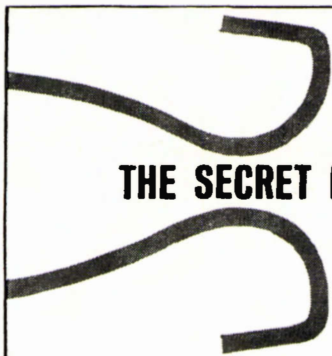
# Electronic Engineering

A MORGAN BROTHERS PUBLICATION

JUNE 1967 3s. 6d.

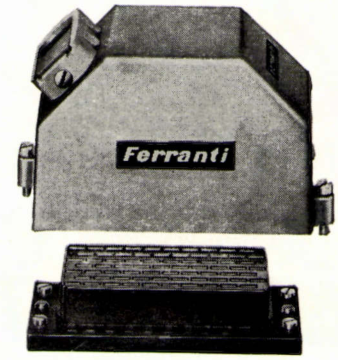
# CONTACT FERRANTI

## FOR THE BEST IN CONNECTOR DESIGN



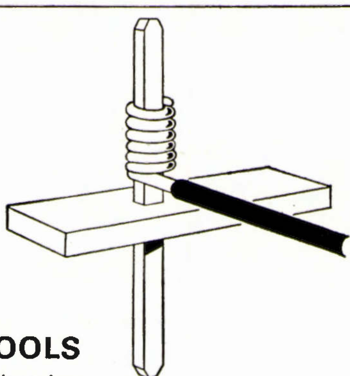

### EDGE CONNECTORS—EWD SERIES

The latest range of Ferranti Edge Connectors offer even greater reliability. The design incorporates a unique rolling-leaf spring contact, which has a low rate stress limiting characteristic, giving controlled contact pressure and very low insertion and withdrawal force.
AVAILABLE WITH 8, 16, 24, 32 or 40 POLE POSITIONS SINGLE OR DOUBLE SIDED CONTACTS.



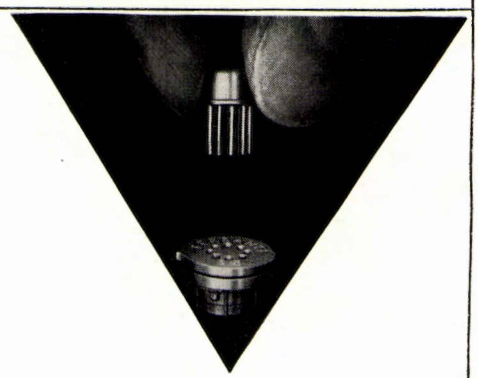

### CONNECTORS LFC SERIES

Ferranti LFC Connectors are designed for use either as rack and panel connectors, or, when used with the appropriate connector cover, as a free plug or free socket. LFC Connectors offer: UNSURPASSED RELIABILITY · VERY LOW INSERTION FORCE · LOW RATE FLOATING SPRING · CONTROLLED CONTACT FORCE.
Available in 35, 70 and 91 pole sizes.



### WRAPPING TOOLS

The Ferranti range of Wrapping Tools enable wrapped joints to be made quickly and easily with the minimum of staff training. Wrapped joints are the most reliable joints known, take less space and completely eliminate the possibility of damage caused by heat.
A full range of Hand and Power operated tools is available for making standard or miniature joints. Standard power tools are driven by compressed air and miniature power tools by electricity.



### CIRCUIT-SAVERS

Ferranti Circuit-Savers overcome the problem of replacing quickly and easily integrated circuits in printed circuit boards. The Circuit-Saver is inserted into the board and soldered into the circuit. Integrated circuits in 8 or 10-lead TO-5 encapsulations can then be inserted into the Circuit-Saver in a matter of seconds. Thus integrated circuits can be replaced quickly, without damage.

**FERRANTI**
*First into the Future*

FERRANTI LTD · KINGS CROSS ROAD · DUNDEE · SCOTLAND · Telephone: (ODU2) Dundee 89311

DS/T154/2

VOLUME 39 · NUMBER 472

# Electronic Engineering

*Incorporating*
ELECTRONICS, TELEVISION and SHORT WAVE WORLD
Managing Editor: H. G. FOSTER, *M.Sc., F.I.E.E., M.I.E.R.E.*

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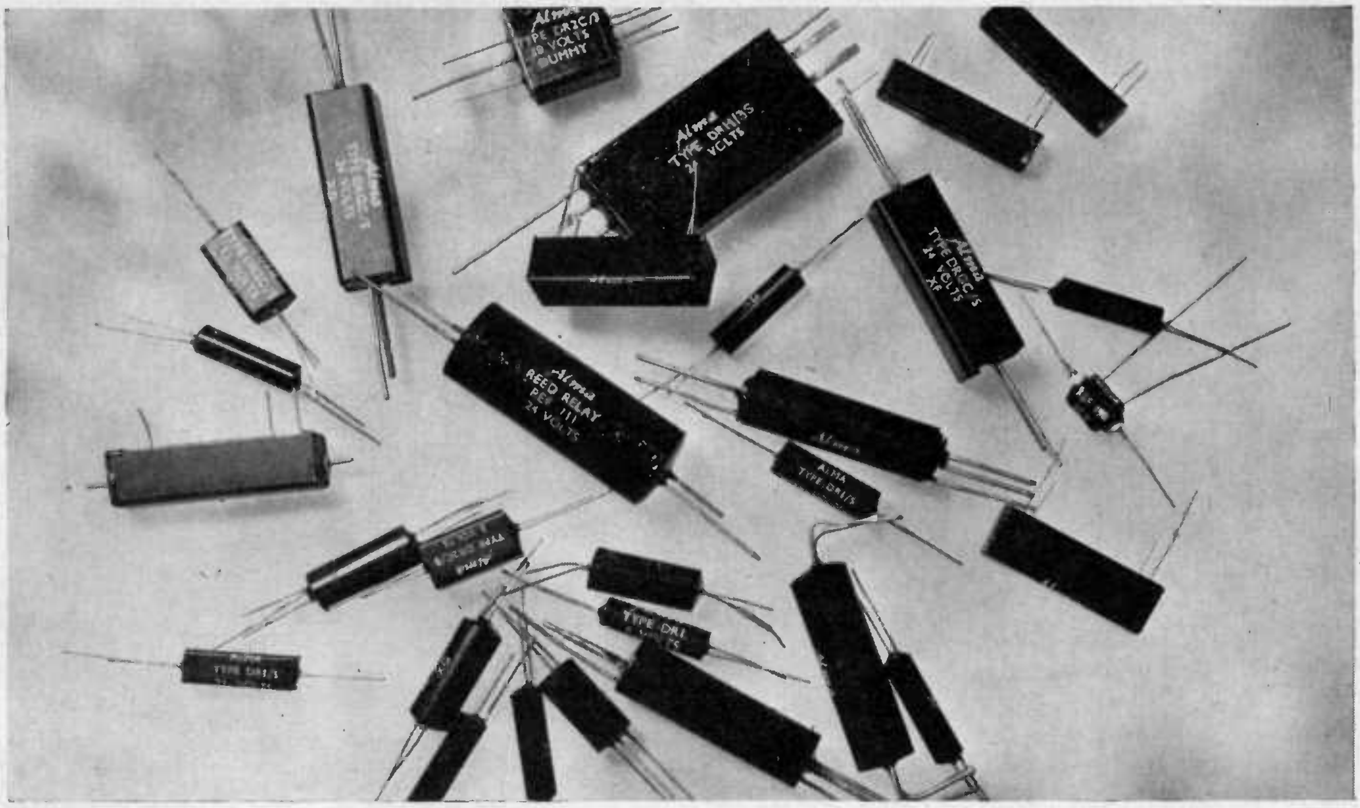
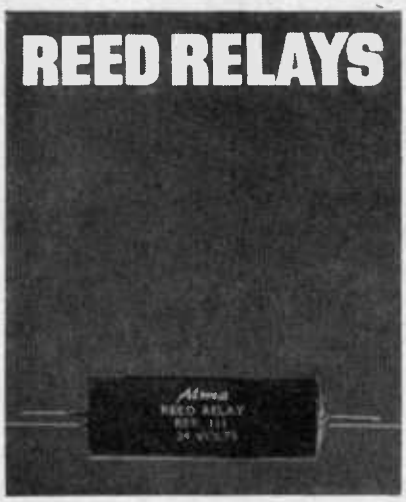
ABC MEMBER OF THE AUDIT BUREAU OF CIRCULATIONS



# Oneoffmanship



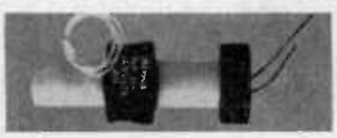

## Alma design nearly three hundred reed relays in three years

Many of these specials have now been incorporated into our standard range of relays. Alternatively, they have been adapted to form the basis of other specials. So if you have a call for a reed relay quite unlike anything you've heard of before—call Alma. Chances are we already have a design to suit exactly. And that will save you TIME!

## Typical examples


Very sensitive type - 25mW


Low capacitance - 1pF



High voltage - 25kV


High insulation - $10^{14}\Omega$

Not shown - Reed Uniselector for sequential switching

**ALMA COMPONENTS LIMITED**

Park Road · Diss · Norfolk · Telephone: Diss 2287 · Telex 98162

M.O.A. & A.R.B. Approved Inspection

For more information circle No. 9

# "There's a fortune in the idea-if only we had the capital to exploit it"

Sometimes it seems that the easiest stage of a real technical advance is the idea itself, and that the more difficult stages are its application, development and eventual marketing. For these stages finance is needed, and it is frustrating to conceive the idea and then be unable to raise the money to exploit it.

T.D.C. was formed to meet this need—to assist companies, associations or individuals who have made a substantial technical advance but are held up for finance at the commercial stage. If you think that T.D.C. could help, begin by sending for our booklet 'Capital for Technical Innovation'.

**Technical Development Capital Limited**

7 Copthall Avenue, London, E.C.2.

*A Subsidiary of Industrial and Commercial Finance Corporation.*

TDC

For more information circle No. 10

# By how much will FREON* cut the cost of your industrial cleaning?

# 10%? 50%? 80%?

Because it removes grease and oils without harm to metal, plastic or elastomeric parts, and has high density and low surface tension, FREON is the ideal cleaning solvent.

Du Pont makes over 40 different FREON fluorocarbons—all *tailored* to handle specific needs.

Many companies have reported dramatic savings in labour and solvent costs after using FREON solvents—and they have received the benefit of Du Pont's considerable technical experience in effectively analysing cleaning problems.

Find out how FREON solvents can give you better cleaning at lower cost by completing the coupon and posting it today.

To: **Du Pont Company (United Kingdom) Ltd.,**
'Freon' and 'Valclene' Products Department,
Du Pont House, Fetter Lane, London, E.C.4.

I am interested in receiving information on the advantages of using FREON solvents in cleaning

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*(State application)*

Name______________________________

Position______________________________

Company______________________________

Address______________________________

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**Enquiries from outside the U.K. and Eire should be addressed to: Du Pont de Nemours International S.A., 'Freon' and 'Valclene' Products Department, 81 route de l'Aire, Geneva, Switzerland.**

**FREON* SOLVENTS**

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Better Things for Better Living ...through Chemistry

* *Du Pont's registered trademark*

EE6/67

DP275

For more information circle No. 11

# meet a high class family that's not afraid of hard work. real hard work

First meet the 'baby' logger—the 620. He can handle up to four 4-digit measurements a second with $1\mu V$ resolution, and has floating differential inputs, insensitivity to noise, expandability and versatility.
Not enough? Try the 630 system—similar to the 620, but with added capabilities, fully programmable functions, 6 digits and up to six measurements a second.
Still too slow? Then go for the EI-85 system. This cool sophisticate can log up to 25 measurements a second—d.c. from $1\ \mu V$ to 750V, d.c./d.c. ratios, a.c. volts, ohms. He has a successive—approximation multimeter with $\pm 0{\cdot}01\%$ accuracy, plus a high CMR and active Bessel filter for noise protection.
Or perhaps you'd prefer to meet their close relations, the digital voltmeters . . . like the EI-85 which provides a full 5-digit measurement in 10 milliseconds with $\pm 0{\cdot}005\%$ accuracy, and the 630 which has 6-digit display and $1\mu V$ resolution.
These are just some of the Honeywell test gear family. Get to know them—and the others, by writing for details to
Honeywell Controls Limited, Test Instruments,
Great West Road, Brentford, Middlesex. ATLas 9191.

**Honeywell**

**Honeywell International:** Sales and Service offices in all principal cities of the world. Manufacturing in U.K., U.S.A., Canada, Netherlands, Germany, France, Japan.

For more information circle No. 12

# PULSE GENERATOR PG-2

## THE MOST VERSATILE PULSE GENERATOR IN ITS PRICE RANGE

VARIABLE RISE AND FALL TIMES
VARIABLE WIDTH AND DELAY
VARIABLE BASE LINE OFFSET
1 c/s TO 16 Mc/s.


MADE IN ENGLAND

**The type PG-2 is an all solid state pulse generator with true positive or negative single or double pulse outputs, offering the widest combination of PRF, pulse width, delay and rise and fall times in any single instrument in this price range.**

- PRF 1 c/s to 10 Mc/s (16 Mc/s double pulse)
- WIDTH 10 ns to 200 ms
- DELAY −10 ns to +200 ms
- RT & FT 10 ns to 20 ms
- OUTPUT 0 to 20 V (12 V into 50Ω)
- BASELINE SHIFT 0 to 4 V
- POSITIVE, NEGATIVE OR COMPLEMENTARY PULSE
- PANEL HEIGHT 3½″
- PRICE £380.

## CLAUDE LYONS LIMITED

Instruments Division,
Valley Works, Hoddesdon, Herts. *Hoddesdon 67161. Telex 22724*
76 Old Hall Street, Liverpool 3. *MARitime 1761. Telex 62181*

ID13

For more information circle **No. 13**

# *the family–*
# *a range of X–Y graphic recorders.*
# *the planning–by Honeywell*

Start with the 'baby', the 320M. Although it is low priced and measures only 12" x 19", the 320M is packed with big instrument features that make it an accurate, versatile and reliable choice. For example, it has a repeatability of ± 0·1%, 80 μV/cm sensitivity, and can have a built-in time base.

Then there's the 520, a high sensitivity recorder (50 μV/cm) with isolated inputs and potentiometric input impedance for use with delicate signal sources. It has an accuracy of ± 0·15%, internal time bases on both axes and remote control facilities.

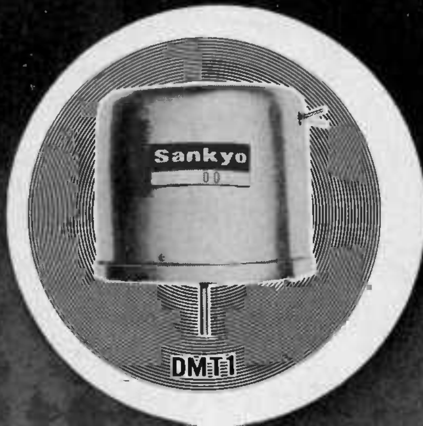
For plug-in module versatility, there's the top performance 400. In one or two (X-YY') pen versions. This instrument has been designed to make 'obsolescence' a forgotten word !

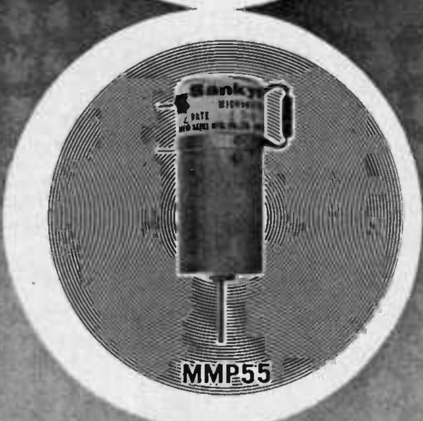
The full story about the family of X-Y recorders is yours for the asking: simply drop a line to Honeywell Controls Limited, Test Instruments, Great West Road, Brentford, Middlesex. ATLas 9191.

**Honeywell**

**Honeywell International:** Sales and Service Offices in all principal cities of the world. Manufacturing in U.K., U.S.A., Canada Netherlands Germany, France, Japan

# See For Yourself By Testing Sankyo's Family Of Micro Motors!




**DMT1**.....,.Single speed (2400 rpm). With flat sealed case.
**MMS51**.....Single speed (300 rpm). For high grade tape recorders.
**MMP55**.....2 speeds (2700,4800 rpm). For 8mm cameras.
**MMZ6**...... Without governor. For still and cameras.

**SPECIFICATION FOR SANKYO MICRO MOTORS**

| | Dimension (mm) ϕ Length | Rated Voltage (V) | Range of Voltage (V) | Rated Torque (gr-cm) | Rated Speed (rpm) | No Load Current (mA) | Load Current (mA) | Starting Torque (gr-cm) | Life (Hr) |
|---|---|---|---|---|---|---|---|---|---|
| DMT1 | 42×37 | 6 | 4.5— 6 | 9 | 2400 | 40 | 130 | 25(4.5V) | 600 |
| DMY16 | 42×37 | 6 | 4.5— 6 | 15 | 2400 | 50 | 200 | 50(4.5V) | 600 |
| DMY81 | 42×37 | 12 | 8 —12 | 20 | 2400 | 50 | 150 | 50(8V) | 600 |
| MMS44 | 25×55.5 | 9 | 6 —10 | 10 | 3000 | 40 | 140 | 20(6V) | 600 |
| MMS51 | 25×55.5 | 6 | 4.5— 6.3 | 7 | 3000 (2900) | 50 | 140 | 25(4.5V) | 1000 |
| MMP55 | 20×45 | 4.5 | 4 — 6 | 10 | 2700 4800 | (110) | (290) | 60(4.5V) | 50 |
| MMZ6 | 16×29 | 4 | 4 — 6 | 2 | 6000 11500 | (100) | 120 | 15(4V) | 100 |

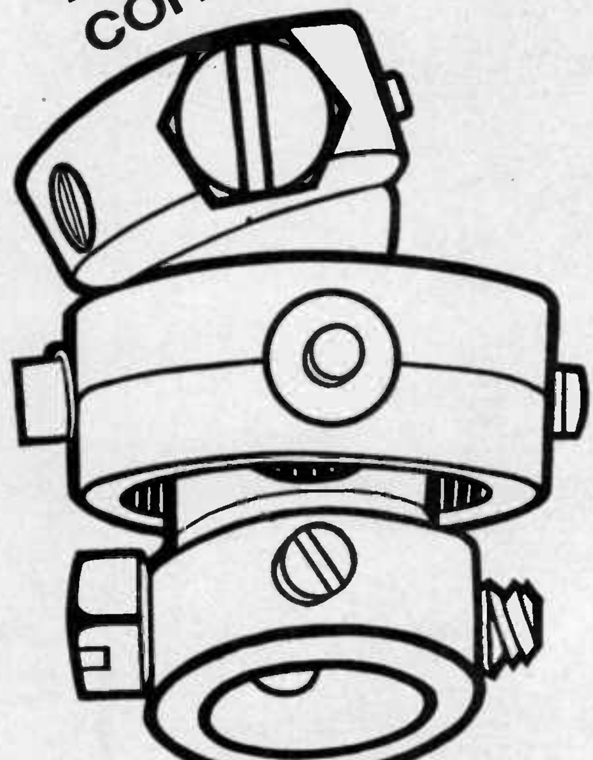
# Sankyo

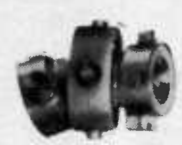
**Motor Division**

**SANKYO SEIKI MFG. CO., LTD.:** 17-2, Shimbashi 1-chome, Minato-ku, Tokyo, Japan. Tel: 591-8371 Cables: SANKYORGEL TOKYO
**AMERICAN SANKYO CORP.:** Rm. 808-10, 95 Madison Ave., N.Y.C., N.Y., U.S.A. Tel: LE-2-8020
**SANKYO (EUROPE) EXPORT und IMPORT G.M.B.H.:** 4 Düsseldorf, Bahnstraße 45-47, W. Germany. Tel: 25652/3



# A new angle in-direct control





TYPE A

TYPE 1B

*(British Patent No. 864992— British Foreign patent)*

## UNIVERSAL MINIATURE COUPLINGS

The extensive range of Oxley precision made mechanical spindle couplers have been specially designed for leading out shafts and spindles from inconveniently mounted components.

They are unique in utilising P.T.F.E. as bearings. Thus reducing friction and consequently eliminating wear and backlash to a minimum ensuring long life.

The units can be supplied with taper and grub screws or with socket clamp arrangement.

The shaft sizes range from .125" to .250"—other shaft sizes can be accommodated upon request.

*For full technical particulars write to*

**OXLEY®**
**DEVELOPMENTS COMPANY LIMITED**
*Priory Park, ULVERSTON, North Lancs.*
*Tel. Ulverston 2621*

For more information circle No. 18

# what's a little one doing in an advertisement about big Daddy?

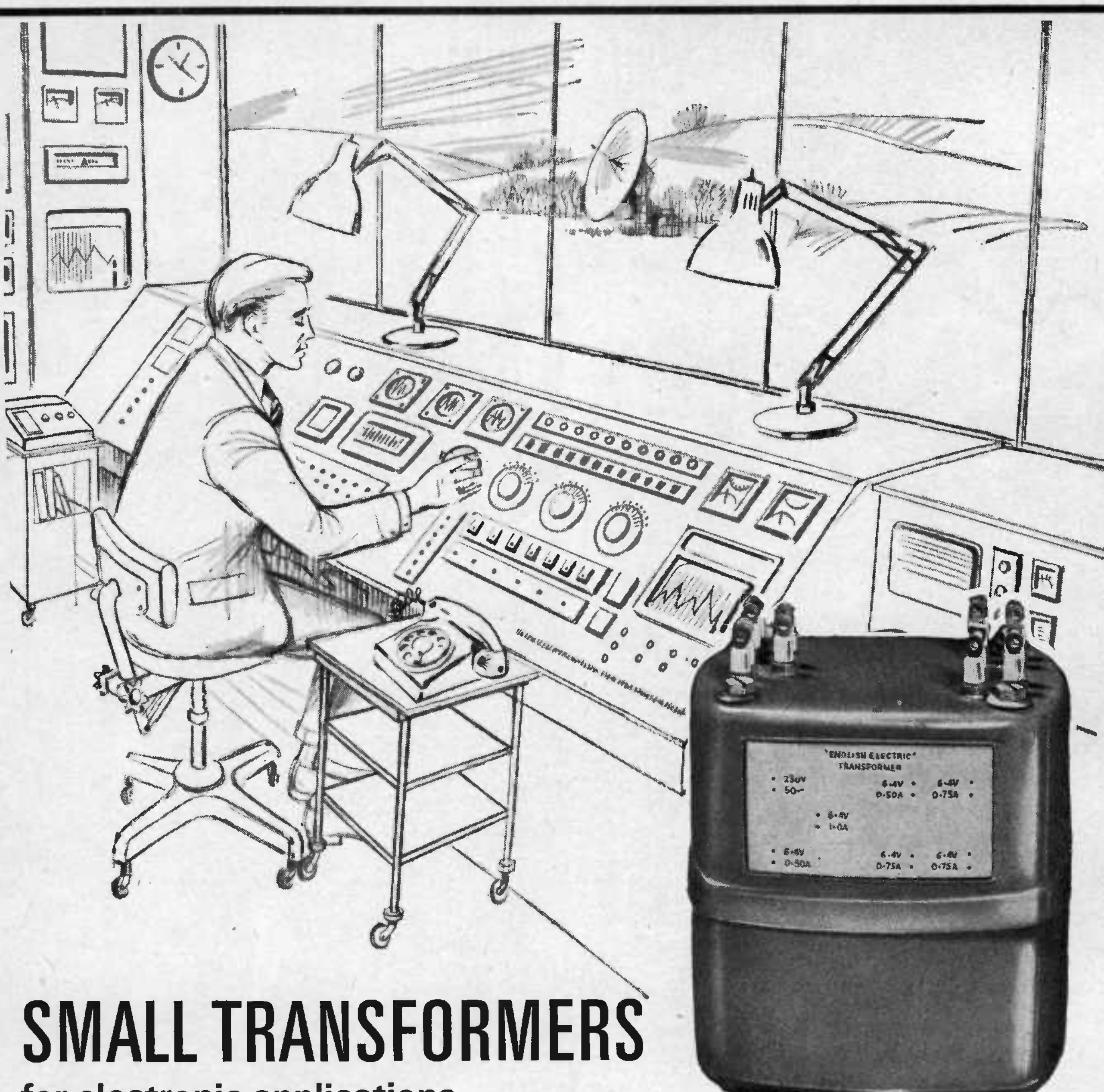
You can't beat family pride ! So when we decided to show you the 2106 Visicorder, we thought let's bring the little one in as well.
The 1706 was after all the first ever low cost UV recorder, and what a fine picture it makes with its newly added electronic flash timing, grid lines and 5 sets of 8 paper speeds . . .
But it's the 2106 that's the real big Daddy of Honeywell's oscillograph family.
For a heart it has the world's best galvos—range 0 to 13,000 Hz. Its 12 (+ 2 auxiliary) channels, chart speeds up to 300 cm/s and full width electronic time and grid lines add up to a pretty powerful versatility.
. . . And it's not as expensive as it looks ! The 2106 is just one of the Visicorder family about which we'd like to tell you more. We will gladly add your name to the mailing list if you drop a line to Honeywell Controls Limited,
Test Instruments, Great West Road, Brentford, Middx. Atlas 9191.

**Honeywell**

**Honeywell International:** Sales and service offices in all principal cities of the world. Manufacturing in U.K., U.S A., Canada, Netherlands, Germany, France, Japan.

