as variable resistors, relays, solid state switches, etc. Further details can be obtained from the British agents of the Raytheon Company, Messrs. Walmore Electronics Ltd., 11-15, Betterson Street, Drury Lane, London, W.C.2.

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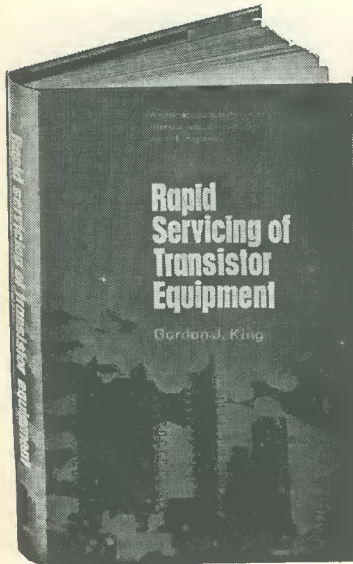
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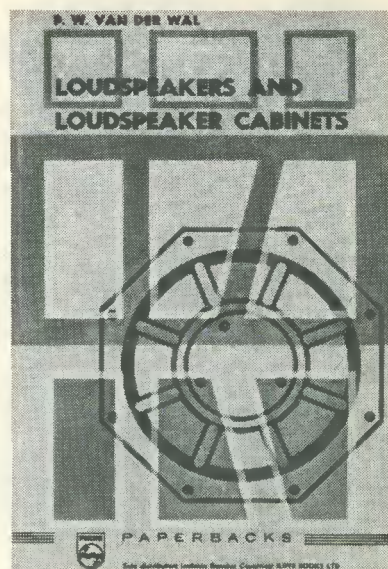
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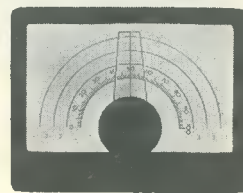
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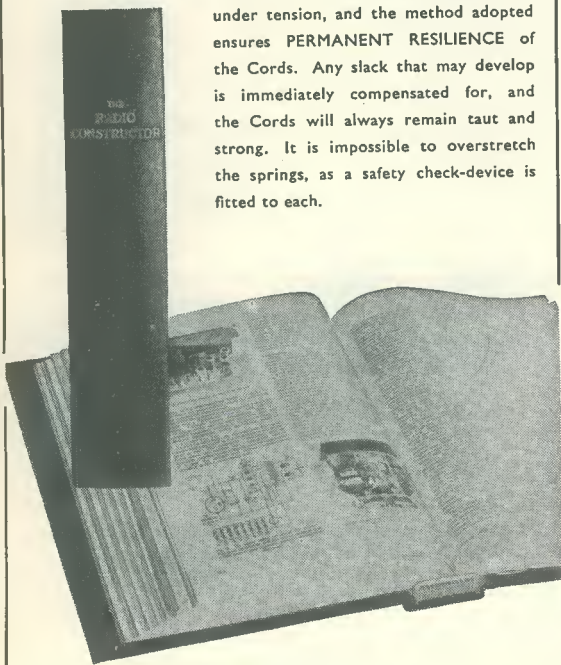
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continued from page 509

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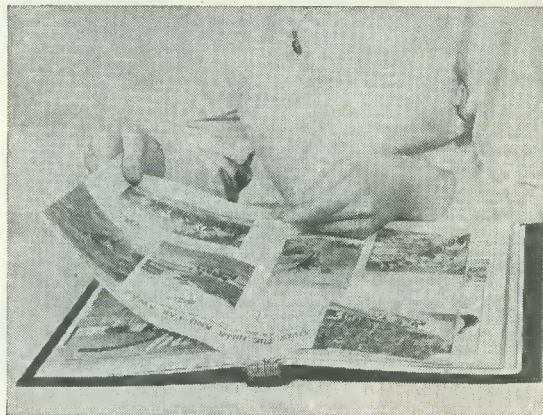
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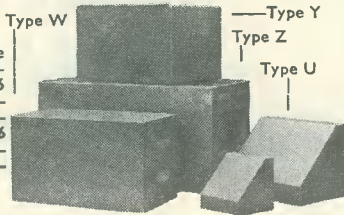
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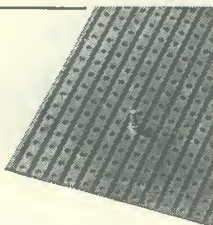
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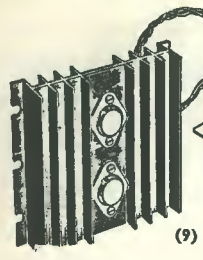
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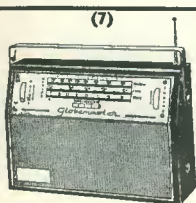
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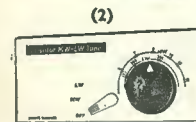
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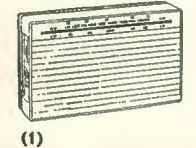
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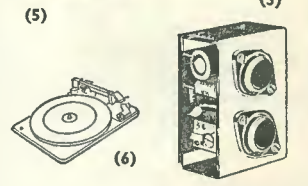
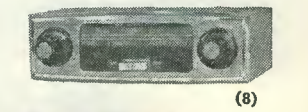
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