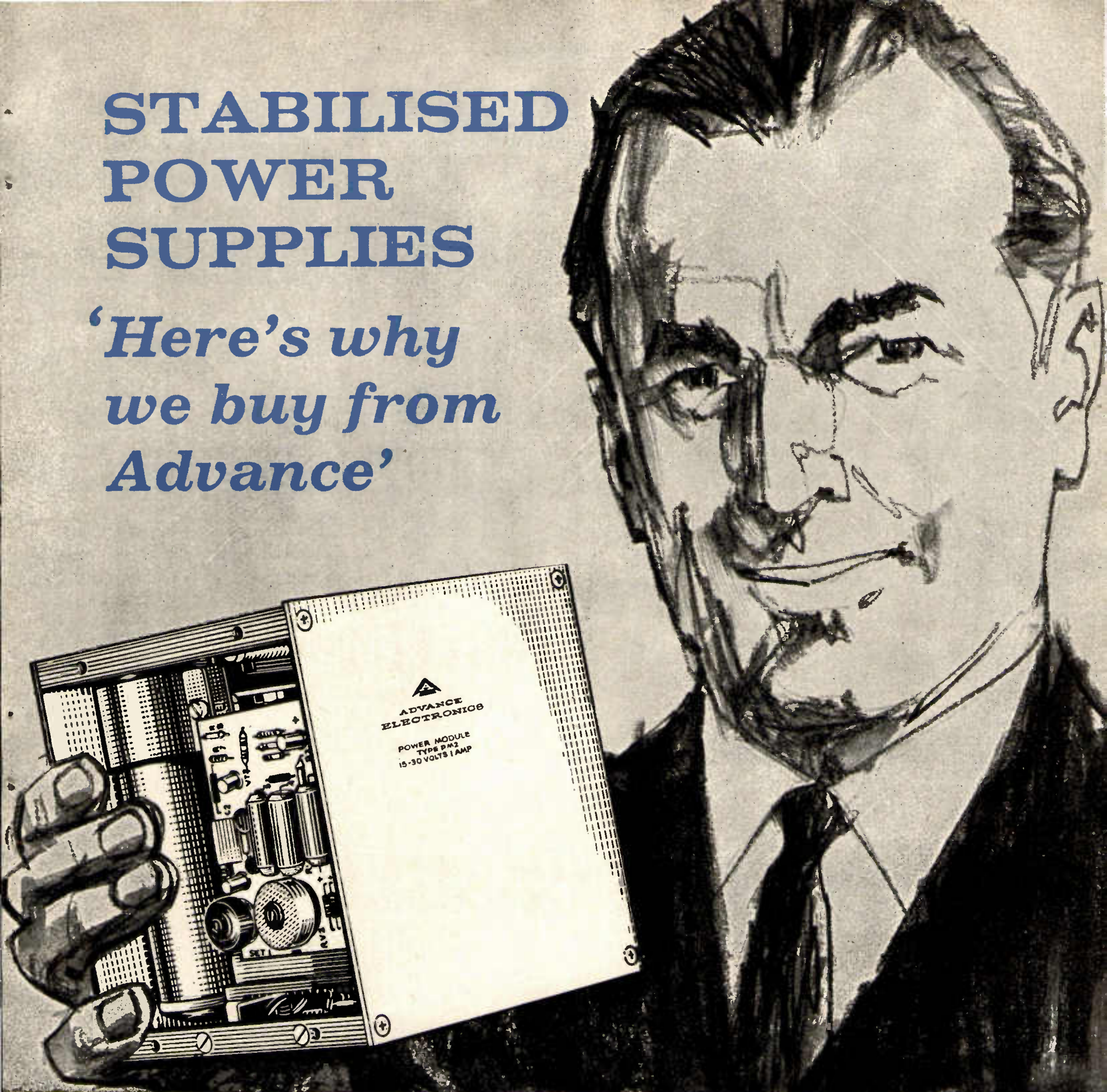

YOU
helped
to design
this range of
power supply units...
SO THEY MUST BE GOOD!
S.B. RANGE VARIABLE OUTPUT D.C. STABILISED POWER UNITS. Based on proven electronic circuitry, incorporating a current limiting protection circuit and designed with the emphasis on reliability.
SB 30/10 Variable 0.30 Volts at 10 Amps. U.K. PRICE £105
SB 30/5 Variable 0.30 Volts at 5 Amps U.K. PRICE £78
SB 30/2 Variable 0.30 Volts at 2 Amps. U.K. PRICE £58
Farnell
INSTRUMENTS LTD.
SANDBECK WAY WETHERBY
YORKSHIRE TEL. 2691/2/3/4

*For more information circle* **No. 23**



# STABILISED POWER SUPPLIES

## *'Here's why we buy from Advance'*




We insist on exact specifications. We demand real assurance of service. We get firm quotations, and we make ruthless comparisons. Advance come out on top every time, with their outstanding range of modular power supply units and their extraordinary skill and enthusiasm when it comes to designing specials. Before you place an order for power supplies, at least find out what Advance have to offer.

**ADVANCE ELECTRONICS LIMITED**

Volstat Division, Roebuck Road, Hainault, Ilford, Essex. Telephone: 01-500 1000. Grams: Attenuate, Ilford.

*For more information circle* **No. 24**

AV2.

B

For more information circle No. 28

# MEGOHM METER


## Type 2565

***For MAXIMUM reliability***

**The modern method of measuring insulation in workshop, laboratory and on location.**

- No skill required: Just press keys and read Megohms.
- Widely used in schools, factories, Laboratories, Colleges, Government Establishments.
- Standard Model reads 20M to 50,000M @ 250-1000 volts; Lower and Higher Resistance & Voltage Ranges to Order.
- Mains Operated: Stabilised Voltage: Will read insulation Resistance of Capacitors.
- Proven Circuitry. Gives years of trouble-free Service.
- Easily Portable. Weight 14½lb. Dimensions 12½×8½ ×9 inches. Steel Case.
- Alternatively available in Polished Teak Case with Lid and Carrying Handle.

*Write or phone for full details to*

**ALLIED ELECTRONICS LTD**

*Telephone* VANDYKE 1856

28 UPPER RICHMOND ROAD, LONDON, S.W.15

*All units actual size*

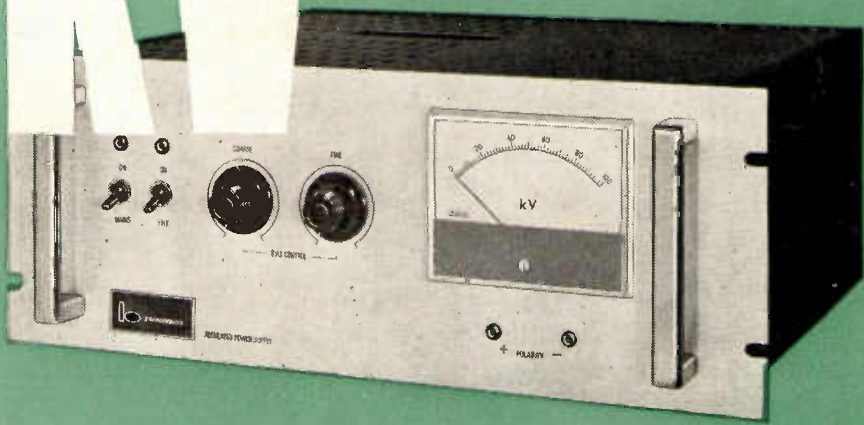

**SERIES 1″—600**
With wide choice of characters, variable grouping, and colour; a really flexible display unit. 12 Lamp positions.

**SERIES 3¾″—650**
For prominent display of single numbers or combination of words, digits and symbols. 12 or 27 Lamp positions. Easily read at any angle.

For more information circle No. 29

*For more information circle* **No. 30**

# now 100kV



Two new models now extend the Beta Range of stabilised high voltage dc power supplies to 100kV. The new models, 910 and 911, offer positive and negative polarity respectively and provide an output, via a special Brandenburg screened connector, of 250μA at 10kV to 100kV within a 7 inch high x 19 inch front panel.

Despite the high voltage output from a unit having such compact proportions, no compromise in reliability or performance has been made. Very special attention has been given to the design of the high voltage head and encapsulated components, areas where Brandenburg's experience and production facilities are unrivalled.

**BRIEF SPECIFICATIONS**

**Output**
Model 910: 10kV to 100kV at 250μA positive
Model 911: 10kV to 100kV at 250μA negative

**Stability**
0.1% at ± 7½% mains change

**Regulation**
0.25% zero to max load at full output

**Ripple**
Less than 0.3% peak to peak

**Input**
200-250V 50/60cps or 100-125V to special order

**Price**
**£345** complete with 9′ of screened output cable

**Other models of the Beta Range. (Identical in size and appearance)**

**BRIEF SPECIFICATIONS**

**Output**
Model 705: 2mA at 1.5kV to 15kV } Reversible
Model 800: 1mA at 3kV to 30kV } Polarity
Model 905: 400μA at 6kV to 60kV Positive
Model 906: 400μA at 6kV to 60kV Negative

**Stability**
0.1% at ± 7½% mains change

**Regulation**
0.25% zero to max. load at full output

**Ripple**
0.1% peak to peak (Models 905, 906, 0.16%)

**Price**
Models 705 and 800 **£150.** Models 905 and 906 **£245**

Literature, with full details of these and all other models, is freely available from

**brandenburg**

139 SANDERSTEAD ROAD, SOUTH CROYDON, SURREY
POSTAL CODE CR2 OYP Tel: SANDERSTEAD 0225

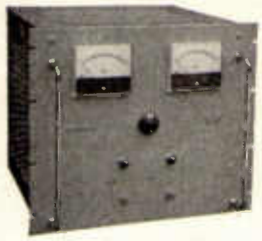
# and still 7" HIGH

and the full range of high voltage power supplies remains available

# brandenburg

### Model 501 Electrophoresis supply £525

| | |
|---|---|
| Output | 0-5kV, continuously variable at 500mA max. |
| Metering | Both voltage and current. |
| Regulation | Zero to full load at max. output: 9%. |
| Input | 380/440 volts 3 phase and neutral 50 c/s. |
| Protection | Mains fuses and overload trip. |
| Facilities | 220v single phase outlet to drive hour clock. External and zero current interlocks. Overcurrent relay trip. |



### Models 620, 750 and 803 £285 each model

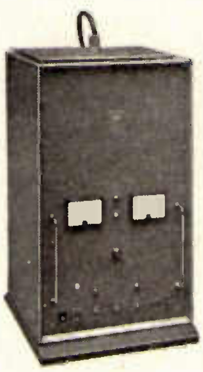
| | |
|---|---|
| Output | Model 620 1-10kV at 10mA. Model 750 2-20kV at 5mA. Model 803 3-30kV at 3mA. |
| Polarity | Positive or Negative (not reversible). |
| Ripple | Less than 12 volts Peak to Peak. |
| Stability ±10% Mains Change | Better than 0.5%. |
| Protection | Preset overload cut-out. Mains and H.T. Fuses. |
| Regulation Zero to Full Load | Both Models better than 0.5% at full output. |
| Input | 200/250 volts 50 c/s. |
| Size | 19" wide x 17" deep x 15" high (cm. 48 x 43 x 38). |

### Model 825 £745

| | |
|---|---|
| Output | 4 to 40kV continuously variable. |
| Output Current | 5mA max. |
| Polarity | Positive or Negative (NOT Reversible). |
| Ripple | Less than 40 volts. |
| Stability, ±7% Mains Change | 0.5%. |
| Effective Resistance | Less than 20 Kilohms. |
| Voltage Control | Potentiometer on Front Panel. |
| Metering | Voltage and Current. |
| Protection | Fuses and overload trip. |
| Input | 200/250 volts. 50 c/s. |
| Size | 24" x 22" x 44" (cm. 61 x 56 x 112). |

### Model PM2500R £149

| | |
|---|---|
| Output | 100-2,500 volts d.c. |
| Output Current | 5mA d.c. maximum. |
| Polarity | Reversible. |
| Ripple | Less than 10mV. |
| Stability ±7% Mains Change | Not worse than 0.03%. |
| Drift with Time | Less than 0.01% per hour. |
| Voltage Control | 5 Switched Ranges with overlapping fine control. |
| Input | 200/250 volts 50 c/s. |
| Size | 19" wide x 17" deep x 7" high (cm. 48 x 43 x 18). |

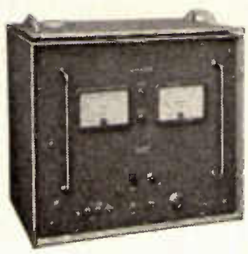


### Model MR100R £585

| | |
|---|---|
| Output | 10kV to 100kV. |
| Output Current | 0 to 1mA. |
| Polarity | Reversible. |
| Ripple | Less than 0.1% at full output. |
| Stability, ± 10% Mains Change | Better than 0.1%. |
| Protection | Current Overload Cut-out, Preset Overvolt Trip, H.T. Line Fuse, Mains Fuses. |
| Regulation Zero to Full Load | Better than 1%. |
| Input | 200/250 volts 50 c/s. |
| Size | 24" wide x 19" deep x 40" high (cm 61 x 48 x 102). |

### Model MR50HS £785

| | |
|---|---|
| Output | 5kV to 50kV. |
| Output Voltage Control | 10-turn Calibrated Helical Potentiometer. |
| Output Current | 0 to 1mA. |
| Polarity | Reversible. |
| Ripple | Less than 2 parts in $10^5$. |
| Stability ± 10% Mains Change | Better than 2 parts in $10^5$. |
| Effective Resistance | 60 Kilohms. |
| Drift-Short Term | Better than 4 parts in $10^6$ over 15 minutes. |
| Drift-Long Term | Better than 1 part in $10^4$ over 24 hours. |
| Protection | Pre-set overload cut out. |
| Input | 200/250 volts 50–60 c/s. |



### Model MR50R *(above)* £270

| | |
|---|---|
| Output | 5kV to 50kV. |
| Output Current | 0 to 1mA. |
| Polarity | Reversible. |
| Ripple | 20 volts P to P. |
| Stability, ± 10% Mains change | Better than 0.5%. |
| Protection | Pre-set overload cut out. Mains and H.T. fuses. |
| Regulation Zero to Full Load | Better than 0.5% at full output. |
| Input | 100/125 volts and 200/250 volts 50–60 c/s. |
| Size | 24" wide x 18" deep x 23" high (cm 61 x 46 x 58). |

For more information circle **No. 32**

# How far do you go to ensure quality and reliability?

# Stop right here!

ACTUAL SIZES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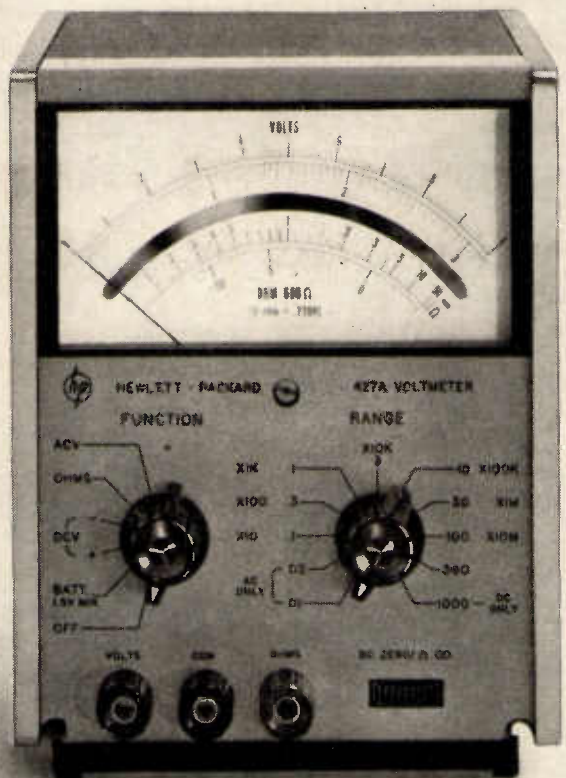


*The model 427A is an all-solid-state, multi-function meter for measuring Voltage (AC or DC) and resistance. With it you can make AC voltage measurements over a choice of ten ranges, DC over nine ranges and resistance over seven ranges.*
*This instrument was designed for broad application in any laboratory, production line or service department—or in the field. It operates off battery or mains, depending on your needs.*
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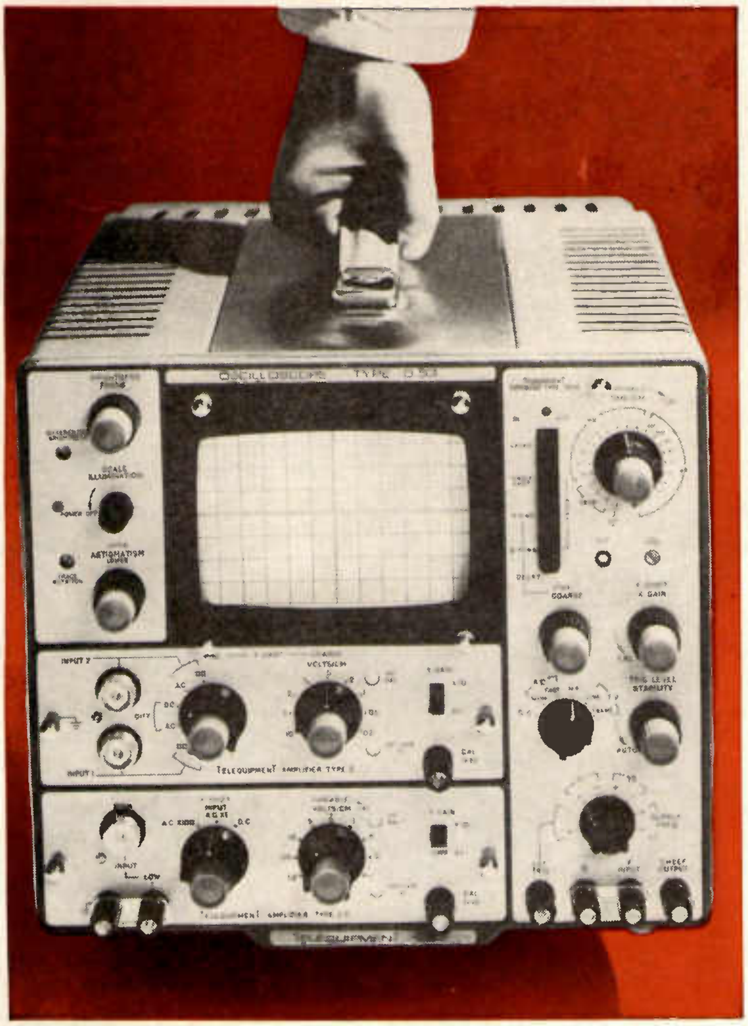

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*Typical Characteristics:*
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10mA : 10mA : 1mA = 2nS
Stored charge at 10mA = 35pC

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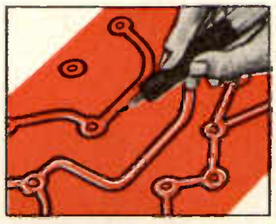
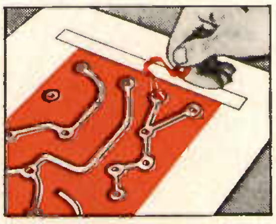


With sharp blade, outline the areas to be masked. Do not cut through the backing sheet. The Ulano Swivel Knife does the job quickly, easily.



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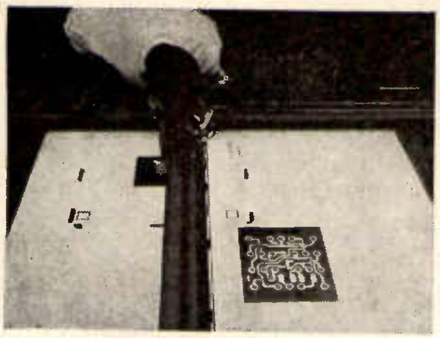

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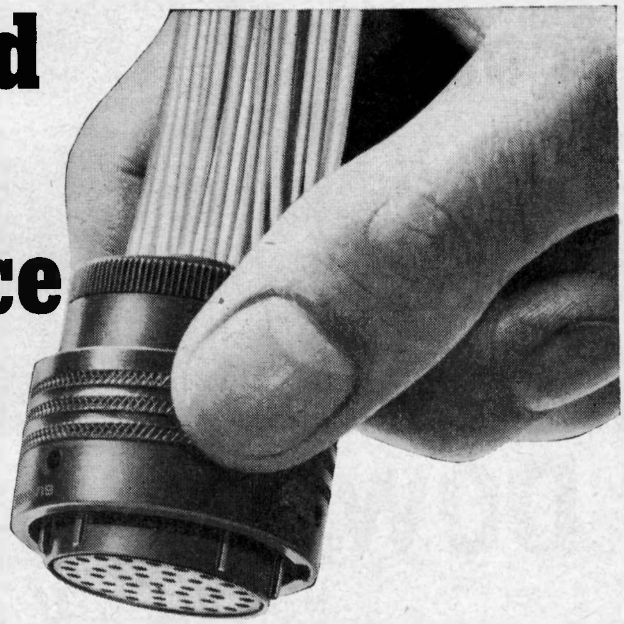
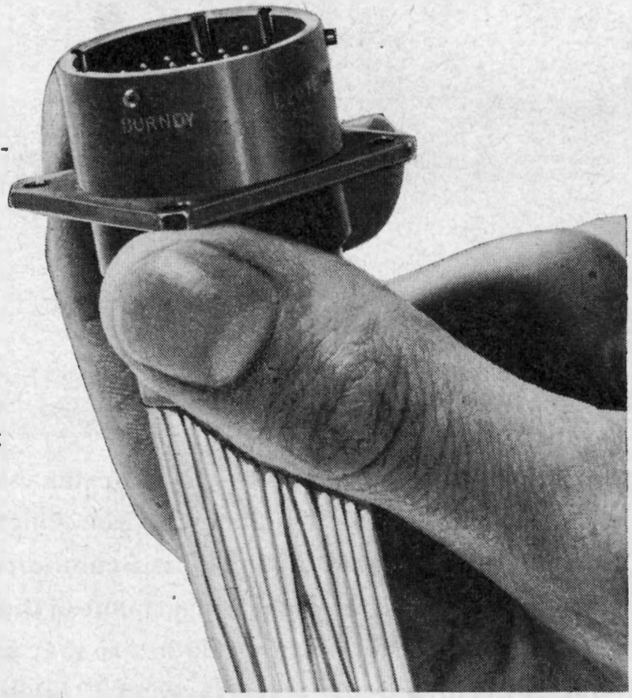


It's the **Bantam**

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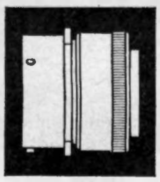
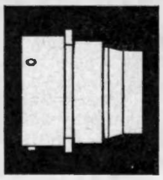
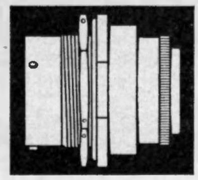


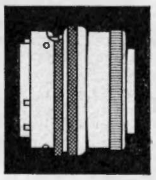
0 Wall mounting flange receptacle



2 Box mounting flange receptacle



4 Jam nut rear mounting receptacle



6 Straight cable plug

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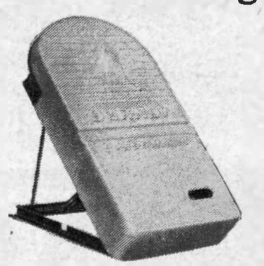

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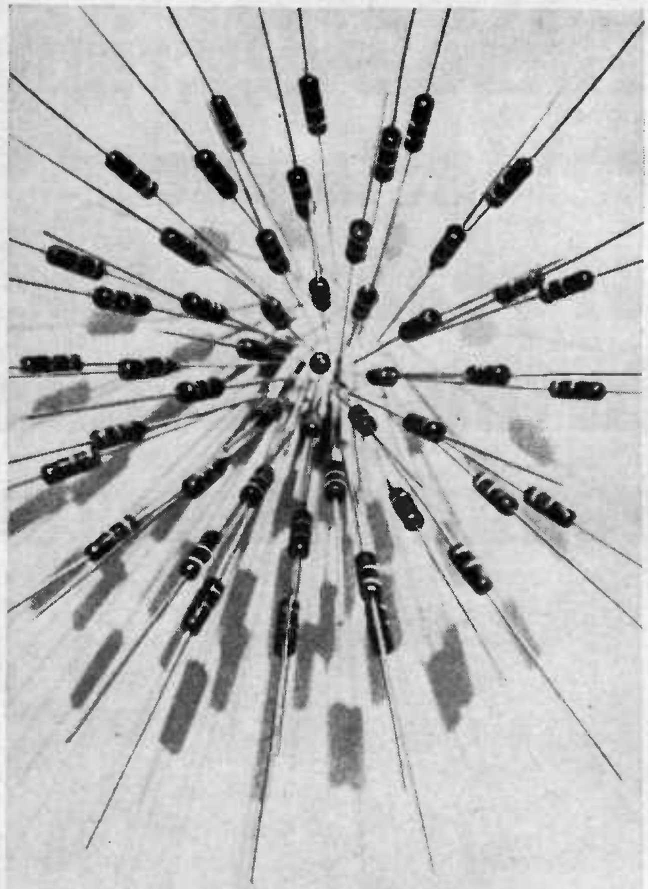

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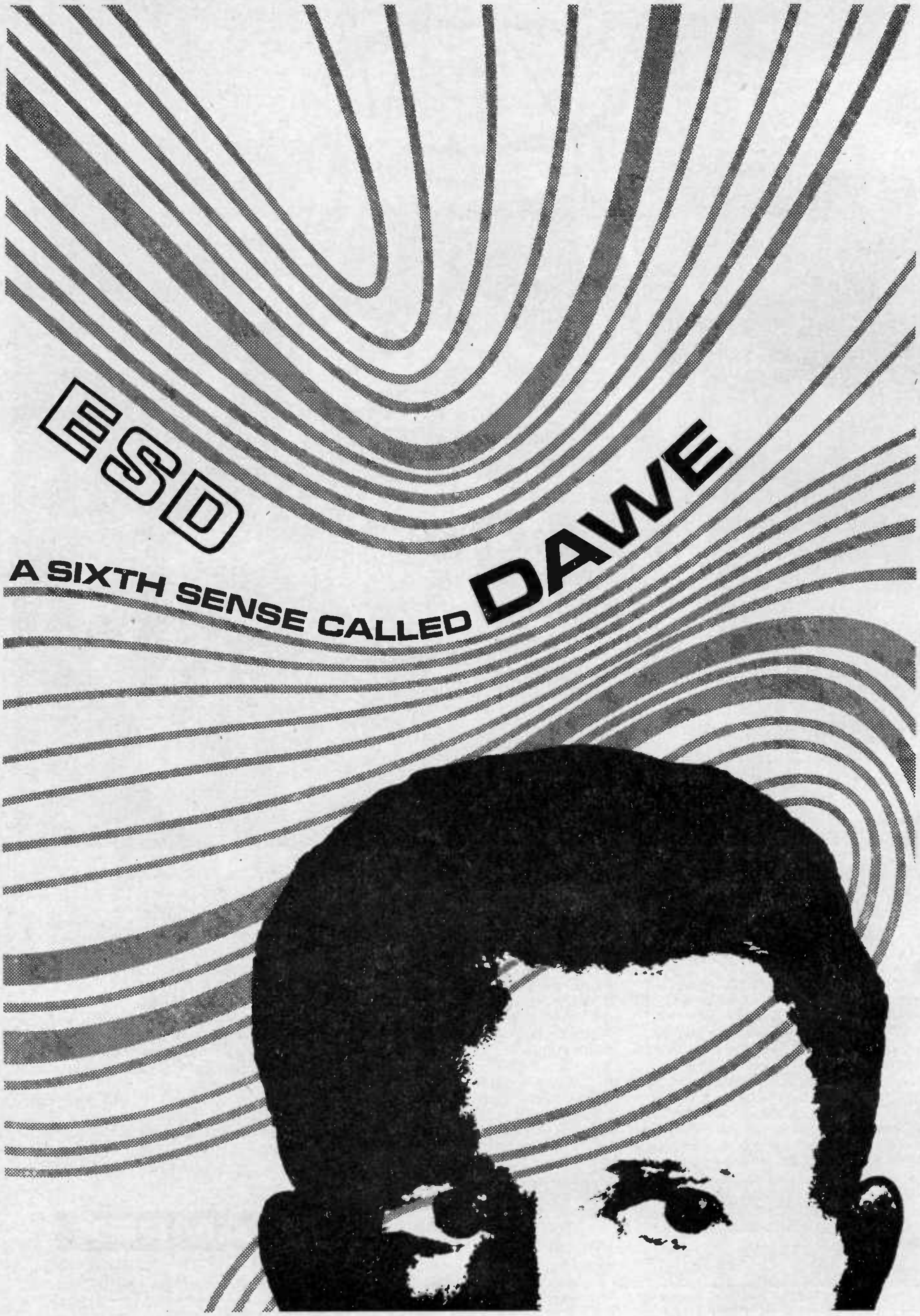
PAM EML 8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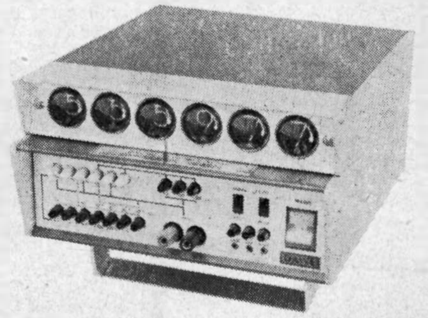

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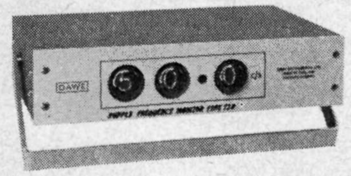


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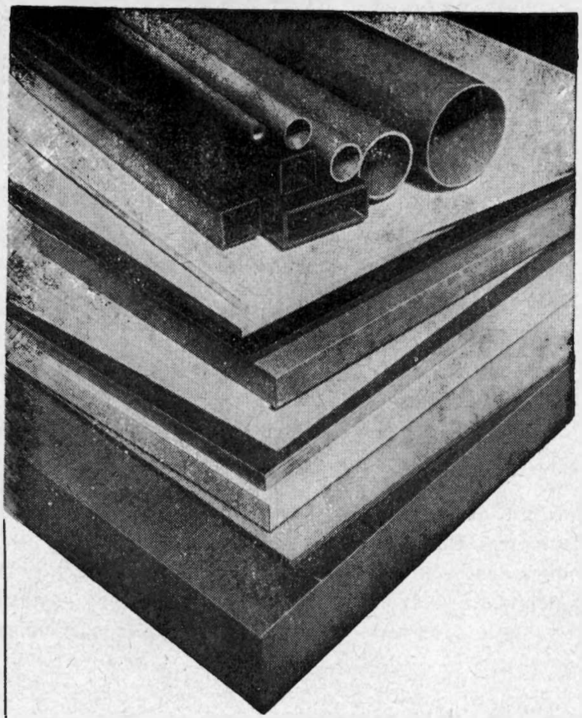

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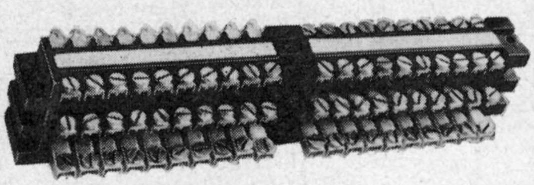

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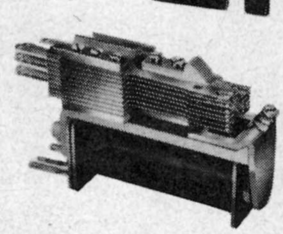

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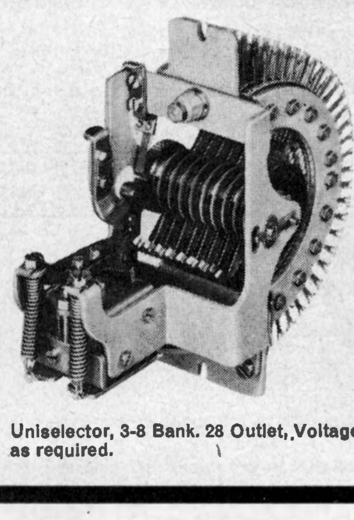

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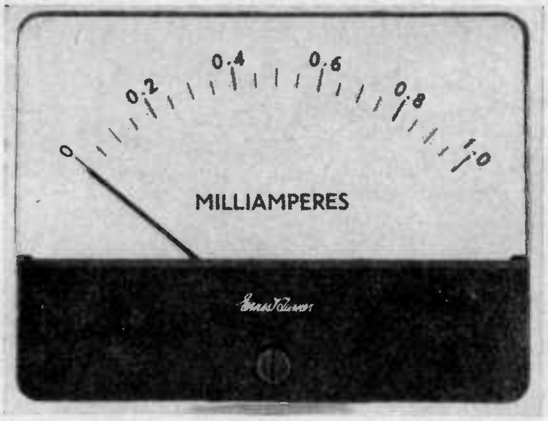


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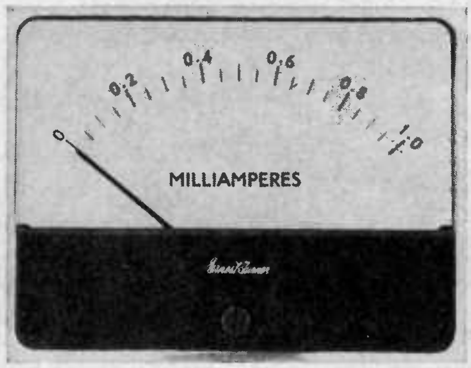


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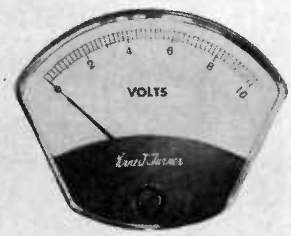
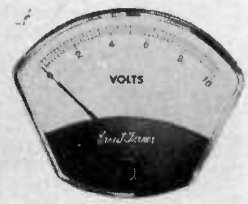



MODEL 403


MODEL 643

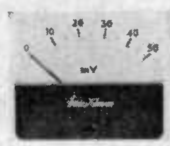

MODEL 6425


MODEL 642




MODEL 402


MODEL 641



MODEL 640

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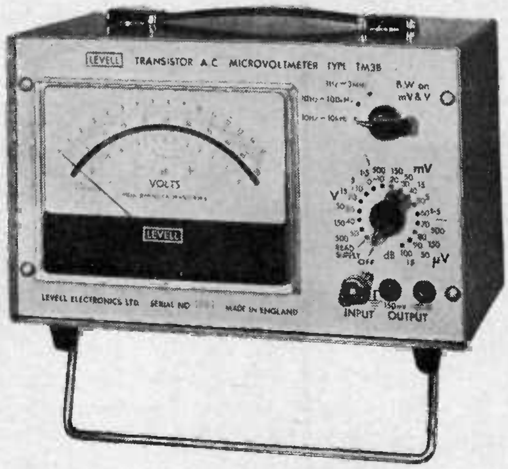


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**A.C. Power Unit £7.10.0.**

**VOLTMETER RANGES**
$15\mu$V, $50\mu$V, $150\mu$V . . . . 500V f.s.d. Accuracy $\pm1\%$ $\pm1\%$ f.s.d. $\pm1\mu$V at 1kHz.

**dB RANGES**
$-100$dB to $+50$dB in 10dB steps. Scale $-20$dB to $+6$dB. 0dB=1mW into 600Ω.

**FREQUENCY RESPONSE**
Above $500\mu$V: $\pm3$dB from 1Hz to 3MHz.
$\pm0{\cdot}3$dB from 4Hz to 1MHz.
On $500\mu$V: $\pm3$dB from 2Hz to 2MHz.
On $150\mu$V: $\pm3$dB from 4Hz to 1 MHz.
On $50\mu$V: $\pm3$dB from 7Hz to 500kHz.
On $15\mu$V: $\pm3$dB from 20Hz to 200kHz.

**AMPLIFIER OUTPUT**
150mV at f.s.d. on all ranges. Will drive a load of 200kΩ and 50pF without loss.

**POWER SUPPLY**
One type PP9 battery, life 1000 hours; or, A.C. mains when Power Unit is fitted.

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### BROADBAND VOLTMETERS

As A.C. Microvoltmeters plus H.F. probe to extend response to 450MHz. Two versions differ only in meter size and L.F. bandwidth switch on type TM6B.

**TYPE TM6A**


**£85**

**complete with battery and input lead**
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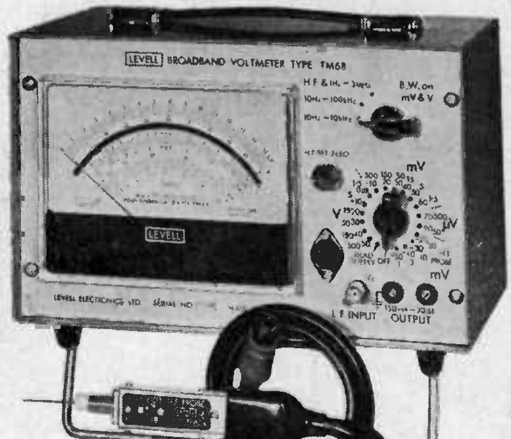
**Leather Case £4.10.0.**
**A.C. Power Unit £7.10.0.**

**TYPE TM6B**




**£99**

**complete with battery and input lead**
**OPTIONAL EXTRAS**
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**H.F. VOLTAGE RANGES**
1mV, 3mV, 10mV . . . . 3V f.s.d. Square law scales. Accuracy $\pm4\%$ of reading $\pm1\%$ of f.s.d. at 30MHz.

**H.F. dB RANGES**
$-50$dB, $-40$dB, $-30$dB . . . . $+20$dB. Scale $-10$dB to $+3$dB. 0dB=1mW into 50Ω.

**H.F. RESPONSE**
$\pm0{\cdot}7$dB from 1MHz to 50MHz.
$\pm3$dB from 300kHz to 400MHz.
$\pm6$dB from 400MHz to 450MHz.

**L.F. RANGES**
As TM3A and TM3B except for the omission of $15\mu$V and $150\mu$V.

**POWER SUPPLY**
One type PP9 battery, life 1000 hours on L.F. ranges and 400 hours on H.F. ranges; or A.C. mains when Levell Power Unit fitted.

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*For more information circle* **No. 57**

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| Output Current: | up to 2 Amps. |
| Protection: | Re-entrant current, any level up to 2 Amp and selected by a single external resistor. |
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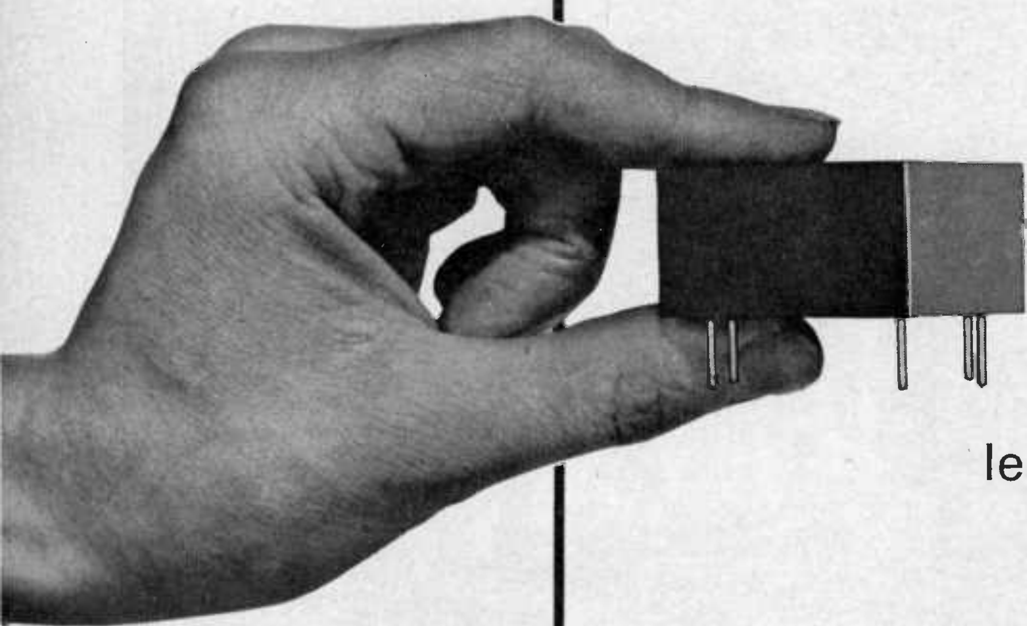

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## the superlative range

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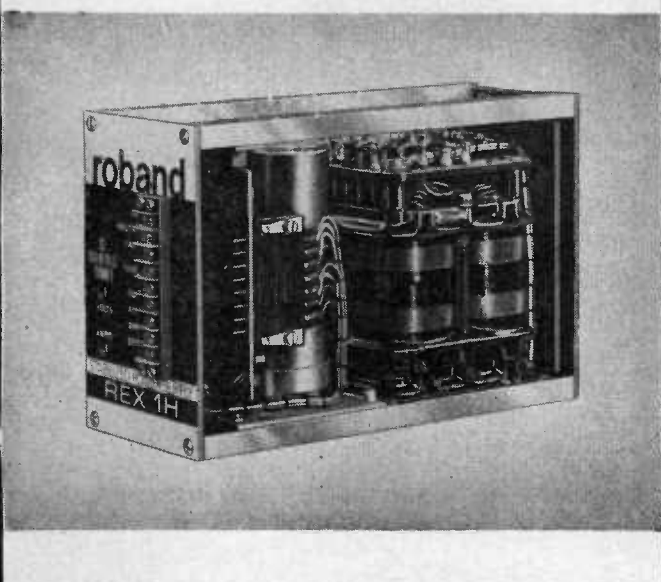
|---|---|---|---|---|---|---|
| REX 2L | 4–12V | 2A | 3⅛" | | 7⅜" | £36. 10. |
| REX 5L | 4–12V | 5A | 4⅛" | | 11 13/16" | £48. 15. |
| REX 10L | 4–12V | 10A | 4⅛" | | 16 13/16" | £66. 10. |
| REX 15L | 4–12V | 15A | 8¼" | | 14⅜" | £84. –. |
| REX 1H | 6–30V | 1.25A | 3⅛" | | 7⅜" | £34. 10. |
| REX 2.5H | 6–30V | 2.5A | 4⅛" | | 9¼" | £44. 5. |
| REX 5H | 6–30V | 5A | 4⅛" | 5" | 14⅜" | £57. 5. |
| REX 10H | 6–30V | 10A | 8¼" | | 14⅜" | £84. – |
| REX 1S | 30–50V | 1A | 3⅛" | | 7⅜" | £37. 15. |
| REX 2.5S | 30–50V | 2.5A | 4⅛" | | 12⅜" | £55. 15. |
| REX 5S | 30–50V | 5A | 8¼" | | 12⅜" | £77. 5. |

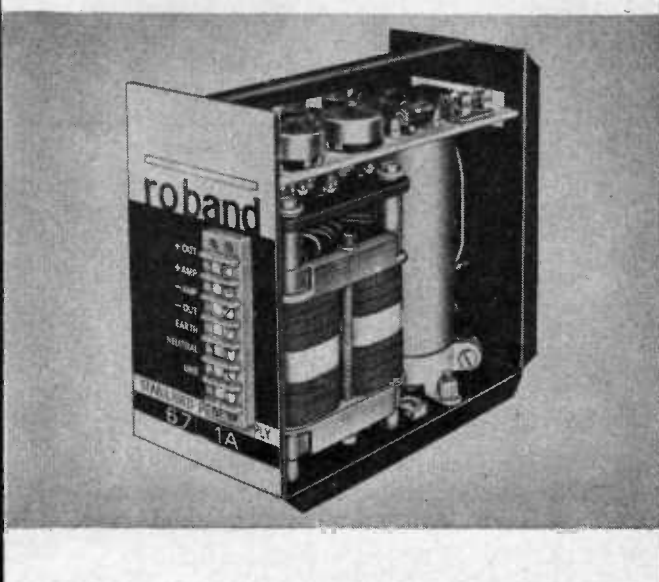
**PERFORMANCE** Input Voltage: 100-125/200-250 ± 7½%. 48-400 cps. Stability (15% mains variation): 0.01%. Stability (0 to full load): 0.05%. Ambient Temperature: —10°C to + 60°C. Temperature Coefficient: 0.01%°C. Protection: Self-resetting current limiter preset to operate at 110% of maximum current rating. Ripple and noise: 500 microvolts, peak to peak.




# 67 series
## lowest cost M.T.B.F. >8000 hrs

| TYPE NO. | OUTPUT VOLTS (pre-set) | CURRENT RATING | WIDTH | HEIGHT | DEPTH | PRICES (carriage paid UK) |
|---|---|---|---|---|---|---|
| 67–1A | 6–24V | 1A | 2½" | 5" | 6" | £22 |
| 67–2A | 6–24V | 2A | 3⅜" | 5¾" | 6⅜" | £27 |
| 67–5A | 6–24V | 5A | 4 3/16" | 6⅛" | 9¾" | £40 |
| 67–10A | 6–24V | 10A | 5½" | 6 31/32" | 12" | £55 |
| 67–Twin 1A | 6–24V (X2) | 1A–1A | 4 3/16" | 5¾" | 6⅜" | £41 |
| 67–Twin 5A | 6–24V (X2) | 5A–5A | 5½" | 6 31/32" | 12" | £76 |

**PERFORMANCE** Input Voltage: 100-125/200-250 ± 7½%. 48-400 cps. Stability (15% mains variation): 0.1%. Stability (0 to full load): 0.2%. Ambient Temperature: —10°C to +60°C. Temperature Coefficient: <0.05%/°C. Protection: Self-resetting current limiter preset to operate at 110% of maximum current rating. Ripple and noise: 2mV RMS.



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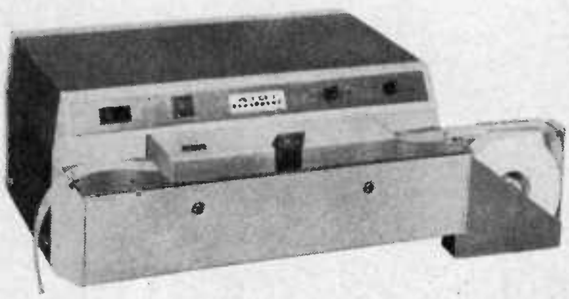
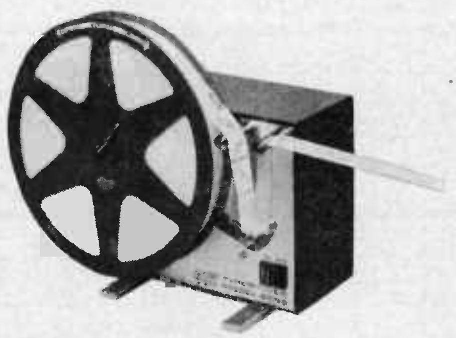
# GNT



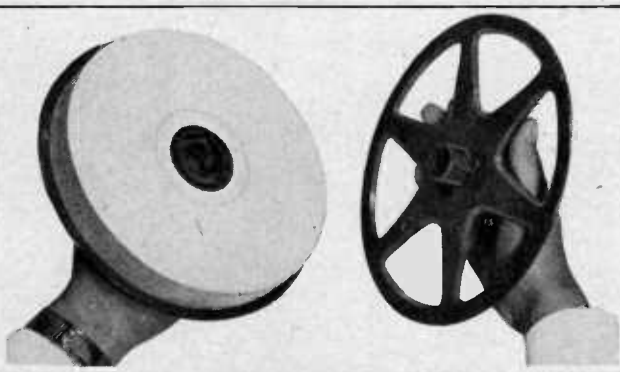
MODEL 4101

**PHOTO-ELECTRIC TAPE READER 0-1000 CH/SEC.**



MODEL 2082

**TAPE WINDER 0-125 CH/SEC.**



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**8½" O/D. 11/16", 7/8" & 1" TAPE**

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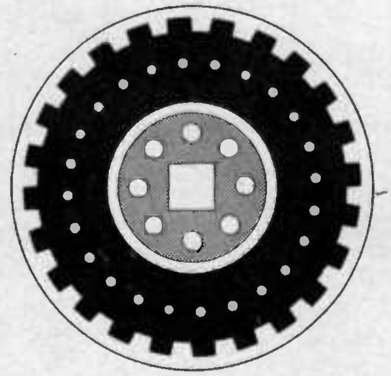

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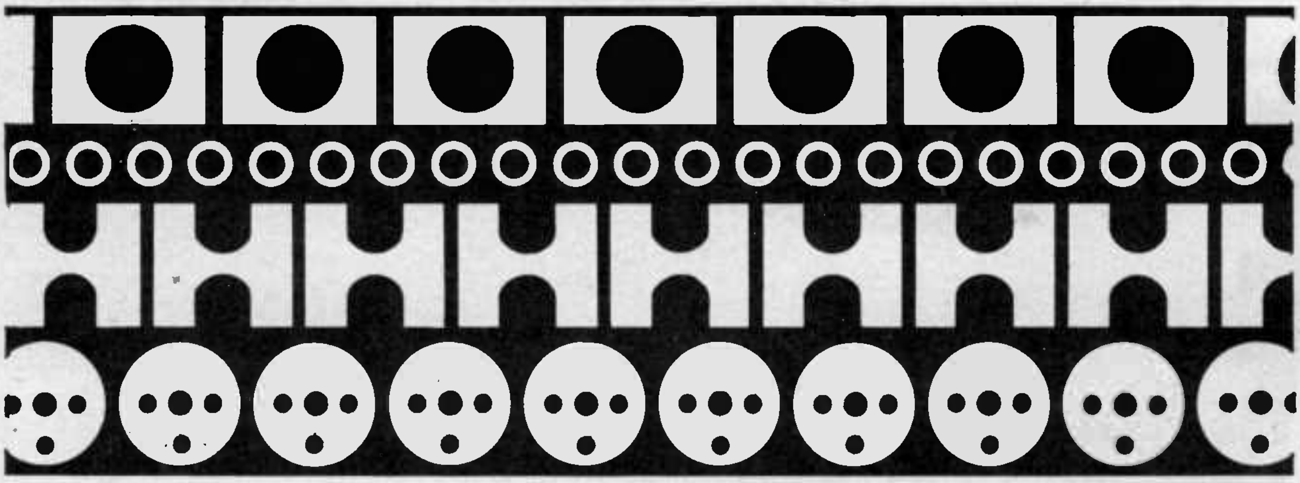



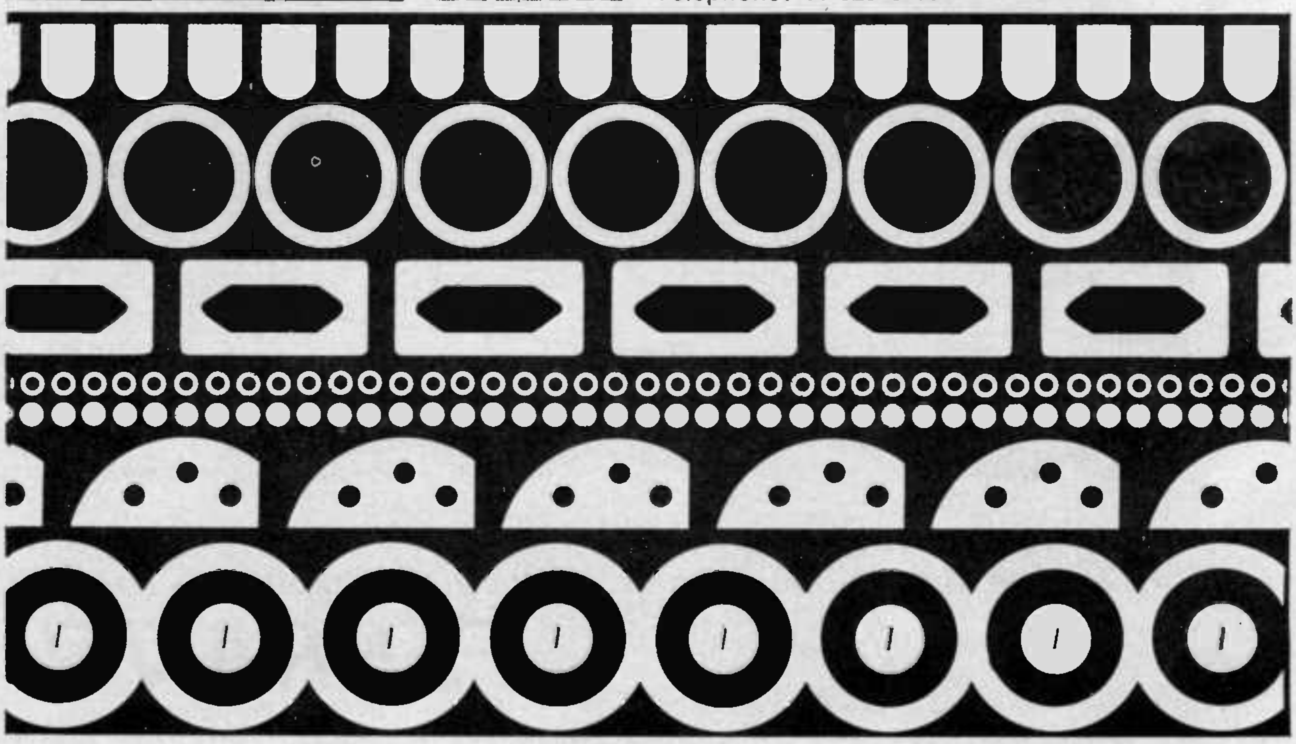

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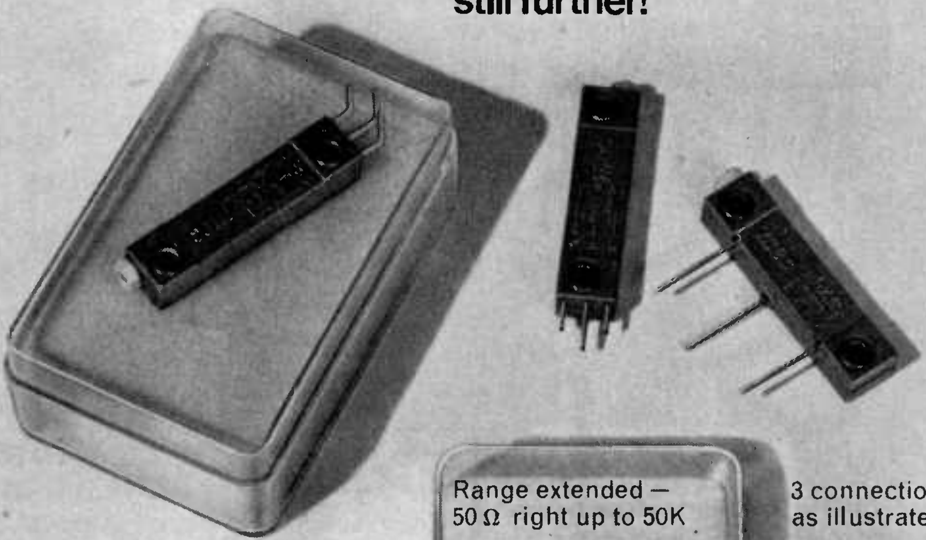
**Improve the performance still further!**



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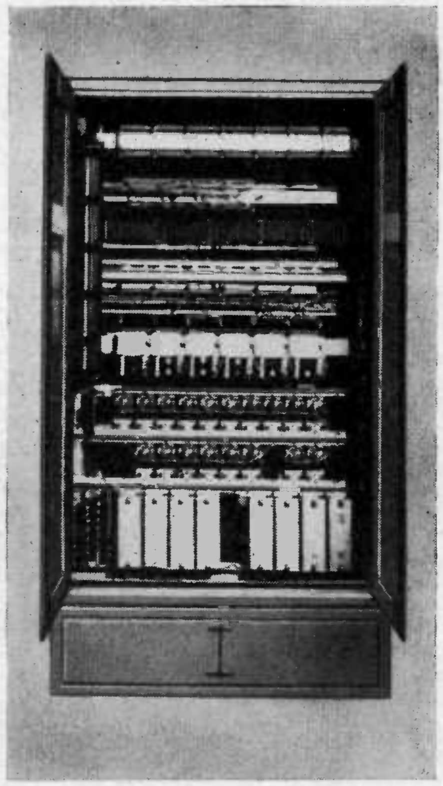

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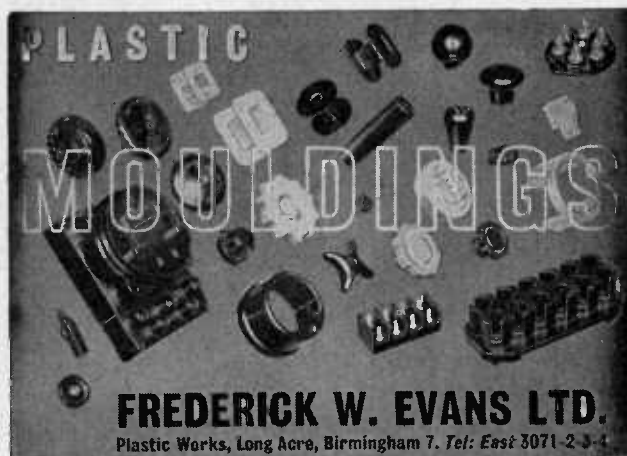


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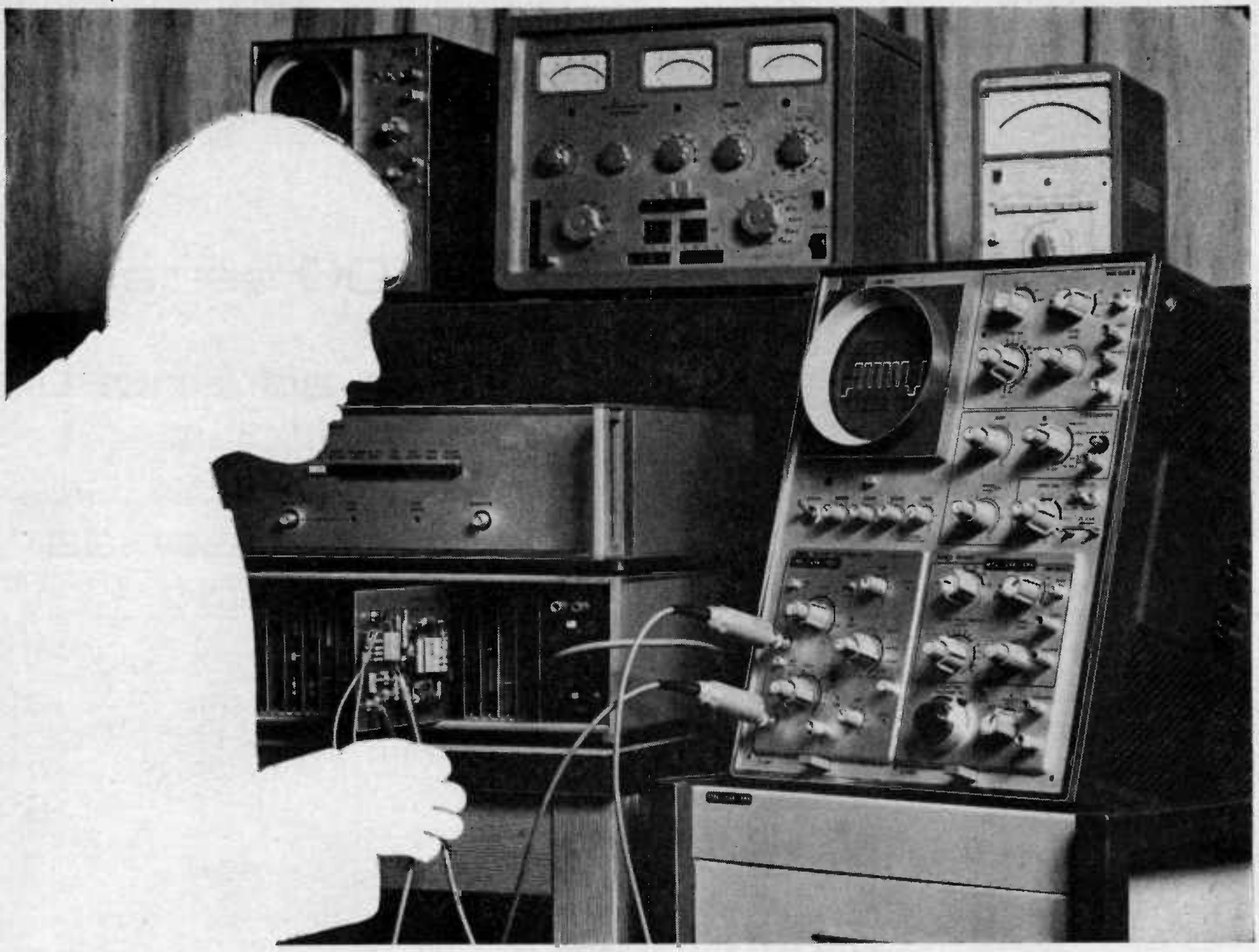

# Move in and plug in

## to 500 μV/cm at 50 MHz single trace or 10 mV/cm at 50 MHz four trace

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As first production models are subjected to life tests under normal user conditions, new circuits to prolonged testing, and none of the components are run at more than 70% of their rating, a high level of reliability is assured.

A full description of this oscilloscope is set out in an 8-page brochure. We will be glad to send you a copy.

**Illustration: The PM 3330 used in conjunction with the PM 5530 TV pulse generator and the PM 5540 TV pattern generator.**

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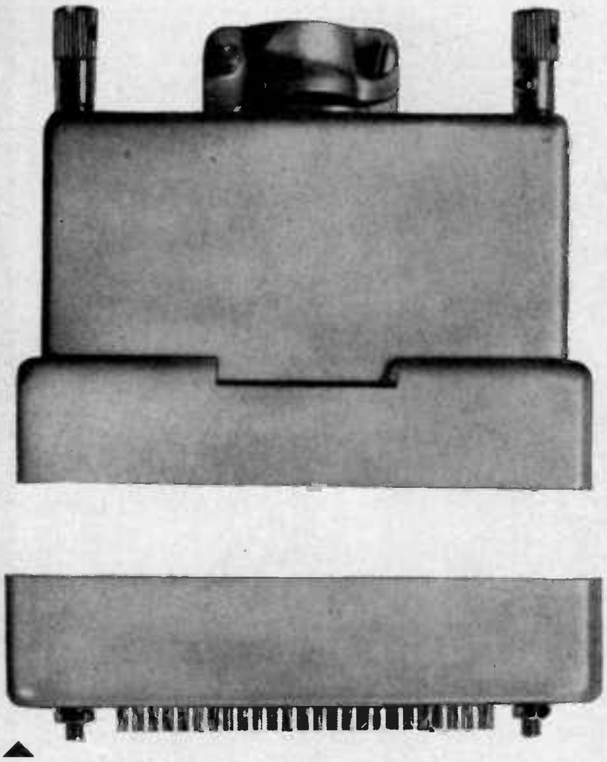

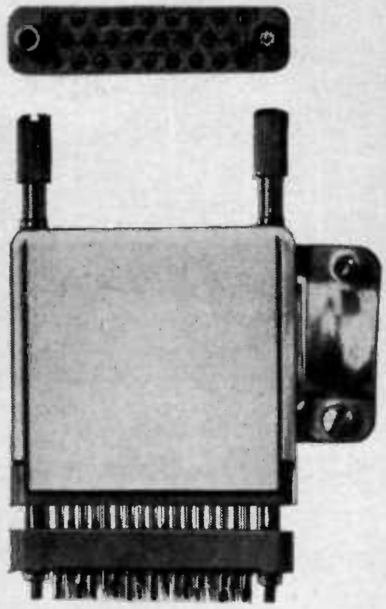

### MICRO-MINIATURE RANGE

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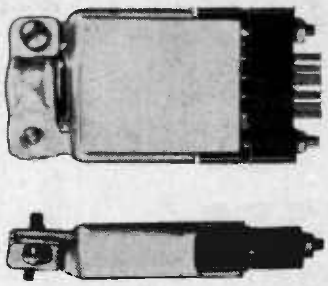

### SUB-MINIATURE RANGE

- ☐ Diallyl Phthalate mouldings
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- ☐ 5, 7, 11, 14, 20, 26, 29, 34, 42, 50, 75 and 104 contacts

### MINIATURE RANGE

- ☐ High dielectric and mechanical strength with Diallyl Phthalate mouldings
- ☐ Screwlocks give fast, easy connections and safeguard against accidental disconnection
- ☐ Plating Gold flash or Heavy gold Military Standard
- ☐ Stainless steel guides and screwlocks—non corrosive
- ☐ Hoods, shells, cable clamps to suit any application
- ☐ 5, 6, 7, 8, 9, 11, 14, 18, 20, 21, 26, 34, 41, 50, 75 and 104 contacts

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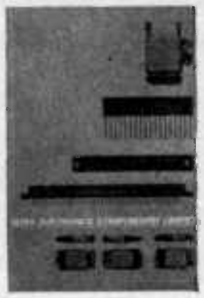

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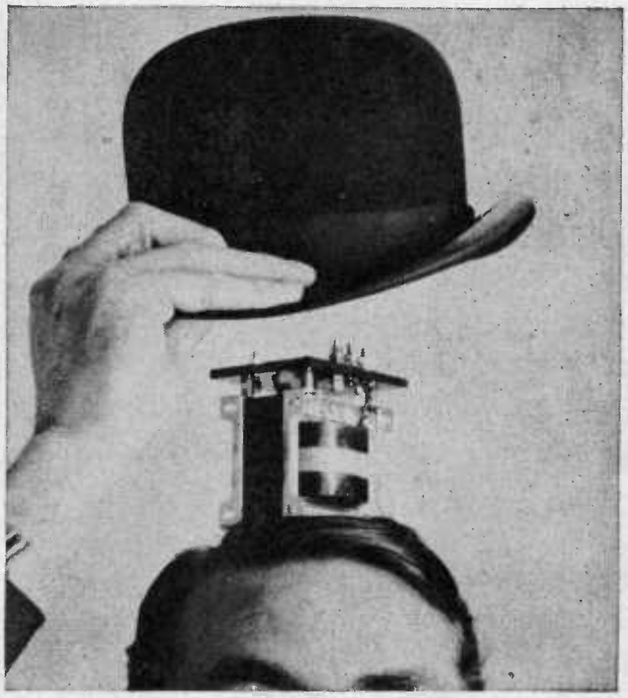
**TECHMASHEXPORT**



## "Don't keep it under your hat" the man said



At ARW we have been making quality transformers and coils for 20 years, mainly for Scottish industry. From time to time we get complaints from new customers in the South that they have only just heard of us, sometimes with the terse advice that we should shout louder about out service □ Well here goes! □ In our modern factory at Thornliebank, we have an experienced design staff and up-to-date facilities for producing transformers from miniature to 20KVA □ Single phase and 3 phase · auto and double wound · open, enclosed, and oil immersed · ring current transformers □ Applications : electrical apparatus · electronic equipment · switchgear · auxiliary supplies □ Vacuum varnish impregnation □ Epoxy resin encapsulation. The price is right and delivery performance is almost a fetish.

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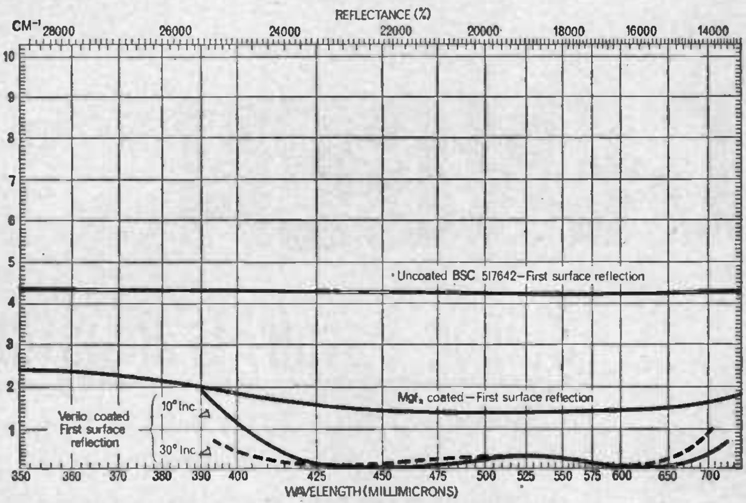
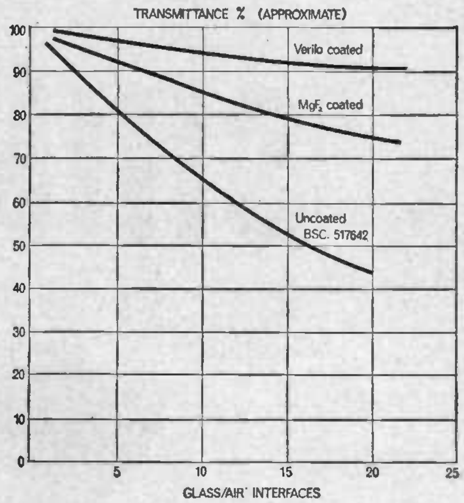
# Lens surface reflectivity cut to 0.5%* by VERILO® coating




**GRAPH A**




**GRAPH B**

Increased performance and the reduction in size of many optical systems is now made possible by the use of Pilkington Perkin-Elmer Verilo (Very Low Reflection) Coatings. As graph A shows reflectivity is very much reduced when compared with conventional magnesium fluoride coatings, and the high efficiency of Verilo coating becomes more beneficial as the number of glass/air interfaces in optical systems increase (see graph B).

For example in a lens system with ten glass/air surfaces Verilo coating in place of magnesium fluoride increases transmittance by almost 10%.

**Specification:** Pilkington Perkin-Elmer Verilo coating is a durable, multilayer dielectric thin-film system for application on substrate systems having an index of refraction between 1.46 and 1.75. These coatings meet the following specular reflectivity limits from each coated surface:

| Spectral Region | Reflectivity 0-15° | Reflectivity at 30° |
|---|---|---|
| 425 to 650 millimicrons | 0.5% Absolute | 1.0% Absolute |
| 425 to 700 ,, | 0.5% Average | 1.0% Average |

**Applications:** As well as transmission enhancement, Verilo coatings can be used with advantage to reduce ghost images in photographic systems; reduce glare on lighting wedges and instrument windows; improve signal-to-noise ratios in electro-optical systems; reduce standing noise in coherent systems; improve high-frequency characteristics of optical transfer functions.

**Availability:** The Verilo coating process is carried out in the modern Pilkington Perkin-Elmer plant at St Asaph, and is available for customers' own optical systems as well as Pilkington Perkin-Elmer products.

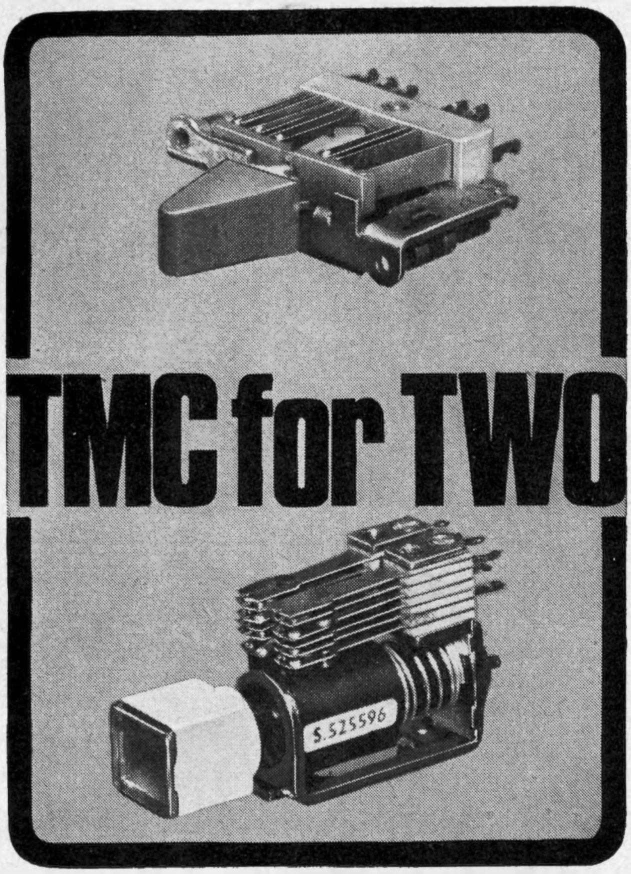
**Other Vacuum Coated Products include laser mirrors, interference filters, and reflective layers.**

**Information:** Please call Mr Elwyn Williams, Pilkington Perkin-Elmer Ltd., St Asaph 3301 for further information on Verilo coatings, other vacuum deposited layers or specialised products.

(*average lens surface reflectivity in range 425 – 700 millimicrons, angle of incidence 0 – 150°).

PILKINGTON
PERKIN-ELMER

**ST ASAPH, FLINTSHIRE**
Tel: St Asaph 3301; Telex 61291: Telegrams Optical St Asaph



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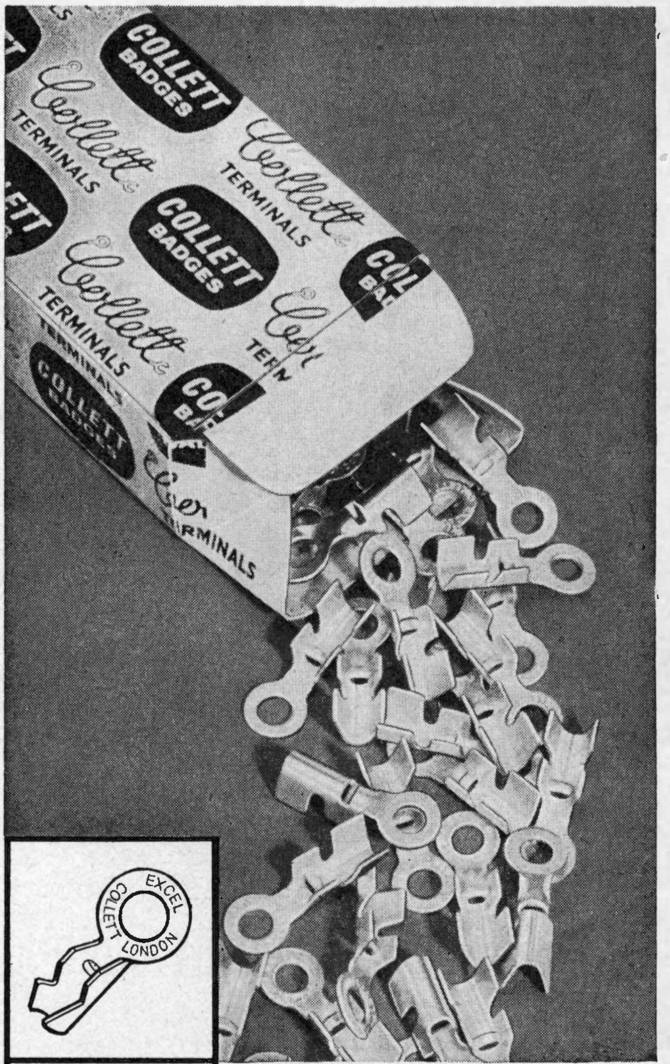



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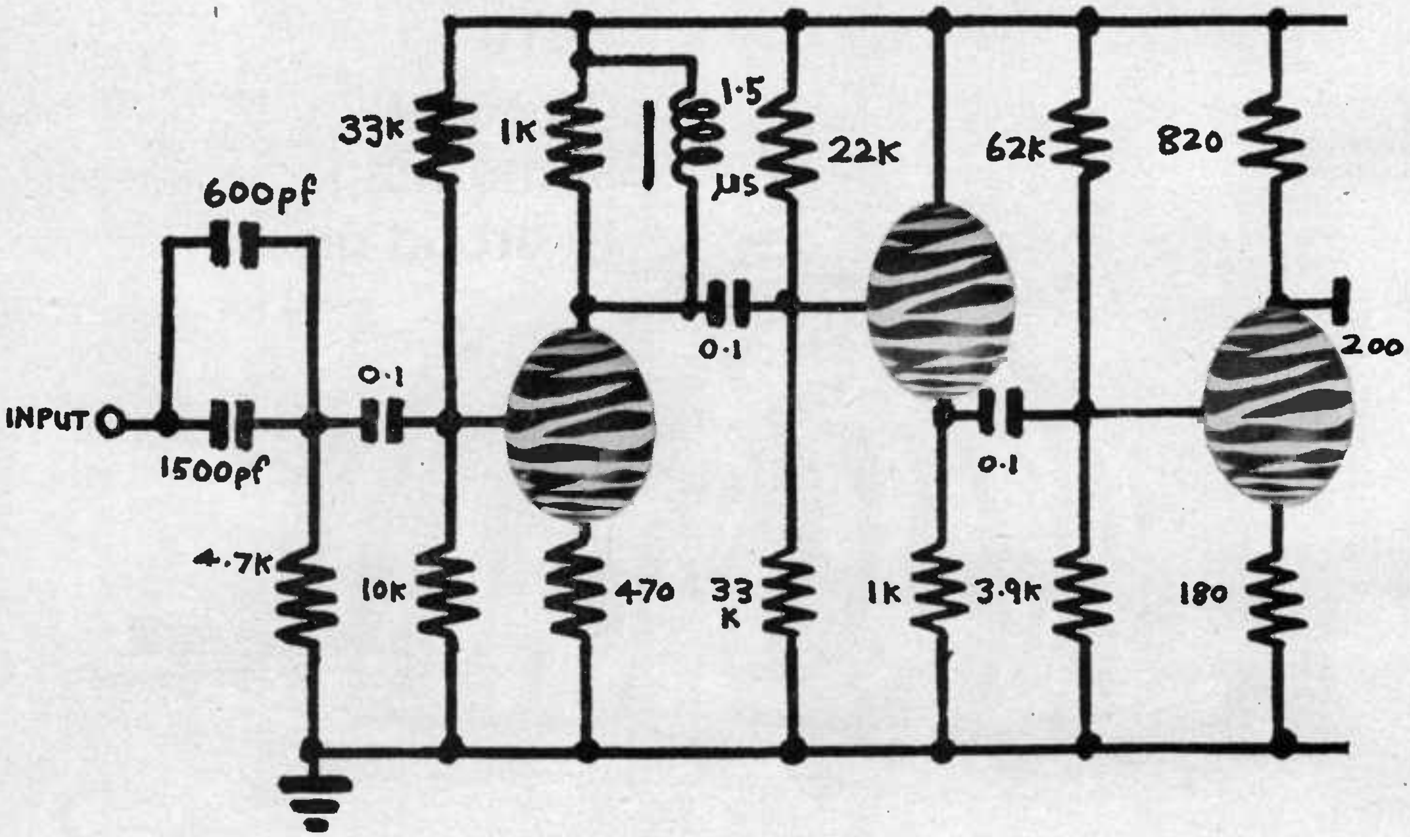


*For more information circle* **No. 89**

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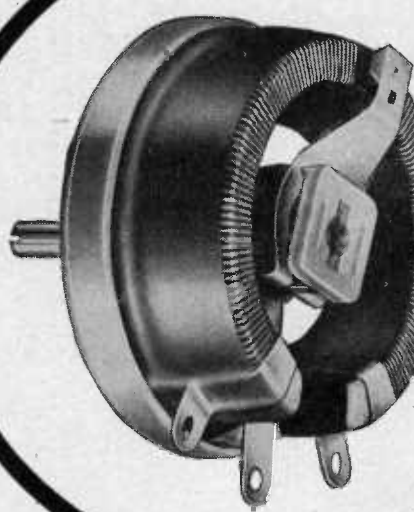

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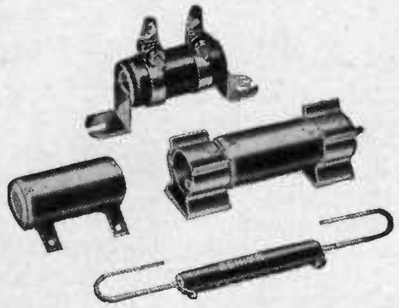

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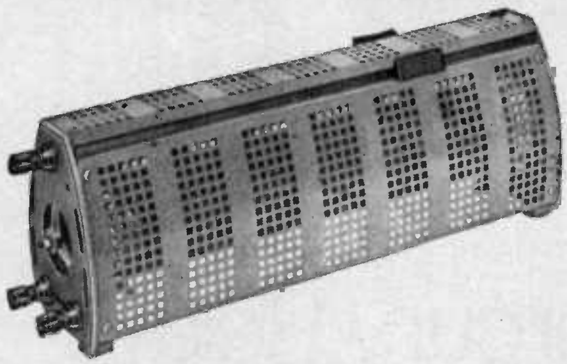

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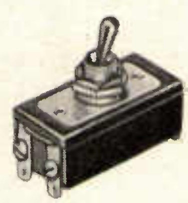

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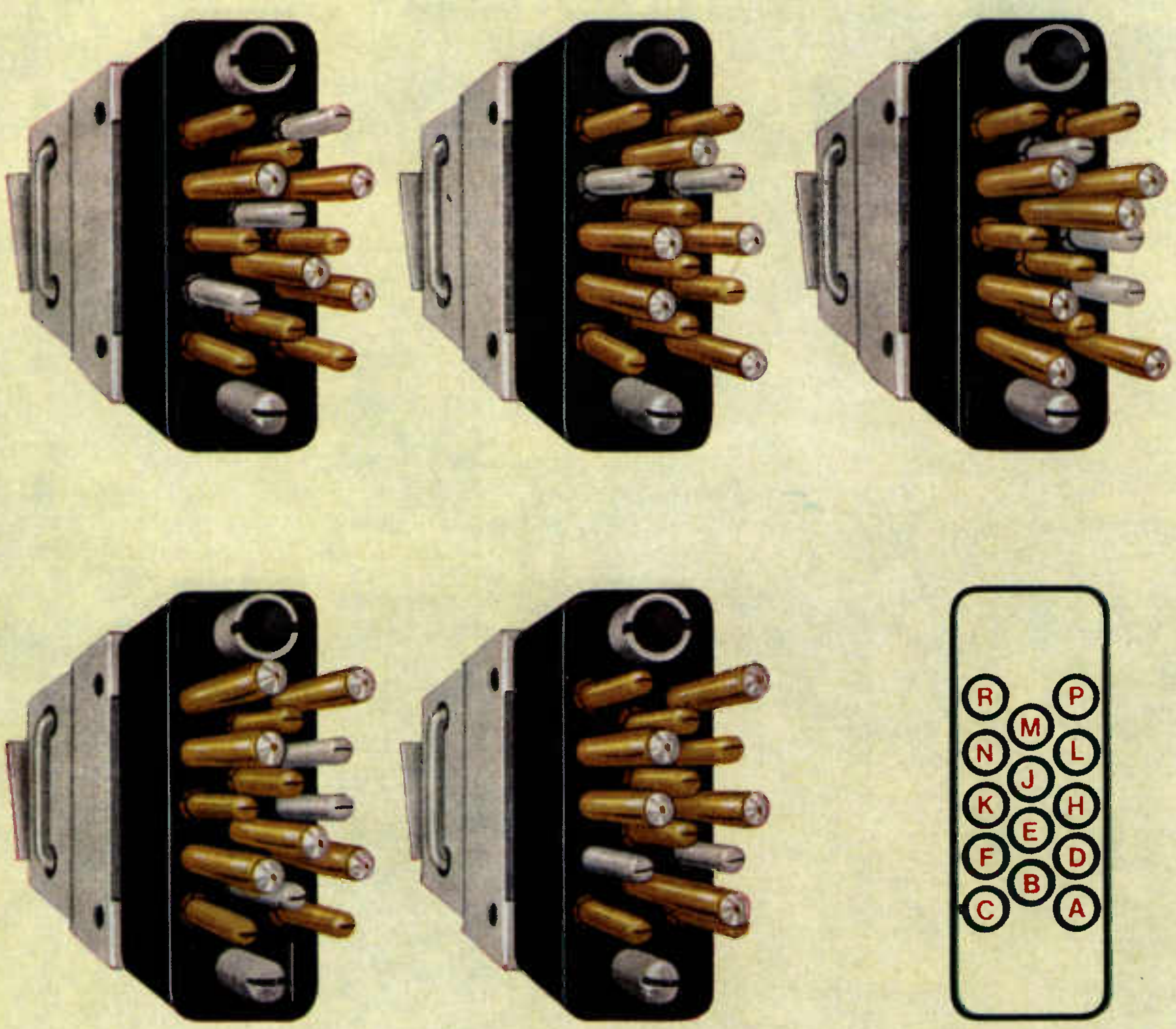


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# Electronic Engineering

VOL. 39. No. 472.


## COMMENTARY

'MICROELECTRONIC Technology Points the Way Ahead' was the title of an article published in this journal about a year ago and the truth of that statement is becoming apparent with increasing rapidity. It is also becoming clear that the various forms of microcircuit, that is thin films, thick films and silicon integrated circuits (s.i.c.) are not so much rivals in the race for supremacy, but rather are they complementary.

The thin and thick film processes both produce passive components only and so require the addition of active devices, which may be included in a common package or connected externally by the user. In each case, conductive, resistive and dielectric layers are applied to an insulating substrate to form one or more circuits. This highlights the basic concept of microelectrics as involving a sequence of operations each of which is applied in parallel to all appropriate parts of the circuit, until the circuit is complete, as opposed to building up a circuit, component by component, using either printed circuits or point-to-point wiring. The thin film process uses evaporation or sputtering and is capable of producing highly stable components which may be adjusted to give precise values. The thick film process uses cermets and glazes applied by silk screening or similar techniques to produce components of moderate accuracy.

As pointed out in a recent article* the basic advantage of s.i.c. technology is that it produces active devices—diodes and transistors—which dominate the performance and cost of electronic equipment, particularly in the digital field. The passive components produced by this process are less accurate and more restricted in range of values than in the film processes, but complete circuit functions can be performed by s.i.c.'s. The major investment in microelectronics by many large organizations throughout the world (as, for instance, the Plessey Company) is being made in s.i.c. technology not primarily because of the reduction in size or improvement in reliability offered—valuable though these features may be—but because it offers a way of reducing the cost of electronic equipment.

One of the basic reasons for the economic advantage of the s.i.c. process may be seen by comparing it with silicon planar epitaxial transistors. A typical npn transistor of this type for small signal linear applications up to a few hundred megahertz or switching applications with times of around 10nsec, has an active area of about ·002in × ·002in. The area of silicon used is normally increased to around 0·02in × 0·02in to provide room to attach lead wires and to ease the other assembly problems. Thus, only 1 per cent of each silicon 'chip' is really being used for the semiconductor device; the remaining 99 per cent being sacrificed for mechanical convenience. The process appears even more inefficient when transistors, processed in the form of slices of silicon 1 to 1·5in in diameter, each containing some 1000 transistors, are divided and separately encapsulated only to be assembled together again by the user.

The use of integrated assemblies of transistors does, in general, require a more complex technology as the orthodox planar transistor process automatically gives a common collector connexion which has limited use. Extra stages in processing are required to produce transistors with electrically isolated collectors, though resistors, capacitors and diodes may be formed by the basic transistor process, by including them on the appropriate photo-engraving mask. The net result of this more complex processing is to increase the cost of producing a finished silicon slice by rather less than a factor of two while improving the utilization from 1 per cent to between 10 per cent and 50 per cent depending on chip size and the area required for isolation diffusion.

The question of yield is, of course, very important and greater process complexity and number of components per chip must reduce the yield of s.i.c.'s compared with single transistors produced with the same level of technology. Firstly, each additional stage increases the number of whole slices rejected. Secondly, the increased active area per chip will increase the chance of a fault occurring in that area. Once a reasonable level of process control has been established, it is the second mechanism which largely controls the yield. The larger the chip, the greater is the chance of a fault occurring in it but on the other hand, the greater the complexity of the s.i.c., the greater is the economic advantage over discrete components, and from this it can be seen that costwise there is an optimum size of chip. Naturally, as further technological advances are made, higher and higher yields can be expected and s.i.c.'s of greater complexity will become feasible and economically viable.

There is no doubt that microelectronics in one form or another is going to bring about profound changes in the electronic industry and that s.i.c.'s are going to play an important part in this. Indeed, it can be seldom in the history of technology that one element has emerged so dramatically from the background to play such a unique and dominant role in re-shaping the course and future of an industry and, moreover, with the exception of oxygen it is the most abundant element of the earth's crust!

* HOLT, W. Microelectronics Reliability and Economics. *Component Technology* (The Plessey Co. Ltd.) 2, 6 (May 1967).

# Calculation of V-I Curves for N.T.C. Thermistors

## A Method of Calculating the Steady-State Voltage/Current Characteristics of Directly-Heated Negative-Temperature-Coefficient Thermistors

**(Part 1)**

By M. R. McCann*, C.Eng., A.M.I.E.R.E.

*When the current through an n.t.c. thermistor is slowly increased, the corresponding voltage increases to a maximum $E_{max}$ and then starts to fall. This characteristic is due to the self-heating of the thermistor.*

*The basic formulae involving resistance, temperature and power are used to derive a graph which shows curves of normalized voltage and current for negative temperature coefficient thermistors. From these curves and a knowledge of the cold resistance and $E_{max}$, the voltage-current characteristic of the thermistor can be determined in any ambient temperature. An example of this calculation is shown for a thermistor mounted in different enviromental conditions. The analysis includes a simple method of calculating $E_{max}$ (on the V-I characteristic) and its variation with ambient temperature. A universal resistive load line is derived for use in conjunction with voltage-current characteristics plotted with logarithmic scales.*

*Several examples of the voltage-current characteristic of different types of thermistors are shown measured under practical conditions and they are compared with characteristics calculated from the normalized curves.*

*The analysis is extended to include thermistors mounted in a vacuum, where there is a considerable deviation in the behaviour of the thermistors from that of those mounted in other media.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 408)

IN applications of the negative temperature coefficient thermistor where a significant amount of power is dissipated in the device, the basic analysis for an electrical design requires a knowledge of the steady state operating point under conditions of thermal equilibrium. Such applications include temperature compensation in electrical circuits, measurement of electrical power, amplitude control of oscillators and amplifiers, manometers, katharometers, anemometers and other flow measurements. This operating point can be described electrically if any two of the following quantities are known:

(a) Voltage across the thermistor
(b) Current through the thermistor
(c) Its resistance
(d) The amount of power being dissipated in the device

The voltage and current are perhaps the most convenient quantities since when plotted in a graphical form, a resistive load line may easily be drawn to obtain the operating point.

In the following analysis it will be shown how the voltage-current characteristic under any conditions can be calculated simply without resorting to complex exponential equations, assuming that the thermal conditions of the surrounding medium and the 'characteristic' parameters of the thermistor are known. The analysis also shows how the voltage-current characteristic is dependent upon the variation of the ambient temperature, the conductivity of the surrounding medium, the resistance of the thermistors, etc., and other variables.

### Calculation of $E_{max}$

When electrical power is dissipated in a thermistor, its mean temperature rise above that of its surroundings is in direct proportion to the magnitude of the dissipation. This relationship holds when the environmental conditions of the thermistor are such that the heat flow to its surroundings obeys Newton's law of cooling. This law is closely followed when the thermistor is in free air and the temperature of the thermistor is sufficiently low to neglect the radiation of heat from its surface.

In the majority of the following analysis, it is assumed that the above relationship holds.

It has been found that the following relationship for negative temperature co-efficient thermistors holds over a wide temperature range:

$$R_T = R_\infty \exp (B/T) \quad \dots\dots\dots (1)$$

where $R_T$ is the resistance of the thermistor when its mean temperature is $T$ degrees Kelvin; $R_\infty$ is a constant equal to the thermistor resistance when the temperature $T$ is infinitely high; and $B$ is the constant that will be referred to in the text as the characteristic temperature of the thermistor.

It has been found by experiment that the variation in $B$ for STC thermistors over the temperature range from −40°C to +120°C is less than 1 per cent. This was verified experimentally by accurate measurement of resistance and temperature. The equipment that was used did not permit verification outside this range of temperatures.

In an ambient temperature of $T_a$°K with zero power dissipated in the thermistor, equation (1) becomes:

$$R_o = R_\infty \exp (B/T_a) \quad \dots\dots\dots (2)$$

This article relates only to the steady state condition of the thermistor. The thermistor is allowed to regain equilibrium after any change in the conditions. All voltages and currents are effective (r.m.s.) values.

Consider a power of $W$ watts dissipated in the thermistor such that its mean temperature is raised to $\Delta T$ above that of the surroundings (which are assumed to be at $T_a$°K). Then

$$W = k\Delta T \quad \dots\dots\dots\dots\dots\dots (3)$$

where $k$ is the dissipation constant (i.e. the reciprocal of the thermal resistance). The resistance of the thermistor $R_T$ now becomes:

$$R_T = R_\infty \exp\left(\frac{B}{T_a + (W/k)}\right) \quad \dots\dots\dots\dots (4)$$

from equations (1) and (3). This assumes that the resis-

* *Standard Telephones & Cables Ltd.*

tance of the thermistor obeys equation (1) where $T$ is the mean temperature of the thermistor. From equations (2) and (4).

$$R_T = R_o \exp\left(\frac{B}{T_a + (W/k)} - (B/T_a)\right) \quad \ldots. \ (5)$$

Now:

$$E = IR_T \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (6)$$

where $E$ is the voltage across the thermistor and $I$ is the current through the thermistor.

Equation (5) becomes:

$$E = IR_o \exp\left(\frac{B}{T_a + (W/k)} - (B/T_a)\right) \quad \ldots. \ (7)$$

If equation (7) is differentiated with respect to $I$, it can be determined whether the voltage has maxima or minima.

Rearranging equation (7) and substituting $W = EI$:

**SYMBOLS**

| | Units | |
|---|---|---|
| $B$ | °K | The 'characteristic temperature' of the thermistor |
| $D$ | °K | A parameter which is a function of $T_a$ and $B$ |
| $D'$ | °K | The value of $D$ at ambient temperature $T_a'$ |
| $E$ | V | The d.c. voltage across the thermistor |
| $E_{max}$ | V | The maximum voltage across the thermistor |
| $E_{max}'$ | V | The maximum voltage across the thermistor at an ambient temperature $T_a'$ |
| $F$ | °K | Another parameter which is a function of $T_a$ and $B$ |
| $I$ | A | The direct current through the thermistor |
| $I_m$ | A | A parameter which is a function of $E_{max}$ and $R_o$ |
| $I_{max}$ | A | The current through the thermistor at $E_{max}$ |
| $k$ | W/°C | The dissipation constant of the thermistor |
| $R_{max}$ | Ω | The resistance of the thermistor at $E_{max}$ |
| $R_o$ | Ω | The resistance of the thermistor at zero power dissipation |
| $R_o'$ | Ω | The value of $R_o$ at $T_a'$ |
| $R_T$ | Ω | The resistance of the thermistor at temperature $T$ |
| $R_\infty$ | Ω | The resistance of the thermistor at infinite temperature |
| $T$ | °K | The mean temperature of the thermistor |
| $T_a$ | °K | The ambient temperature of the medium surrounding the thermistor |
| $T_a'$ | °K | An arbitrary ambient temperature different from $T_a$ |
| $T_c$ | °C | The temperature $T_a$ expressed in degrees Celsius |
| $W$ | W | The mean power dissipated in the thermistor |
| $W_{max}$ | W | The mean power dissipated in the thermistor at $E_{max}$ |
| $\Delta T$ | °C | A change in the mean temperature of the thermistor |
| $\Delta T_a$ | °C | A change in the ambient temperature $T_a$ |

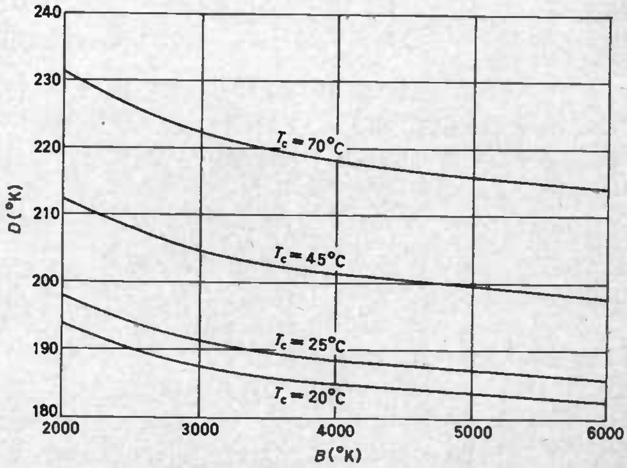



**Fig. 1.** Variation of the parameter $D$ with the characteristic temperature $B$ at various ambient temperatures

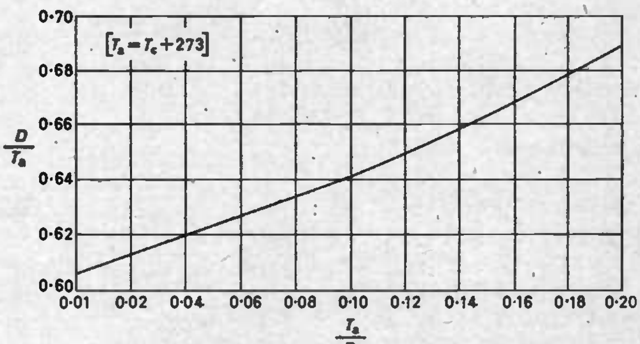



**Fig. 2.** Normalized relationship between the parameter $D$, the characteristic temperature $B$, and the ambient temperature

$$\ln E - \ln I = \ln R_o + \frac{B}{T_a + (EI/k)} - (B/T_a)$$

Therefore:

$$1/E\,(dE/dI) - (1/I) = -B \left[\frac{E/k}{(T_a + (EI/k))^2} + \frac{(dE/dI)(I/k)}{(T_a + (EI/k))^2}\right]$$

Hence:

$$dE/dI = E/I \left[\frac{(EIB/k) - (T_a + (EI/k))^2}{(EIB/k) + (T_a + (EI/k))^2}\right]$$

Now $dE/dI = 0$, corresponding to a voltage maximum or minimum for a slow change of $I$, when:

$$(T_a + (EI/k))^2 = EIB/k$$

hence:

$$(EI)^2 + EI\,k(2T_a - B) + k^2T_a^2 = 0 \quad \ldots\ldots \ (8)$$

and:

$$EI = K\,[(B/2)\{1 \pm \sqrt{(1 - (4T_a/B))}\} - T_a] \quad \ldots. \ (9)$$

The roots of equation (8) are given in equation (9). These are real, since in practice $4T_a < B$. (If $4T_a > B$ there will be no voltage maximum or minimum.) The negative sign of the square root in equation (9) corresponds to a voltage maximum. The positive sign is of no interest and refers to a voltage minimum occurring at a temperature well above the maximum permissible for the thermistor.

Under the conditions when the voltage is a maximum, let:

$$E = E_{max}$$
$$EI = W_{max}$$
$$T = T_{max}$$
$$E/I = R_{max}$$

Then:

$$\Delta T = T_{max} - T_a \quad \ldots\ldots\ldots\ldots\ldots \quad (10)$$

and:

$$\Delta T = W_{max}/k \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (11)$$

Then from equation (9):

$$W_{max} = k\,[(B/2)\{1 - \surd(1 - (4T_a/B))\} - T_a] \quad \ldots \quad (12)$$

and:

$$T_{max} = B/2\,[1 - \surd(1 - (4T_a/B))] \quad \ldots\ldots (13)$$

Now:

$$E_{max}{}^2 = W_{max} R_{max} \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

and:

$$R_{max} = R_o \exp((B/T_{max}) - (B/T_a)) \quad \ldots\ldots \quad (15)$$

From equations (12), (13) and (15):

$$E_{max} = [R_o k(T_{max} - T_a)]^{\frac{1}{2}} \exp((B/2T_{max}) - (B/2T_a)) \quad \ldots \quad (16)$$

This equation is a complex function of the variables $B$, $T_a$, $R_o$, $k$ and $E_{max}$ ($T_{max}$ is a function of $B$ and $T_a$). However, the following equation (evolved in Appendix A) shows that a very much simpler relationship exists between these quantities, i.e.:

$$E_{max} = D\,\surd(R_o k/B) \quad \ldots\ldots\ldots \quad (17)$$

where:

$$D \simeq (T_a/\surd e)\,[1 + (T_a/2B) + (5/8)\,(T_a{}^2/B^2) + (53/48)\,(T_a{}^3/B^3) + (903/384)\,(T_a{}^4/B^4) + (21769/3840)\,(T_a{}^5/B^5)] \quad \ldots\ldots\ldots \quad (18)$$

$D$ is a function of $T_a$ and $B$, and does not vary appreciably over a wide range of values of $B$ (2 500°K to 6 000°K). It has the dimensions of absolute temperature.

When the ambient temperature $T_o$ is 20°C, ($T_a$ = 293°K), $D \simeq 186$ and

$E_{max} = 186\,\surd(R_o k/B)$ within ±2 per cent over a range of $B$ from 2 500°K to 6 000°K .......... (19)

When $T_c$ is 25°C ($T_a$ = 298°K)

$E_{max} = 190\,\surd(R_o k/B)$ within ±2 per cent over a range of $B$ from 2 500°K to 6 000°K .......... (20)

It also follows that for most practical purposes:

$$E_{max} = 0{\cdot}64\,T_a\,\surd(R_o k/B) \quad \ldots\ldots\ldots\ldots \quad (21)$$

This can be seen in Fig. 1 where $D$ is plotted against $B$ for various ambient temperatures.

Fig. 2 is a plot of $D/T_a$ versus $T_a/B$, which enables $D$ to be determined easily for any given values of $T_a$ and $B$.

Therefore, $E_{max}$ can be calculated from Fig. 1 or 2 and equation (17) for given values of ambient temperature, characteristic temperature ($B$), dissipation constant and cold resistance ($R_o$) or, if the ambient temperature is 20°C or 25°C, by the approximate formulae in equations (19) and (20).

### The Generalized Voltage-Current Characteristic

Equation (17) enables the generalized voltage-current characterisic to be determined as follows:

Now $E^2 = WR_T$

$= k\Delta TR_T$ from equation (3)

$= k\Delta TR_o\,(R_T/R_o)$

Hence $E = (E_{max}/D)\,\surd(B\Delta TR_T/R_o)$ by substitution for $k$ from equation (17).

Therefore:

$$E/E_{max} = \frac{\surd(B\Delta T)}{D} \exp\left[(B/2)\left(\frac{1}{T_a + \Delta T} - (1/T_a)\right)\right] \quad \ldots\ldots\ldots\ldots \quad (22)$$

Also $I = \surd(W/R_T)$

$$= \sqrt{\left(\frac{k\Delta TR_o}{R_o R_T}\right)}$$

Substitution for $k$ gives:

$$I = (E_{max}/DR_o)\sqrt{\left(\frac{B\Delta TR_o}{R_T}\right)}$$

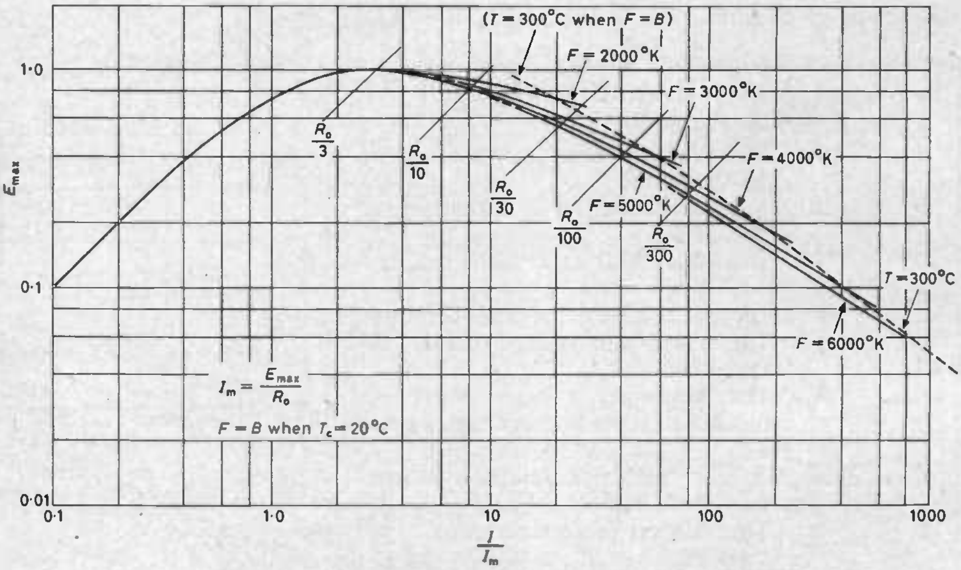



**Fig. 3. Relations between the normalized voltage and the normalized current as a function of the parameter $F$**

Hence:

$$IR_o/E_{max} = \frac{\surd(B\Delta T)}{D} \exp\left[-(B/2)\left(\frac{1}{T_a + \Delta T} - (1/T_a)\right)\right] \quad \ldots\ldots\ldots\ldots\ldots \quad (23)$$

Let:

$$I_m = E_{max}/R_o \quad \ldots\ldots\ldots\ldots\ldots \quad ((24)$$

Then:

$$I/I_m = \frac{\surd(B\Delta T)}{D} \exp\left[-(B/2)\left(\frac{1}{T_a + \Delta T} - (1/T_a)\right)\right] \quad \ldots\ldots\ldots\ldots \quad (25)$$

$E/E_{max}$ and $I/I_m$ can be calculated for given values of $B$, $T_a$ and $\Delta T$ from equations (22) and (25). They have been plotted graphically in Fig. 3 for an ambient temperature of 20°C. It is interesting to note that the current ($I_{max}$) at maximum voltage is approximately equal to $3I_m$. Also contours can be plotted in Fig. 3 showing lines of equal mean thermistor temperature. Note, however, that these lines apply only for an ambient temperature of 20°C.

Once the relationship between the normalized voltage $E/E_{max}$ and the normalized current $I/I_m$ is known, the relationship between voltage $V$ and current $I$ can be determined for any given or calculated values of $E_{max}$ and $R_o$. (In some calculations more accuracy is obtained if

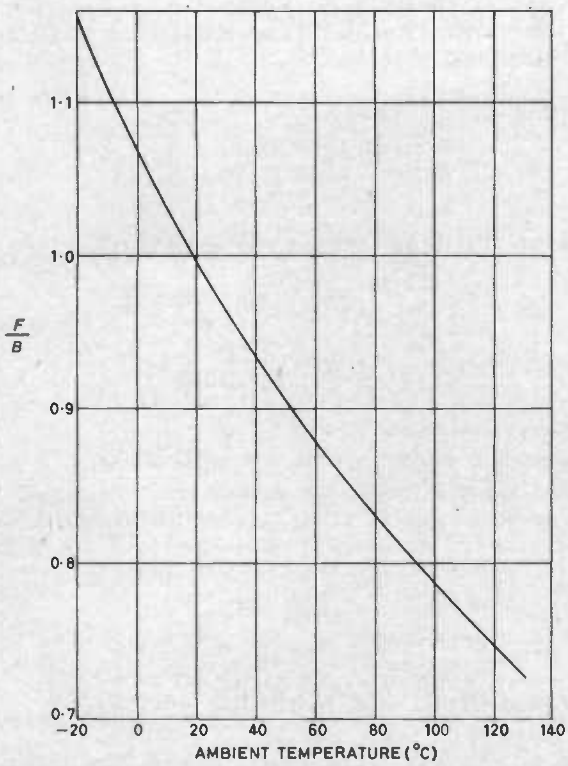



**Fig. 4. Relation between the parameter $F$ and the characteristic temperature $B$ over the range −20°C to +130°C in ambient temperature**

Fig. 3 is plotted in a different form, as described in Appendix C.)

At first sight, it is necessary to plot equations (22) and (25) again for each different ambient temperature. However this is unnecessary, as shown below.

### The Normalized Voltage-Current Curves for Different Ambient Temperatures, $T_a$

Consider equation (7):

$$E = IR_o \exp\left(\frac{B}{T_a + (W/k)} - B/T_a\right)$$

then:

$$E/E_{max} = (IR_o/E_{max}) \exp\left[(B/T_a)\left(\frac{1}{1 + (E/E_{max})(IR_o/E_{max})(E_{max}^2/kT_aR_o)} - 1\right)\right]$$

and:

$$E/E_{max} = (I/I_m) \exp\left[(B/T_a)\left(\frac{1}{1 + (E/E_{max})(I/I_m)(E_{max}^2/kT_aR_o)} - 1\right)\right]$$

But:

$$E_{max}^2 = D^2(kR_o/B)$$

Hence:

$$E/E_{max} = (I/I_m) \exp\left[(B/T_a)\left(\frac{1}{1 + (E/E_{max})(I/I_m)(D^2/BT_a)} - 1\right)\right] \quad \ldots\ldots\ldots\ldots (26)$$

Now from equation (18) the expression $D^2/BT_a$, which occurs in equation (26) can be written:

$$D^2/BT_a = (T_a/eB)\left[1 + 1/2\,(T_a/B) + 5/8\,(T_a^2/B) + \ldots\ldots\right]^2$$

The value of $D^2/BT_a$ is therefore a function of $T_a/B$ only, and is independent of the actual value of either $T_a$ or $B$. The only other place where $T_a$ and $B$ enter equation (26) is again as the ratio $T_a/B$. Since there are no other quantities involved in (26) except the normalized current and voltage, it follows that $T_a/B$ is the only parameter involved in the relation between $E/E_{max}$ and $I/I_m$.

Therefore, the effect of any change in $T_a$ may be exactly replaced by the effect of a reciprocal change in $B$. The curves in Fig. 3 have therefore been plotted in terms of a parameter $F$ where:

$$F = B(293/T_a) \quad \ldots\ldots\ldots\ldots\ldots (27)$$

$$= B\,\frac{293}{T_C + 273} \quad \ldots\ldots\ldots\ldots\ldots\ldots (28)$$

where $T_C$ is the ambient temperature $T_a$ expressed in degrees Celsius (so that for $T_C = 20°C$, $F = B$).

Fig. 4 shows the variation of $F$ in terms of the ambient temperature $T_C$.

When determining the voltage-current characteristics of a thermistor from the normalized curves it might be noted that $E_{max}$ and $I_m$ are also functions of temperature. Both these quantities can be calculated from equations (2) and

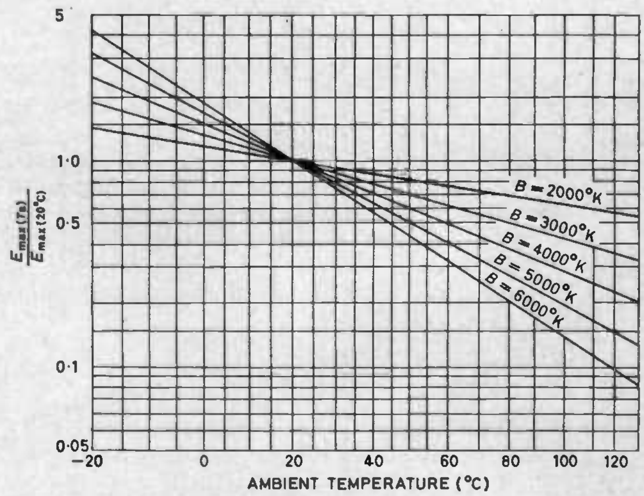
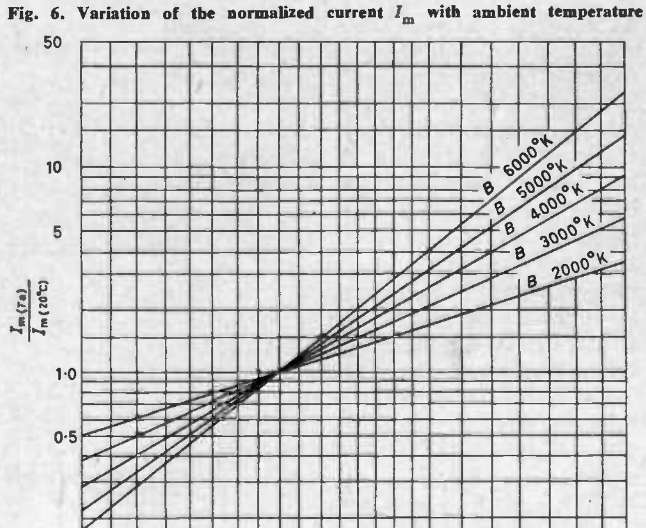



**Fig. 5. Variation of the normalized maximum voltage $E_{max}$ with ambient temperature**

**Fig. 6. Variation of the normalized current $I_m$ with ambient temperature**



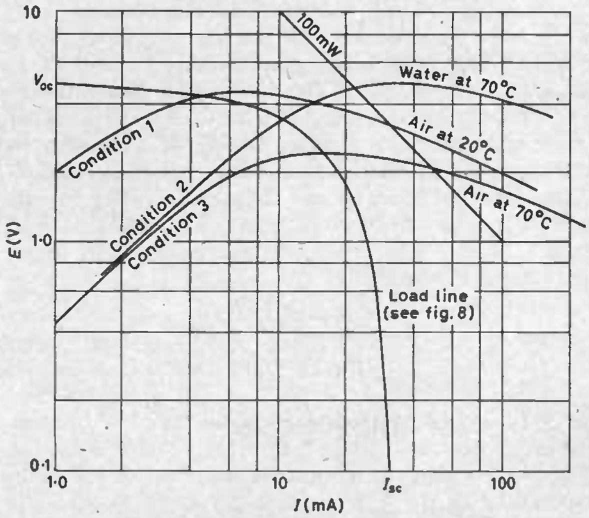



**Fig. 7. Examples of the voltage-current characteristic of an F23D thermistor under various operating conditions**

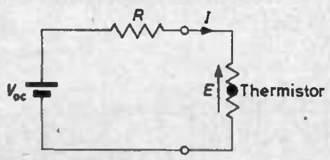



**Fig. 8. General equivalent circuit for a load line (example shown in Fig. 7)**

*$V_{oc}$ = open-circuit voltage,*
*$I_{sc}$ = short-circuit current,*
*R = load line resistor where*
*$V_{oc}/I_{sc} = R$*

(21). However if $E_{max}$ and $I_m$ are known at some particular ambient temperature (20°C or 25°C), $E_{max}$ and $I_m$ at some other ambient temperature can be determined if it is known how the ratio of these quantities varies with ambient temperature.

For $E_{max}$ this has been solved (in Appendix B) using the approximation:

$$\ln\left(\frac{E_{max}''}{E_{max}'}\right) = -\frac{\Delta T_a}{T_a'' T_a'}[(B/2) - 310] \quad \ldots\ldots \ (29)$$

where $E_{max}'$ corresponds to an ambient temperature $T_a'$ and $E_{max}''$ corresponds to an ambient temperature $T_a''$

If $T_a'$ is 293°K (20°C) the apparent error in the characteristic temperature over a range of ambient temperature from −20°C to +120°C would be in the order of ±2 per cent. Since the tolerance on the characteristic temperature is of the order of ±5 per cent, it is a reasonable approximation to say that for most practical purposes equation (29) is correct over this temperature range.

Comparison of equation (29) with a re-arrangement of equation (2) shows that the variation of $E_{max}$ from its stated value of 20°C obeys the same law as the variation of the thermistor resistance with temperature (with no power dissipated), except that the apparent value of the characteristic temperature is $((B/2) - 310)$°K. It can be shown by a similar argument to that in Appendix B that the variation of $I_m(=E_{max}/R_o)$ with temperature is determined by an apparent characteristic temperature of $((B/2+310)$°K. Figs. 5 and 6 show the variation of $E_{max}''/E_{max}'$ and $I_m''/I_m'$ with ambient tem-temperature.

Thus with the aid of Figs. 4, 5 and 6, the voltage-current characteristic can be determined from Fig. 3 over a range of ambient temperature from −20°C to +120°C, with no error of practical significance.

### Example of Obtaining the Voltage-Current Characteristic at Various Ambient Temperatures and in Different Surrounding Media

Consider an F23 thermistor in free air (using nominal values):

$R_o$ at 20°C = 2000Ω
$B$ = 3050°K
$k$ = 0·85mW/°C

$E_{max}$ at 20°C from equation (17) and Fig. 1

$$= 187\sqrt{\left(\frac{0\cdot85 \times 2}{3050}\right)} = 4\cdot4\text{V}$$

$I_m = E_{max}/R_o = 2\cdot2$mA.
Since $T_o = 20$°C, $F = B$.

Using Fig. 3, Table 1(a) can be compiled from the numerical values of $E$ and $I$.

Current $I$ is plotted against voltage $E$ in Fig. 7.

To calculate the voltage-current characteristic of this thermistor in free air at an ambient temperature of, say, 70°C it is first necessary to calculate $F$.

i.e. $F = 0\cdot854\ B$ from Fig. 4
$F = 2600$°K

From Fig. 5, $E_{max}$ at 70°C = 0·55 × 4·4
= 2·42V
and from Fig. 6, $I_m$ at 70°C = 2·5 × 2·2
= 5·5mA

Table 1(b) shows the calculation of the voltage-current characteristic under these conditions.

If the F23 is now immersed in water at a constant temperature of 70°C, the dissipation constant $k$ increases from 0·85mW/°C to 3·5mW/°C, which is approximately in the ratio of the thermal conductivity of air to that of water. From equation (17) it can be seen that $E_{max}$ is proportional to the square root of the dissipation constant. Hence, the value of $E_{max}$ under these conditions is 4·92V. Since $I_m$ is modified in the same proportion as $E_{max}$, its value becomes 11·2mA. A new table can be constructed using the new values of $E_{max}$ and $I_m$, and the voltage-current curve drawn and plotted as shown in Fig. 7.

It can be seen that the scales of Fig. 7 are logarithmic. This enables the whole of the useful current and voltage of the thermistor to be plotted with reasonable accuracy. If the characteristic were fitted with linear axes the characteristic could only be drawn accurately over about a decade of voltage and current compared with say three decades with logarithmic axes.

**TABLE 1(a)**
*(In free air at 20°C)*

| $I/I_m$ | $E/E_{max}$ | I (mA) | E (V) |
|---|---|---|---|
| 60 | 0·395 | 13·2 | 1·74 |
| 50 | 0·429 | 11·0 | 1·89 |
| 40 | 0·47 | 8·8 | 2·07 |
| 30 | 0·531 | 6·6 | 2·34 |
| 25 | 0·575 | 5·5 | 2·53 |
| 20 | 0·631 | 4·4 | 2·78 |
| 15 | 0·705 | 3·3 | 3·1 |
| 10 | 0·82 | 2·2 | 3·6 |
| 8 | 0·87 | 1·76 | 3·83 |
| 6 | 0·93 | 1·32 | 4·09 |
| 5 | 0·975 | 1·1 | 4·29 |
| 3 | 1·0 | 0·66 | 4·4 |
| 2 | 0·97 | 0·44 | 4·27 |
| 1·5 | 0·89 | 0·33 | 3·92 |
| 1·0 | 0·74 | 0·22 | 3·25 |
| 0·5 | 0·45 | 0·11 | 2·0 |

**TABLE 1(b)**
*(In free air at 70°C)*

| $I/I_m'$ | $E/E_{max}'$ | I (mA) | E (V) |
|---|---|---|---|
| 40 | 0·51 | 220 | 1·23 |
| 30 | 0·572 | 165 | 1·38 |
| 25 | 0·614 | 137 | 1·48 |
| 20 | 0·67 | 110 | 1·62 |
| 15 | 0·735 | 82·5 | 1·78 |
| 10 | 0·835 | 55 | 2·02 |
| 8 | 0·885 | 44 | 2·14 |
| 6 | 0·945 | 33 | 2·28 |
| 5 | 0·98 | 27·5 | 2·37 |
| 3 | 1·0 | 16·5 | 2·42 |
| 1·5 | 0·89 | 8·25 | 2·15 |
| 1·0 | 0·74 | 5·5 | 1·79 |
| 0·7 | 0·59 | 3·85 | 1·43 |
| 0·5 | 0·45 | 2·75 | 1·09 |
| 0·3 | 0·29 | 1·65 | 0·70 |
| 0·2 | 0·197 | 1·1 | 0·48 |

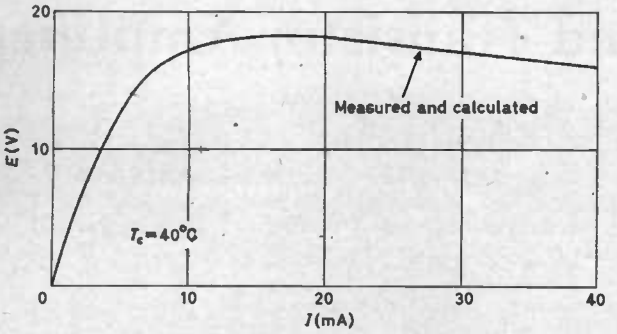



**Fig. 9. Voltage-current characteristic of an STC KS37 disk type thermistor**

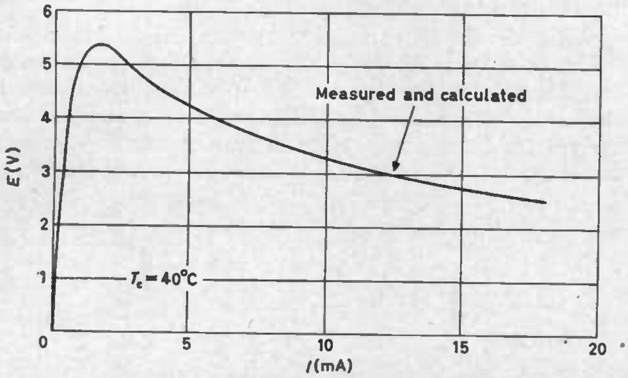
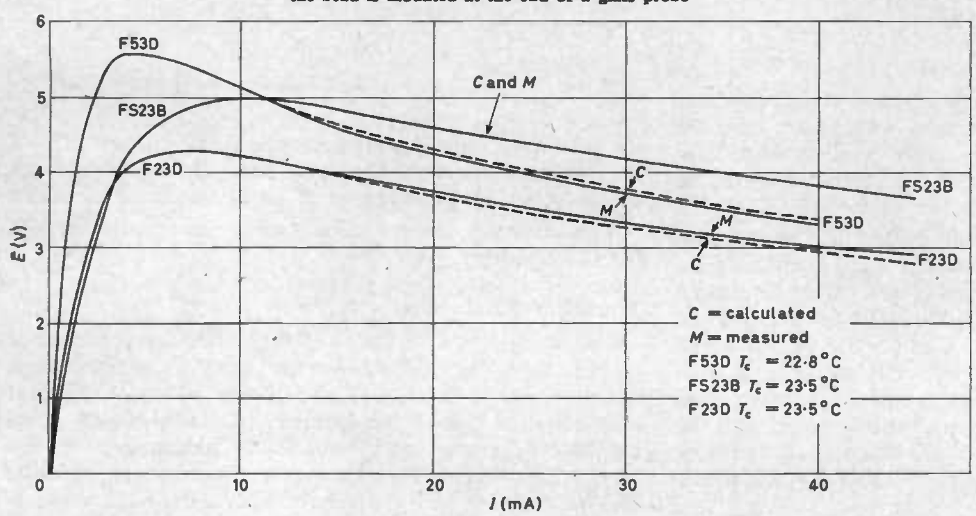



**Fig. 10. Voltage-current characteristic of an STC type A24 bead thermistor (bead inside a gas-filled envelope)**

A resistive electrical load line can be drawn on a characteristic plotted by either method. In the case of the linear axes it is a straight line and for the logarithmic axes it is a rectangular hyperbola of constant shape. The rectangular hyperbola is described in Appendix D and an example of it is shown in Fig. 7. If the thermistor is operated at 5V open-circuit voltage and 30mA short-circuit current its resistance under these conditions can be determined by the intersection of the load line with the voltage-current characteristic, i.e.

| CONDITION (See Fig. 7) | VOLTAGE (V) | CURRENT (mA) | RESISTANCE (Ω) | $T_a$ (°C) | OPERATING MEDIUM | MEAN TEMP. OF BEAD (°C) |
|---|---|---|---|---|---|---|
| 1 | 4·3 | 4·5 | 957 | 20 | Free Air | 43 |
| 2 | 2·4 | 15·5 | 161 | 70 | Free Air | 114 |
| 3 | 3·45 | 10·0 | 340 | 70 | Still Water | 80 |

**Fig. 11. Voltage-current characteristics of various bead thermistors where the bead is mounted at the end of a glass probe**



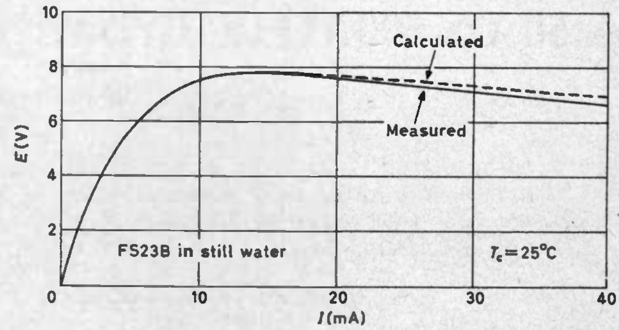




**Fig. 12. Voltage-current characteristic of a bead thermistor immersed in water**

The foregoing method of calculating the voltage-current characteristic is sufficiently accurate if the heat transfer from the thermistor is by convection and conduction when mounted in free air. Figs. 9, 10 and 11 show the comparison between calculated and measured characteristics where the calculated curves are based on the measured values of $E_{max}$ and $R_o$. It would appear that the difference between the measured and the predicted voltage-current characteristic is sufficiently small for most practical purposes even when the mean temperature of the thermistor is 300°C. In the STC range of thermistors, the types whose behaviour can be calculated by this method are:

A, D, E, F, FS, G, GT, KR, KU, KB, M, P and U.

Deviations from the calculated curves occur in practice when the thermistor is immersed in a liquid. This is because heat from the thermistor causes the temperature of the liquid immediately surrounding it to rise. However, the error is not very significant if one compares it with the departure from the nominal characteristic, due to the tolerances on the parameters of the thermistor. Fig. 12 shows the calculated and measured characteristic of an FS23B thermistor immersed in still water.

The curves in Figs. 9, 10 and 11 were obtained by measuring the thermistor inside a small (5in³) enclosure which was in the direct path of a temperature controlled jet of air. These conditions closely simulated those of free air. The temperature of the air within the enclosure was controlled to better than 0·1°C during the plotting of the voltage-current characteristic (to avoid inaccuracies due to thermal time-constant each characteristic took at least two hours to plot on an *X-Y* recorder). The accuracy of the measurement of voltage and current was to within 0·5 per cent. The experiment involving the immersion of a thermistor in water was carried out in a temperature-controlled room.

(*To be continued*)

# A 50 to 500MHz Broadband Transistor Amplifier

By A. E. Hilling*, B.Sc.(Eng)

*A broadband amplifier is described which has a gain of 21·5 ± 1·5dB from 50 to 500MHz. A maximum noise figure of 8dB is obtained at 500MHz and the maximum input and output v.s.w.r. are 2·0:1 with respect to 50Ω. Low third order intermodulation distortion is achieved by using the BFY90 transistor.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 408)

SMALL signal broadband amplifiers are used extensively in multi-channel communication equipments. They can be used, for example, as aerial amplifiers in regions of low field strength and in cable distribution systems to compensate for cable losses. These amplifiers can also be converted to narrower band operation by the use of external filters.

The advantage of the amplifier described here is that, due to its very wide bandwidth, it can cover the requirements of many narrower band amplifiers and in general give a better distortion performance. One amplifier, which can be used without modification or adjustment in many different systems, aids standardization and can bring considerable economic benefits. The use of transistors make the amplifier rugged and reliable and reduce the power supply requirement.

There would appear to be a wide application for a general purpose amplifier of this type.

### Amplifier Requirements

The system for which this amplifier was designed required a gain of approximately 22dB between 50 and 500MHz, a maximum noise figure of 8dB and input and output voltage standing wave ratios not greater than 2·0:1 with respect to 50Ω. The third order distortion generated in the amplifier was required to be as small as possible. Second order distortion was not considered to be important as the amplifier would be preceded by a band-pass filter. The largest signal occurring at the input to the amplifier is likely to be of magnitude −30dBm.

### Transistors

Four Mullard BFY90 transistors are used in this amplifier. These are silicon npn epitaxial planar transistors whose features include low noise, low third harmonic distortion and a minimum $f_T$ of 1GHz between 2 and 20mA collector current.

### Circuit Configuration

The two configurations commonly found in transistor broadband amplifiers are common base and common emitter. To obtain power gain in the common base configuration interstage impedance transformers must be used.

The advantage of this method is that the power gain of a stage is not primarily dependent on the current gain of the transistor. Ferrite core transformers have been tried and were successful in single stage amplifiers covering the band 50 to 500MHz where the load on the output transformer secondary was a 50Ω resistor. However, this technique was not successful when applied to multi-stage amplifiers and a flat gain characteristic could not be obtained. This was attributed to the change of transistor input impedance with frequency presenting a varying load impedance to the previous transistor.

With the advent of transistors with cut-off frequencies ($f_T$) greater than 1GHz, it has become feasible to design broadband amplifiers up to 500MHz using transistors in the common emitter configuration. This technique uses the current gain of the device to provide amplification, the advantage of this method is evident in the elimination of

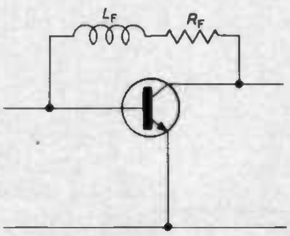



Fig. 1. Stage with shunt feedback

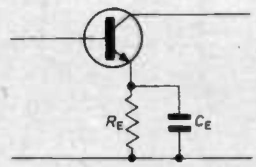



Fig. 2. Stage with series feedback

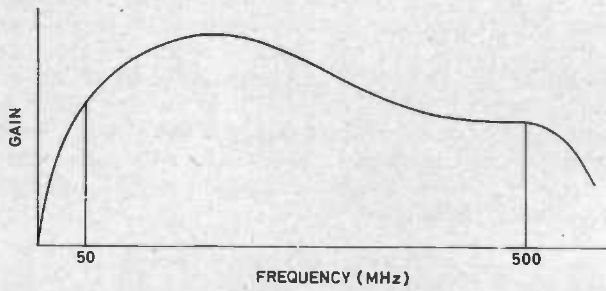



Fig. 3. Sketch of frequency response shunt feedback stage

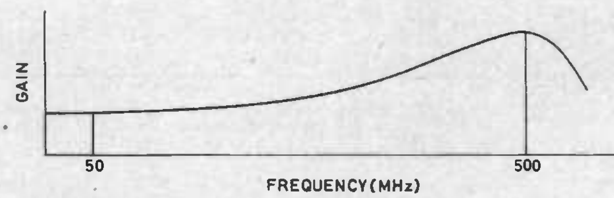



Fig. 4. Sketch of frequency response series feedback stage

the interstage impedance transformer. The current gain of the transistor is, however, frequency dependent but can be controlled by the addition of simple feedback networks.

The common emitter configuration has been adopted for this amplifier.

### Transistor Parameters

The design and analysis of high frequency transistor amplifiers is complicated by the lack of an equivalent circuit which is directly related to the transistor physical characteristics. As a consequence the analysis of these amplifiers is based on 'black box' parameters, in this case $y$ parameters. These have the disadvantage of being current, voltage and frequency dependent. Each stage is

* *Mullard Research Laboratories.*

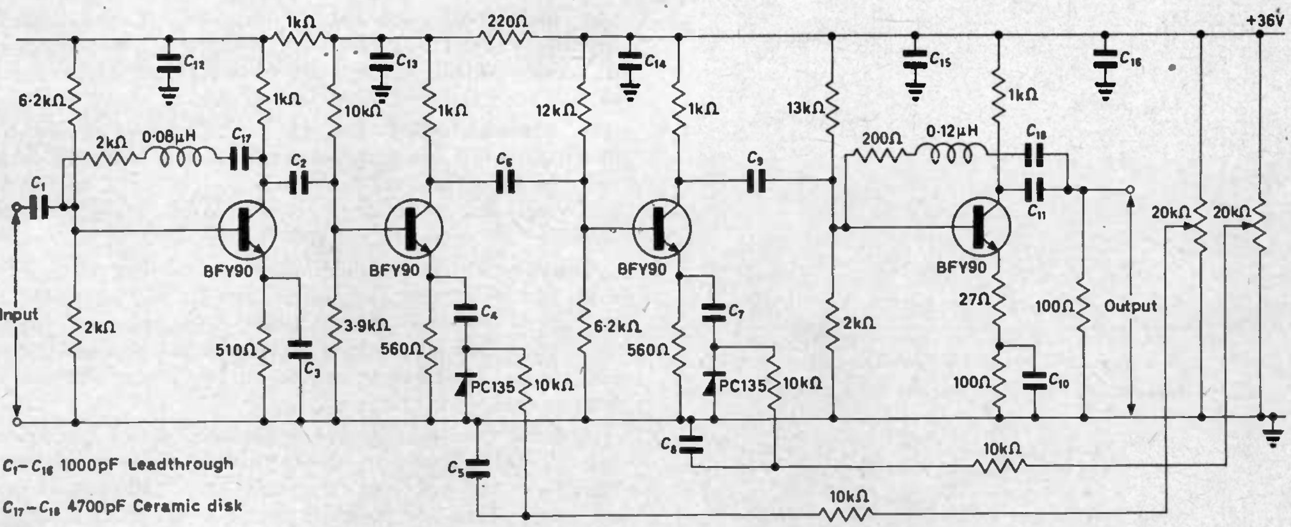



**Fig. 5. A four stage common emitter feedback amplifier using the BFY90 transistor**

complicated by the presence of frequency dependent passive feedback networks.

### Feedback Networks

Two simple forms of feedback have been employed which have been called 'shunt' and 'series' feedback (Figs. 1 and 2). The application of shunt feedback to an amplifier stage lowers the input and output impedances while the application of series feedback increases them. Each feedback network contains a component whose impedance varies with frequency. These components have been chosen so that the effect of the feedback networks decreases with increasing frequency. Further aspects of shunt and series feedback amplifiers are discussed elsewhere[1,3]. Feedback around two stages has also been examined but, due to phase changes in the transistors, can lead to instability at frequencies approaching 500MHz.

### Circuit Design

By taking the values of transistor $y$ parameters at discrete frequencies in the band, at specified operating points, it would be possible, with the aid of a computer, to vary the several feedback elements until the required gain characteristic and input and output v.s.w.r. were obtained. However, the advantage of this very complicated method over an empirical approach is debatable. It is also difficult, at these frequencies, to account for parasitic capacitances and inductances and a purely theoretical design would always require experimental adjustment.

The compromise technique that has been adopted is to assume that, at the highest frequency of interest, the passive feedback networks no longer have any effect and the amplifier is a simple cascade of transistors. The power gain and input admittance of each stage can then be calculated from the transistor $y$ parameters. The sum of the individual stage gains is the maximum gain obtainable from the amplifier at the highest frequency in the band. The gain of the amplifier at the lower frequencies is then reduced to this value by empirical adjustment of the feedback networks. It is assumed that the d.c. bias components and the coupling capacitors have no effect on the operation of the amplifier at the highest frequency in the band.

As an example the calculation of the input admittance and power gain of a BFY90 transistor at 500MHz with a 50Ω collector load is shown in the appendix. This is, in fact, the output stage. Similar calculations can be performed for the preceding stages where the 50Ω load is replaced by the input admittance of the following stage.

From these calculations it would appear that three stages of amplification would meet the gain requirements. However, a minimum of four stages was found to be necessary to obtain adequate control of the amplifier input and output impedances and shape and magnitude of the gain characteristic.

The first stage employs shunt feedback to obtain a good input v.s.w.r. The input impedance of the BFY90 is approximately 400Ω at low frequencies falling to approximately 50Ω at 500MHz. The frequency dependent shunt feedback reduces the input impedance of the first stage to approximately 50Ω over the frequency band. However,

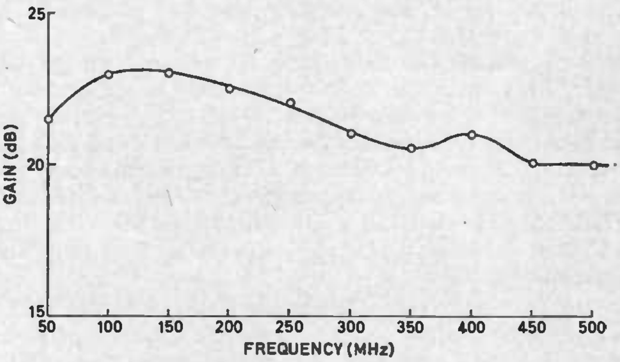
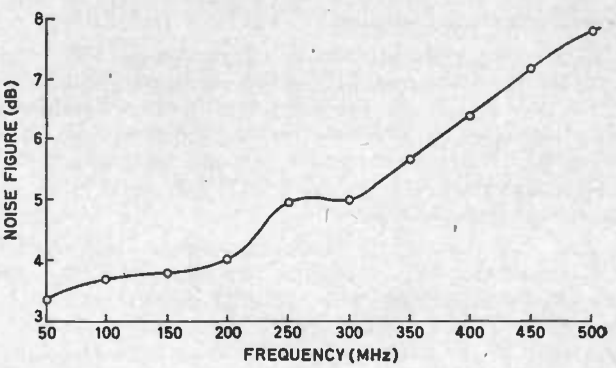



**Fig. 6. Gain-frequency response: amplifier using BFY90 transistors**

**Fig. 7. Noise figure/frequency: amplifier using BFY90 transistors**



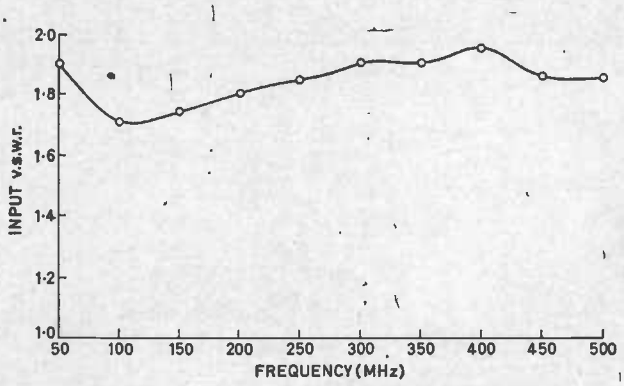



**Fig. 8. Input v.s.w.r./frequency: amplifier using BFY90 transistors**

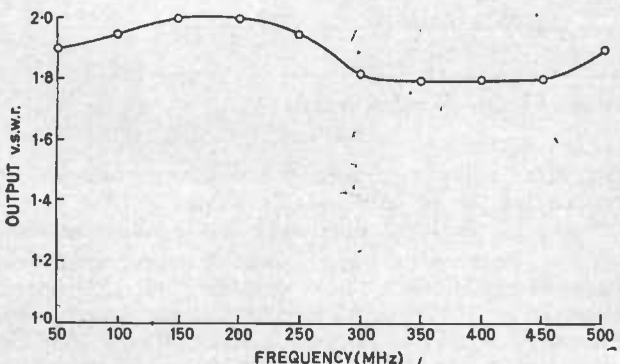



**Fig. 9. Output v.s.w.r./frequency: amplifier using BFY90 transistors**

the frequency response of this stage when adjusted for optimum v.s.w.r. is not flat (Fig. 3). To compensate for this the second stage has series feedback giving a frequency response as shown in Fig. 4. The third stage also uses series feedback and compensates for the fourth stage whose shunt feedback network partially controls the output impedance of the amplifier. It was impossible to obtain an output v.s.w.r. of less than 2·0:1 over the band by only altering the shunt feedback network on the last stage. By plotting the output admittance of the amplifier, for frequencies between 50 and 500MHz, on a Smith chart and by experimental adjustment it was found that an undecoupled emitter resistor of 27Ω in the last stage and a 100Ω resistor across the output gave the required performance. The loss in gain incurred by these resistors partially explains the need for a four stage amplifier mentioned earlier.

The gain and input and output impedances depend on six related variables and to obtain the required performance it was necessary to make several successive adjustments.

The circuit diagram is shown in Fig. 5.

### Intermodulation Distortion

When two input signals, at frequencies $f_1$ and $f_2$, are injected into an amplifier, intermodulation products of these two signals are produced due to non-linearities in the transistors. In addition to harmonics second order terms occur at frequencies $f_1 \pm f_2$ and third order terms at frequencies $2f_2 \pm f_1$ and $2f_1 \pm f_2$. Higher order terms also occur but are less significant.

The magnitude of the output power at the distortion frequencies has been measured relative to the magnitude of the wanted signals with specified input signals to the amplifier. The distortion signals have been referred to the input of the amplifier by the gain at these frequencies and have been denoted 'Equivalent Intermodulation Inputs' (e.i.i.). This is the level of signal, at the distortion frequency, which would have to be present at the input to give the measured power at the output.

A reference signal level of −30dBm was chosen for intermodulation measurements, and the e.i.i. obtained when two signals of this magnitude are injected into the amplifier gives an indication of the dynamic range of the amplifier.

Third order intermodulation products only have been measured on this amplifier as it is intended to be used with a preselection filter with a bandwidth less than one octave. Second order products occur only in an amplifier when the input circuit pass-band is greater than this.

The BFY90 transistors are operated under conditions for minimum distortion, the transistor in the fourth stage being selected for low distortion at a collector current of 20mA. The first stage transistor is operated in conditions which are a compromise between low noise and low distortion.

With the two input signals at frequencies $f_1$ and $f_2$ each of magnitude −30Bm the performance shown in Table 1 was obtained.

These figures for e.i.i. are typically 20dB better than has been obtained with other small signal transistors.

### D.C. Conditions

A supply voltage of 36V was used for this amplifier. The amplifier stages were operated under the conditions shown in Table 2.

The total current requirement for the amplifier was approximately 70mA.

### Construction

Due to the high cut-off frequency ($f_T$) of the BFY90 transistor great care has been exercised in eliminating conditions for spurious oscillation. The amplifier was constructed on a brass chassis which was then mounted in a brass box. The r.f. path through the amplifier was made as compact as possible by keeping lead lengths short. Special care has been taken to minimize emitter lead lengths. Capacitors have been chosen for low parasitic inductance and the coupling capacitors are in fact tubular decoupling capacitors. It is possible for a combination of parasitic capacitances and inductances to cause the amplifier to oscillate at frequencies above 1GHz.

**TABLE 1**

| $f_1$ (MHz) | $f_2$ (MHz) | DISTORTION FREQUENCY (MHz) | E.I.I. (3rd Order) (dBm) |
|---|---|---|---|
| 122 | 156 | 190 | −80 |
| 122 | 92 | 62 | −82 |
| 314 | 256 | 372 | −85 |
| 480 | 440 | 400 | −82 |

**TABLE 2**

| STAGE | COLLECTOR VOLTAGE (V) | COLLECTOR CURRENT (mA) |
|---|---|---|
| 1 | 9 | 7·5 |
| 2 | 9 | 13·5 |
| 3 | 9 | 17·5 |
| 4 | 8 | 20 |

The emitter feedback capacitors in stages 2 and 3 are voltage variable capacitors which allow adjustment to be made away from the r.f. circuit. The transistors are mounted in Jermyn heatsinks which are bolted to the chassis.

**Performance**

The performance of the prototype amplifier is given in Figs. 6 to 9. The 1dB gain compression point of this amplifier occurs at an output power of approximately +7dBm.

**Conclusions**

This design method has the advantage of obtaining a flat gain characteristic and good input and output impedance matches from simple cascaded stages with the minimum theoretical calculation. A feature of this amplifier is the low third order distortion which is achieved by the use of the BFY90 transistor.

**Acknowledgments**

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The author wishes to acknowledge the assistance rendered by S. J. Robinson and S. K. Salmon and to thank the Director of Mullard Research Laboratories for permission to publish this article.

### Appendix

The typical $y$ parameters of development samples of BFY90 transistors at 500MHz are:

$$y_{ie} = 11 + 6{\cdot}5j \text{ mmho}$$
$$y_{fe} = 2 - 50j \text{ mmho}$$
$$y_{oe} = 0{\cdot}5 + 5{\cdot}5j \text{ mmho}$$
$$y_{re} = -0{\cdot}1 - 3j \text{ mmho}$$

Measured at a collector voltage of 10V and an emitter current of 20mA.

*Calculation of Input Admittance and Power Gain*

It can be shown that the input admittance of a transistor in the common emitter configuration can be described by the formula[3].

$$Y_{IN} = y_{ie} - \frac{y_{fe} \cdot y_{re}}{y_{oe} + y_L}$$

where $y_L$ is the load admittance.

The power gain of transistor under these conditions is[3]:

$$\text{P.G.} = \left|\frac{y_{fe}}{y_{oe} + y_L}\right|^2 \cdot g_L/G_{IN}$$

where $g_L$ is the real part of the load admittance.

$G_{IN}$ is the real part of the input admittance.

Substituting the relevant $y$ parameters into these equations gives, with $y_L = 20$ mmho (i.e. $R_L = 50\Omega$)

$$Y_{IN} = 11 + 6{\cdot}5j - \left(\frac{(-0{\cdot}1-3j)(2-50j)}{0{\cdot}5+5{\cdot}5j+20}\right) \text{mmho}$$

$$\underline{Y_{IN} = 18 + 4{\cdot}7j \text{ mmho}}$$

$$\text{P.G.} = \left|\frac{2-50j}{0{\cdot}5 + 5{\cdot}5j + 20}\right|^2 \cdot 20/18$$

$$\underline{\text{P.G.} = 6{\cdot}3 \text{ (8{\cdot}0dB)}}$$

These calculations can be repeated for the penultimate stage where:

$$y_L = Y_{IN} \text{ (of the last stage)}$$
$$= 18 + 4{\cdot}7j \text{ mmho.}$$

### References

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## Thin and Thick Film Circuits

The microengraving machine developed by Standard Telephones & Cables Ltd for film circuits and shown in prototype last year in London, was recently demonstrated in use at the company's Paignton factory. The machine was operated by a punched paper tape produced locally from a remote teleprinter at Harlow in Essex. Both the photomasters for conductor patterns and the actual resistor networks can be cut. The advantage is that no accurate drawings or photographic reduction are required as the paper tape input can obtain all the necessary information. Both thin film resistors and capacitors can be adjusted accurately in value after production by eroding; resistors may be produced to a tolerance of ±0·05 per cent and low-valued capacitors to ±1pF or 0·5 per cent, whichever is the greater.

A reactive sputterer, another development in use at Paignton, produces silicon dioxide layers as the dielectric of thin film capacitors. The source is pure silicon, particles of which are oxidized and eroded by bombardment with oxygen ions from a plasma of argon and oxygen. These ions are initiated by a source of electrons which derive their energy both from the anode voltage and a magnetic field, so that their path in the plasma is quite long. The apparatus has a much higher yield than most sputtering units and produces a silicon dioxide layer of 160cm$^2$ in one operation.

At a conference held at the factory the field of thick and thin film circuits and their relationship with semiconductor integrated circuits, was discussed. Standard Telephones & Cables Ltd announced that full-scale production of thick film circuits has started in their Film Circuit Unit; the target is 0·5 million circuits a year initially, with a possible 6 million in 1972. During the discussion G. Thornton, Products Marketing Manager of the Capacitor Division, pointed out that at an early stage in the development of semiconductor integrated circuits it seemed clear that film circuits were the real solution to miniaturization. Now it appears that s.c.i.c.'s and films are complementary solutions. Some competition is inevitable but broadly speaking film circuits are suited to passive networks and s.c.i.c.'s to digital and other standard circuits. Thornton considered that thin and thick films can compete with the discrete resistor, the market in which is currently estimated at £9M. Thick films may also serve as replacement in equipment involving s.c.i.c.'s for printed circuit boards; thick film substrates with resistors are useful to mount any of the standard s.c.i. circuits. In reply to a question asked on the lateness of S.T.C. in the thick film business, Thornton indicated that their application was in a state of flux and no disadvantage would be apparent.

An extrapolation of market needs in microelectronics to 1972 has been made and the general opinion is that it will be worth about £25M, of which £8M could well be for films. Thin films, largely in military equipment, will account for a little over a half (in value) and the remainder will be thick films in large numbers of lower priced circuits. The film market as a whole is estimated, said Thornton, to split between hybrid, i.e. a combination of film and s.c.i.c., and passive in the following way: for thick films the split will be 50/50 by value and for thin films 2 to 1 in favour of hybrids.

# Comparator-Hold Circuit using the Facing-Coupled Esaki Diode Pair

By Y. Murata

*By the use of the facing-coupled Esaki diode pair circuit, widely applicable circuits with memory and comparing functions are obtained.*

*The operation of the facing-coupled Esaki diode pair circuit was studied through several experiments and by numerical analysis using a digital computer. The operating limitations of the circuit as to load, drive, response speed, etc. was clarified for the case of SONY type-1T1101 Esaki diodes. The circuit was shown to serve as a good memory element and a switching element of high speed.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

THE Esaki diode, invented by R. Esaki in 1957, is a two-terminal semiconductor element having a negative resistance region. The element has so many merits that it has been continually developed and applied to various circuits since its invention.

The circuit applications of the Esaki diode may be divided broadly into two classes, namely single circuit and pair circuit. The former is well known in discriminator circuits for pulse height discriminators, one-shot circuits, etc. As to the latter type, it includes a cascade-coupled Esaki diode pair circuit (Goto pair circuit)[1,2] developed by E. Goto in 1959, and the facing-coupled Esaki diode circuit (f.c.e.d.) developed in 1964. The f.c.e.d. circuit[3] acts as a current switch and a memory.

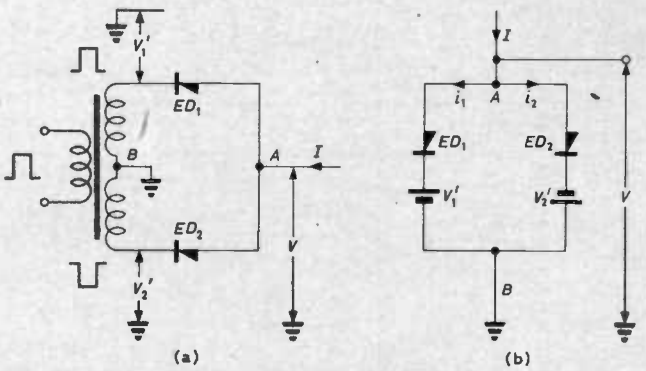



Fig. 1. Basic circuit

The present article describes the operation of the f.c.e.d. circuit and its limitations.

## Action Mechanism

Fig. 1(a) is the basic diagram of this circuit, and for the sake of simplified analysis it is redrawn as shown in Fig. 1(b). In these figures, the notation conventions are as follows:

$ED_j$ = Esaki diode numbered $j$ ($j = 1, 2$),
$I$ = incident current,
$v$ = output voltage,
$i_j$ = current in $ED_j$,
$V_j'$ = drive voltage into $ED_j$.

The operating characteristics of this pair (a closed circuit consisting of $ED_1$ and $ED_2$) viewed from the forward direction of $ED_1$ is shown in Figs. 2(a) and (b).

When the drive voltages $V_j' = 0$, this pair has three equilibrium states $v_{01}$, $v_{02}$, and $v_{03}$ for a certain value of $I$ as shown in Fig. 2(a): effectively, however, $v$ has two states, as $v_{02}$ is in the negative resistance region.

As for $ED_j$, letting

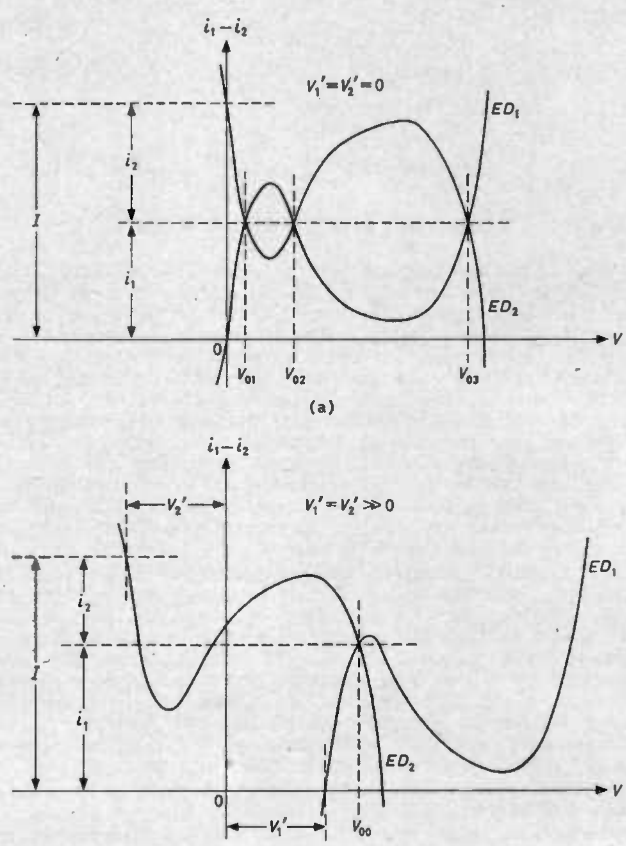
peak current $= I_{jp}$
peak voltage $= V_{jp}$
valley current $= I_{jv}$
valley voltage $= V_{jv}$
forward voltage $= V_{jf}$,

the value of $I$ mentioned above is expressed as:

$$I_{1p} + I_{2p} \, (= I_{max}) > I > I_{1v} + I_{2v} \, (= I_{min}) \quad \dots . \quad (1)$$

where it is assumed that $V_{1p} = V_{2p}$ and $V_{1v} = V_{2v}$. As far as $I$ satisfies equation (1), and provided $V_1' \times V_2' < 0$ and $V_0' (= |V_1'| + |V_2'|)$ is more than a certain value defined below, the number of equilibrium states becomes only one, $v_{00}$, as shown in Fig. 2(b). Though this value of $V_0'$ depends on the characteristic of the diode used and on the value of $I$, the necessary and sufficient condition for the value

Fig. 2. Operating characteristics




* *Musashi Institute of Technology, Japan.*

of $V_o$ is:

$$V_o' > V_{1p} + V_{2f} \ \ldots\ldots\ldots\ldots\ldots\ldots \ (2)$$
$$V_o' > V_{2p} + V_{1f} \ \ldots\ldots\ldots\ldots\ldots \ (2^+)$$

where $^+$ denotes that the polarities of $V_j'$ are opposite to that shown in Fig. 1.

Provided that $V_j'$ increases continuously from zero to the region of equation (2) (Process-I) and subsequently decreases to zero (Process-II) under the condition of equation (1), the equilibrium state changes are as follows:

$$\left.\begin{matrix} \nu_{01} \\ \text{or } \nu_{03} \end{matrix}\right\} \xrightarrow{\textit{Process-I}} \nu_{00} \xrightarrow{\textit{Process-II}} \left\{\begin{matrix} \nu_{01} \\ \text{or } \nu_{03} \end{matrix}\right.$$

In Process-I the final value of $\nu$ becomes $\nu_{00}$, independent of the initial value ($\nu_{01}$ or $\nu_{03}$); while in Process-II the final value of $\nu$ is divided into either of the two cases, $\nu_{01}$ or $\nu_{03}$, and is explained by the up-and-down relation between the peak point of $ED_1$ and valley point of $ED_2$ in Fig. 2(b).

That is, letting:

$$I_{1p} + I_{2v} = I_s\,, \ \ldots\ldots\ldots\ldots\ldots\ldots \ (4)$$
$$I_{1v} + I_{2p} = I_s\,, \ \ldots\ldots\ldots\ldots\ldots\ldots \ (4a)$$

then $I_s$ is the threshold value that defines the final value of $\nu_{01}$ or $\nu_{03}$.

Therefore, when the incident current $I$ satisfies equation (1): (i) if the drives $V_j' = 0$, this pair operates as a memory circuit with two equilibrium states; (ii) if the drive is of suitable pulse shape so that $V_j'_{(min)} = 0$ and $V_j'_{(max)}$ satisfies the condition of equation (2), this pair operates as a current switch with a threshold value of $I_s$.

In this pair, as $ED_2$ is equivalent and complementary to $ED_1$ all the aforesaid statements as to $ED_1$ are applicable to $ED_2$.

The $V' - \nu$ characteristic curve with parameter $I$ calculated by a digital computer is shown in Fig. 3, where $ED_1$ is SONY type-1T1104 ($I_p = 6$mA)*, $ED_2$ SONY type-1T-1101 ($I_p = 2$mA)* and the drive voltage $V' = V_1' = -V_2'$. Fig. 4 shows the $\nu - I$ characteristic curve with parameter $V'$ under the same conditions as Fig. 3.

* *See Appendix.*

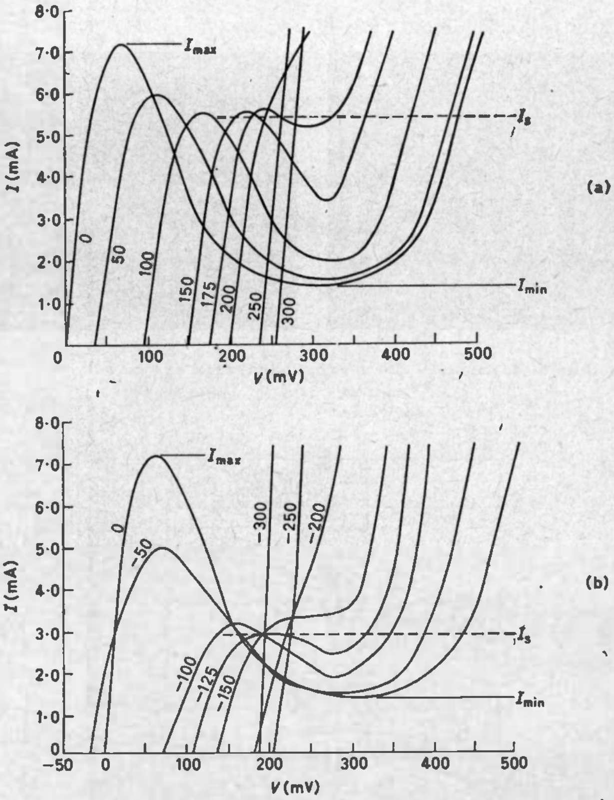
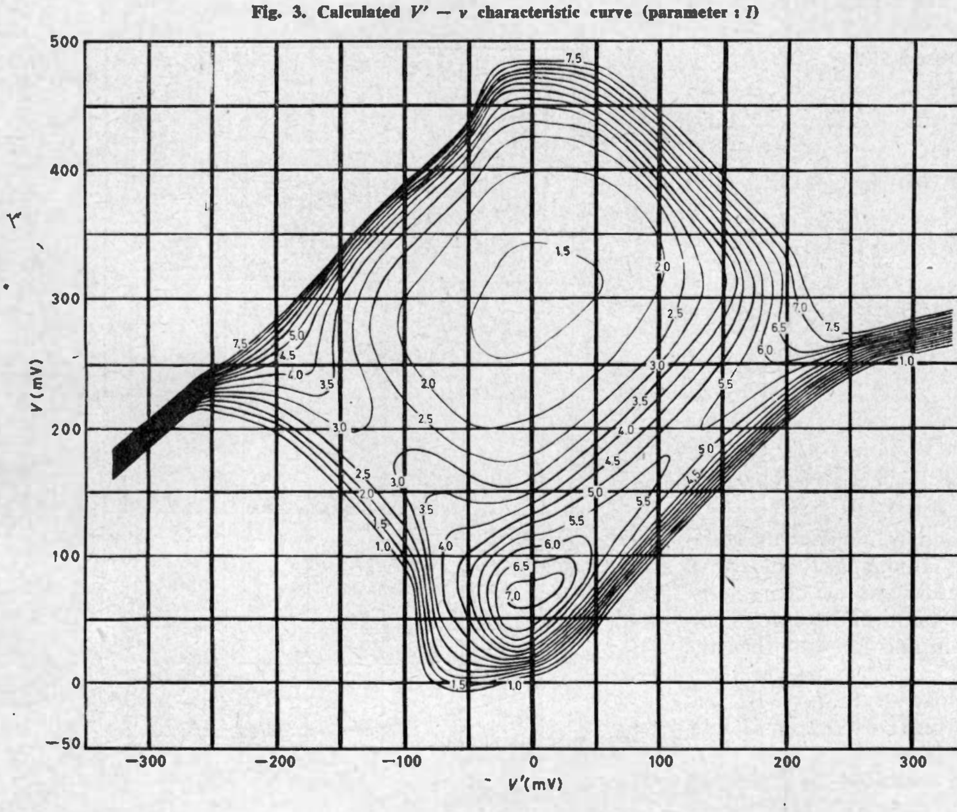



**Fig. 4. (a) Calculated $\nu - I$ characteristic curve (parameter : $V'$)**
**(b) Calculated $\nu - I$ characteristic curve (parameter : $V'$)**

**Fig. 3. Calculated $V' - \nu$ characteristic curve (parameter : $I$)**




From these figures the value $I_s$ is:

(i) $5{\cdot}5\text{mA} > I_s > 5{\cdot}0\text{mA}$ for $V' > 0$

(ii) $3{\cdot}0\text{mA} > I_s > 2{\cdot}5\text{mA}$ for $V' < 0$

thereby equations (1) and (4) are found to be satisfied.

If $ED_1$ and $ED_2$ have the same characteristics, the curve in Fig. 3 becomes symmetric to the axis of $V' = 0$, and Figs. 4(a) and (b) coincide with each other, which is illustrated in reference (3).

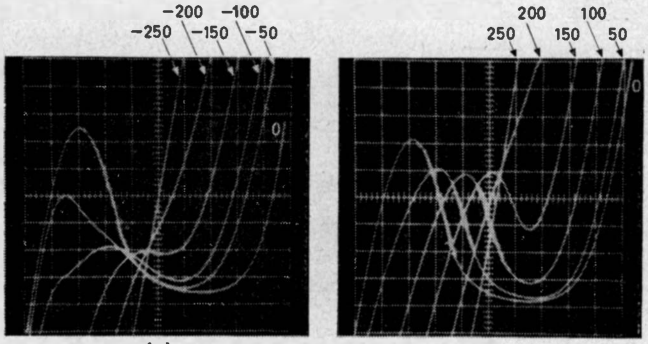
Fig. 5 shows the characteristic curve obtained by the characteristic curve tracer and corresponds to Fig. 4.

Considering the fluctuation of characteristics of Esaki diodes, Figs. 4 and 5 are in excellent accord. This shows that the present interpretation of the action mechanism of the f.c.e.d. circuit may be reasonable.

## Limitation of Operation

### Equivalent Circuit

Fig. 6 shows the equiva-




(a) (b)

**Fig. 5. Experimental $v - I$ characteristic curve (parameter : $V'$)**
*(a) V : 1mA/Div., (b) H : 50mV/Div.*

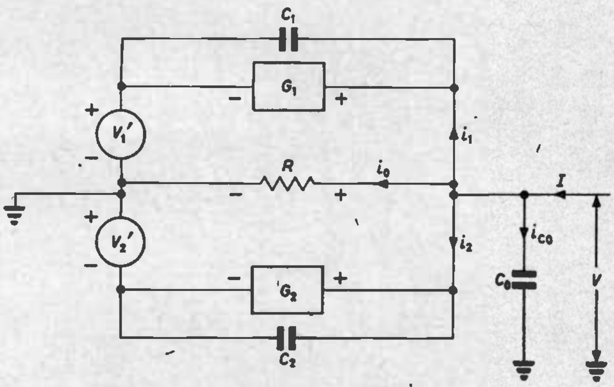



**Fig. 6. Equivalent circuit**

lent circuit of Fig. 1, where the internal inductance and load inductance are neglected, for they appear to be less than $10^{-9}$ H. $R$ expresses the total impedance of the load and incident current source. $G_j$ is static characteristic defined by

$$i = G_j(v) \quad (j = 1,2), \ldots\ldots\ldots\ldots\ldots \quad (5)$$

where $i$ is the current when terminal voltage $V$ is applied to $ED_j$.

The currents in Fig. 6 are expressed as:

$$I = i_{00} + i_0 + i_1 + i_2, \ldots\ldots\ldots\ldots \quad (6)$$

$$i_{00} = C_0 \,.\, dv/dt, \ldots\ldots\ldots\ldots\ldots \quad (7)$$

$$i_1 = C_1 \,.\, d(v - V_1')/dt + G_1(v - V_1'), \ldots\ldots \quad (8)$$

$$i_2 = C_2 \,.\, d(v + V_2')/dt + G_2(v + V_2'), \ldots\ldots \quad (9)$$

$$v = i_0 \,.\, R, \ldots\ldots\ldots\ldots\ldots\ldots \quad (10)$$

where

$C_1 = C_2,$
$G_1(v) = G_2(v) = G(v),$
$V_1' = V_2' = V',$
$C_0 + C_1 + C_2 = C,$
$C \,.\, T = t,$

so that equations (6) and (10) can be expressed by the following equation

$$dv/dT = I - \{ G(v - V') + G(v + V') + v/R \} \quad . \ (11)$$

### Limitation of Load

When $dv/dT = 0$ and $V' = 0$ and $G$ is given by equation (11), the static characteristics among $I$, $v$, and $R$ of the f.c.e.d. are obtained. In the static state, the characteristic of $v$ and $I$ depending on the value of load $R$ is important, for it determines the fan-in and fan-out when the f.c.e.d. operates as a logic circuit.

Fig. 7 shows the relation of load $R$ to $I_{max}$, $I_s$, $I_{min}$, $v_{08(max)}$, $v_{03(min)}$, $v_{01(max)}$, $v_{01(min)}$; where $ED_1 = ED_2 =$ SONY type-1T1101.

These characteristic curves are obtained by computer calculation and are in good agreement with observed data. In this case, the permitted value of load is considered $R > 200$.

### Response Speed

Response speed of this pair is mainly due both to Process-I (rise-up time) and Process-II (fall-down time). In Process-I, as the final value of $v$ is always $v_{00}$, independent of its initial value ($v_{01}$ or $v_{08}$), the only problem is the time required for $v$ to reach a final value. In Process-II the final value is divided into $v_{03}$ or $v_{01}$ according to whether $I$ is larger or smaller than $I_s$. As $I_s$ also depends on the value of $dV'/dt$ (the definition of $I_s$ by equation (4) is the static case), the response speed cannot be expressed only in terms of the time required for $v$ to reach its final value. The response speed in Process-II will be discussed in the later section "change of $I_s$".

For the case that $G_j$ are typical characteristics of SONY type-1T1101 and $V'$ is a unit step drive, the output response in Process-I is calculated by the computer using the Runge-Kutter method.

Examples of analysed results are depicted in Fig. 8: (a) $V_{max}' = 400$mV with $I$ as parameter, (b) $I = (I_{max} + I_s)/2$ with $V_{max}'$ as parameter, (c) $I = (I_s + I_{min})/2$ with $V_{max}'$ as parameter; where the vertical axis is $v$ in mV and the horizontal axis is $T$ $(= t/c)$.

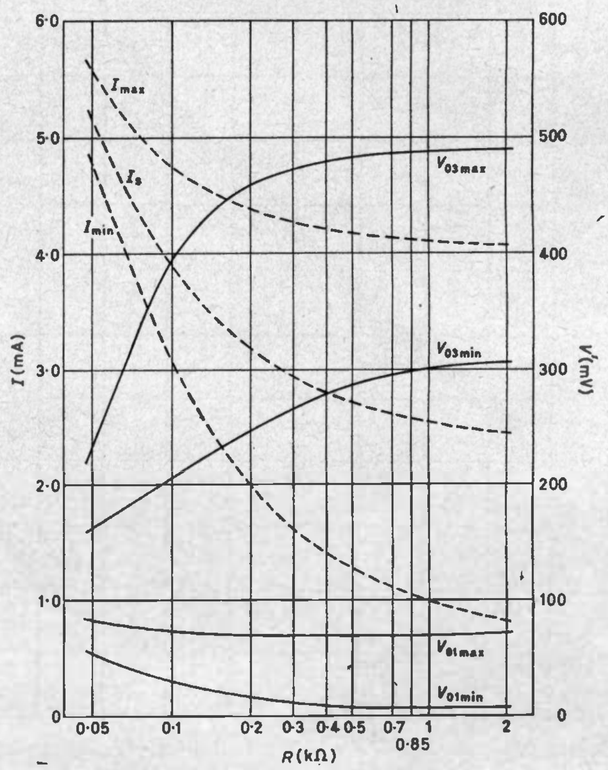
In Fig. 9, the observed data corresponding to Fig. 8 are shown*, and they are in good agreement. This fact shows the validity of the computer calculations as well as of the equivalent circuit shown in Fig. 6 for equation (11).

The calculated results concerning the relation of rise-time† to $R$ and $V_{max}'$ are shown in Figs. 10 and 11 respectively.

* *In experiment, the external capacitance C is added so that the time range may be equivalently enlarged using the relation Ct=t. Results showing the validity of enlarging the equivalent time in this kind of experiment have been reported*[4].

† *The rise time is defined as the time required, after unit step drive is added, for the differences, between v from $v_{08}$ and v from $v_{01}$ to become 1mV.*

**Fig. 7. Static load characteristic**



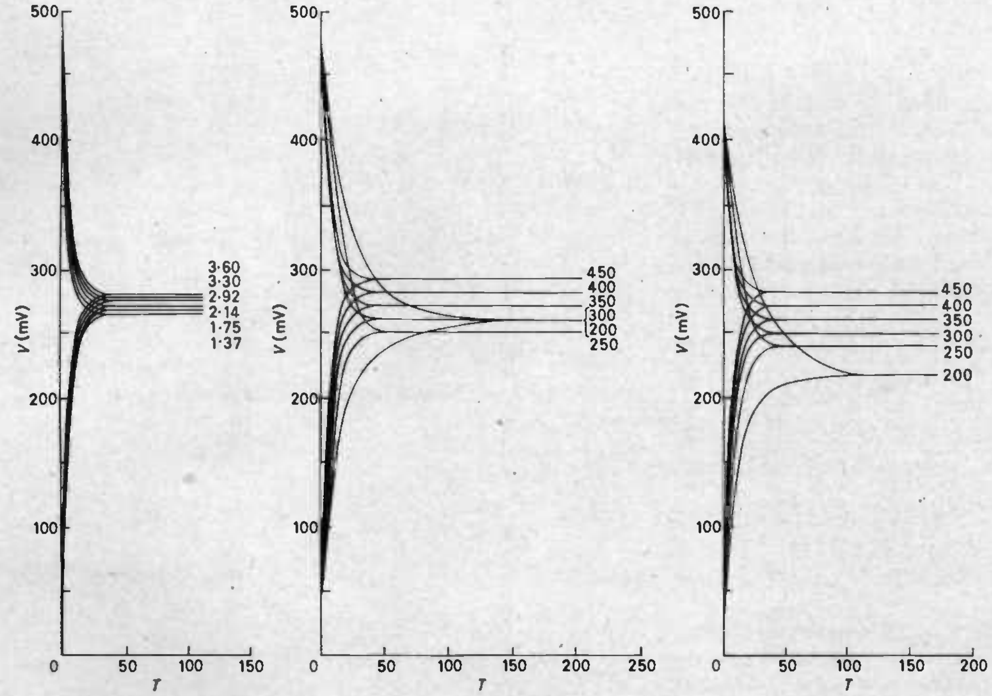



**Fig. 8. Calculated output response to unit step drive**
*(a) parameter : I, (b) parameter : $V_{max}'$, (c) parameter : $V_{max}'$*

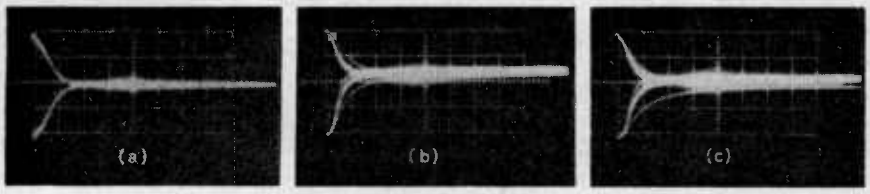



**Fig. 9. Experimental output response to unit step drive**
*(a) V : 100mV/Div. H : 10T/Div., (b) V : 100mV/Div. H : 20T/Div.,*
*(c) V : 100mV/Div. H : 20T/Div.*

From these results, the rise-time is a function of $I$, $V_{max}'$ and $R$, and is regarded as less than $50T$ for the ordinary condition ($V_{max}' \geqslant 300$mV, $R > 200\Omega$).

CHANGE OF $I_s$

The characteristic curve of $V'$ against $v$ is qualitatively explained by Fig. 12 when $I$ is nearly equal to $I_s$. Now in this figure, consider the process in which $V'$ decreases from $V_{max}'$ (point $M$) to zero. When $dV'/dt$ is nearly zero, namely $V'$ decreases quasi-statically to zero, the operation point changes to $v_{01}$ along the line (3).

If $dV'/dt \ll 0$, the capacitance $C (= C_0 + C_1 + C_2)$ in Fig. 6 is inclined to sustain the value of $v$, therefore the operating point moves along the dotted line different from the line (3).

If the operating point crosses the line (2) ($I = I_s$) before it reaches the memory region, the final value of $v$ becomes $v_{03}$. In other words, the stored charge in $C$ serves as a loop current, and the resultant value of $I_s$ seems to be dropped.

The above conclusions are now confirmed by observation and computer analysis for the case $ED_1 = ED_2 =$ SONY type-1T1101.

The $V' - v$ characteristic curves in the interval between point $M$ and the memory region are approximated by parallel straight lines as for $I$ (as illustrated in a previous letter[3] for $V' \leqslant 250$mV). In this region the drop of $I_s$ might be proportional to capacitance $C$ for the same $dV'/dt$.

Fig. 13 shows the observed results of $I_s$ against $dV'/dt$ and $C$, where the abscissa is the added capacitance $C'$ outside the circuit, and the parameter is the frequency of the sine wave used as $V'$(Hz).

In every condition the observed points are fitted to a straight line within the limits of error, and their extrapolated lines coincide at the point $I_s = I_{s0}$ and $C = C_s$. $I_{s0}$ is the value of $I_s$ for a quasi-static change of $V'$, and $C_s$ is the capacitance of the circuit itself. This fact supports the aforementioned.

The change of $I_s$ is normalized to $T (= t/C)$, for it satisfies equation (11). The observed relations between $I_s$ and the frequency of the sine wave ($V'$ in Hz) are given in Fig. 14 with parameter of $C (= C_s + C'$ in pF). It is seen from the figure that the intervals between their curves are expressed by the ratios of the parameters (indicated by "$\longleftrightarrow$"), further it means that $T$ in equation (11) is indeed the ratio of actual time to capacitance. Fig. 15 shows the observed relation between the frequency of $V'$ (normalized to $T$) and the deviation rate $I_s' (= [(I_{s0} - I_s)/(I_{s0} - I_{min})] \times 100)$, which is in agreement with the calculated results.

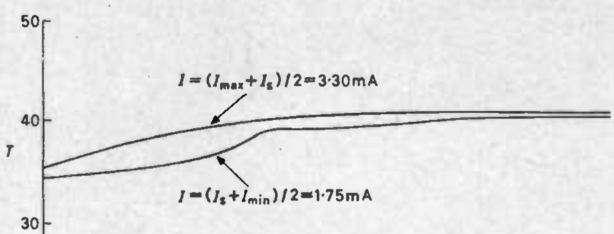
Fig. 16 shows the relation of $I_s$ to $R$ with parameter $V_{max}'$, where $1/C.f$ is $3{\cdot}3 \times 10^3$ and $I_{s0}$ and $I_{min}$ refers to Fig. 7. The figure illustrates that the deviation rate $I_s'$ is constant, independent of $R$ ($250 \leqslant R \leqslant 80$k$\Omega$), within the limits of error.

CORRECTION FOR THE CHANGE OF $I_s$

In the preceding section, the change of $I_s$ caused by $dV'/dt$ was discussed, and various factors were clarified, which now may be corrected for.

Letting $C_1 \neq C_3$, equation (11) becomes:

**Fig. 10. Characteristic curve between load and rise-time (parameter : I)**



$$dv/dT = i - \{G(v - V') + G(v + V') + v/R\} - \frac{C_2 - C_1}{C_0 + C_1 + C_2} \cdot dV'/dT \quad \ldots\ldots\ldots \quad (12)$$

The response of $v$ can be controlled by making the coefficient of the last term in equation (12) to have a proper value, and it is applied to the correction for the change of $I_s$.

From a qualitative viewpoint, so far as the $V' - v$ characteristic curves are parallel straight lines, the equivalent circuit in Fig. 6 is rewritten as shown in Fig. 17. From this circuit, the quasi-static $V' - v$ characteristics are:

$$\partial v/\partial V' = (1 - R_1/R_2)/\{(1 + R_1/R_2) + R_1/R\} \quad \ldots\ldots \quad (13)$$

For the case $ED_1 = ED_2 =$ SONY type-1T1101 and $R = 1\text{k}\Omega$, $\partial v/\partial V'$ is less than unity, then:

$$R_1 \ll R_2 \text{ and } R_1 \leqslant R \quad \ldots\ldots\ldots\ldots \quad (14)$$

The selection of $C_1$, $C_2$ and $C_0$ so as to satisfy the following condition:

$$C_1 . R_1 = C_2 . R_2 = C_0 . R \quad \ldots\ldots\ldots \quad (15)$$

yield the faithful response of $V'$ to $v$ in Fig. 17; which is then experimentally confirmed in consideration of equations (14) and (15).

Fig. 18 shows some examples of experimental results, and corresponds to Fig. 14, where $I_s$ is constant independently of $dV'/dt$ for certain combinations of $C_0$, $C_1$, and $C_2$, (3). This demonstrates the correctness of the representation for the change of $I_s$ and for its correction.

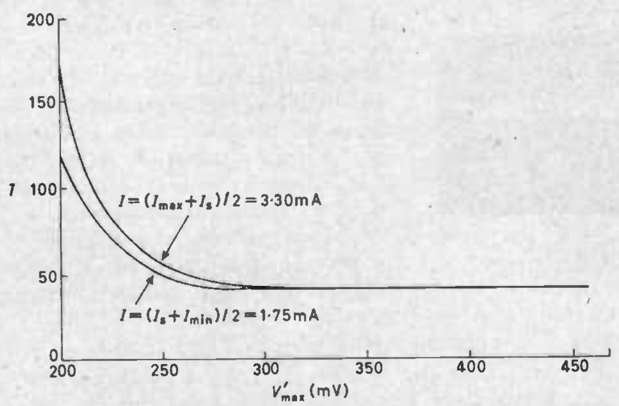
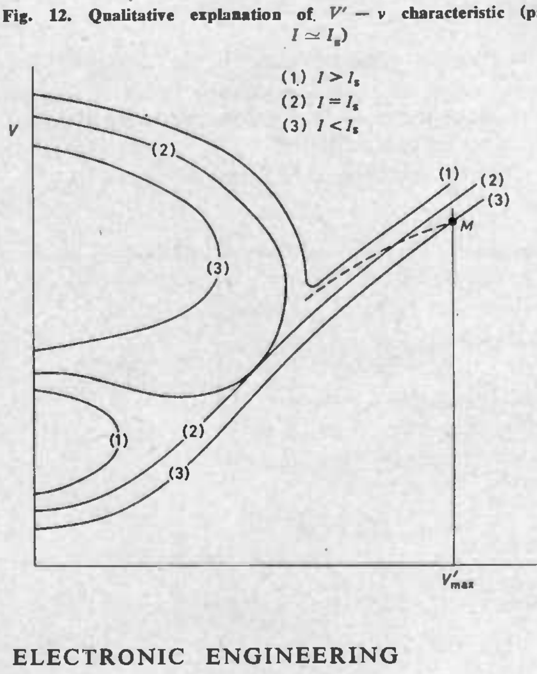



**Fig. 11.** Characteristic curve between $V_{max}'$ and rise-time (parameter : $I$)

**Fig. 12.** Qualitative explanation of $V' - v$ characteristic (parameter : $I$, $I \simeq I_s$)



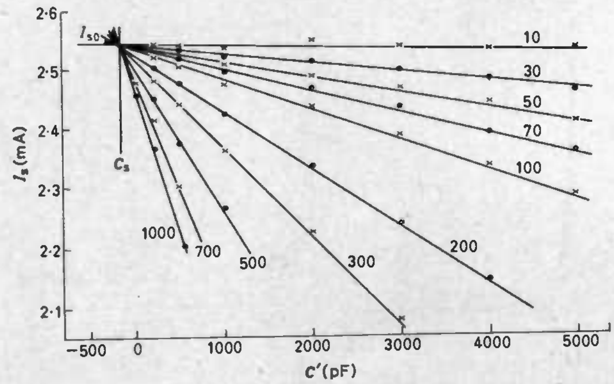




**Fig. 13.** $C' - I_s$ characteristic curve with the parameter of drive frequency

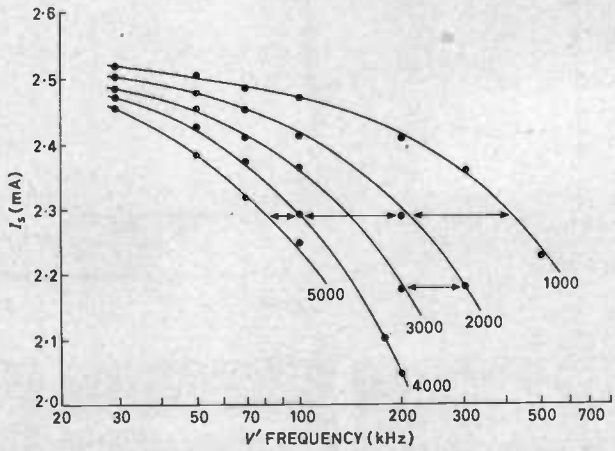



**Fig. 14.** Characteristic curve between drive frequency and $I_s$ (parameter : $C$)

## Conclusions

In the previous sections the operation of the f.c.e.d. circuit under various conditions was investigated through experiments and computer analysis, and these yielded the following conclusions:

(1) The f.c.e.d. circuit operates as a current switch as well as a memory.

(2) The variety of the incident current range for memory and threshold level are obtained in accordance with the combination of diodes.

(3) The limitation of load is decided by the ratio of output levels. In the ordinary range, a considerable amount of fan-in and fan-out is obtained (for diodes of SONY type-1T-1101, $R \geqslant 200\Omega$)

(4) The rise time for switch action is very short (for SONY type-1T1101, the output response to unit step drive is less than $50T (= t/C)$; e.g. 1nsec for a stray capacitance of 20pF).

(5) A change in $I_s$ is caused by $dV'/dt$ (for SONY type-1T1101, a drive of sine wave at period $5 \times 10^3 T$ causes a 10 per cent change in $L_s$; e.g., 10MHz for a stray capacitance of 20pF).

(6) A change in $I_s$ is avoided by the use of the equilibrium of capacitances.

Consequently, it is considered that the f.c.e.d. circuit is suitable for memory and switching elements of high speed.


#### Acknowledgments

The author wishes to express his thanks to Prof. E. Takeda, Prof. N. Yamamuro, Prof. K. Yanagisawa and Prof. M. Toma of Tokyo Institute of Technology for guidance in the course of the work.

He is indebted to Mr. Y. Chiba of the Institute of

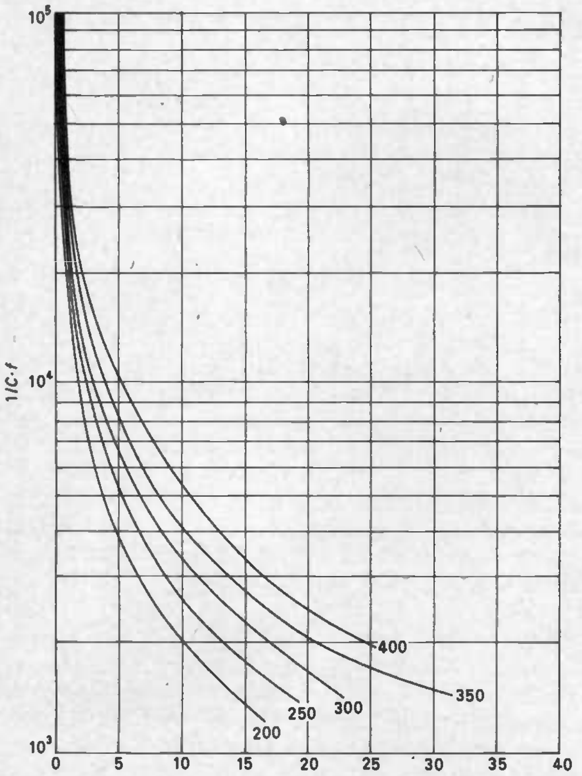



Fig. 15. Characteristic curve between deviation rate of $I_s$ and drive frequency (parameter : $V_{max}'$)

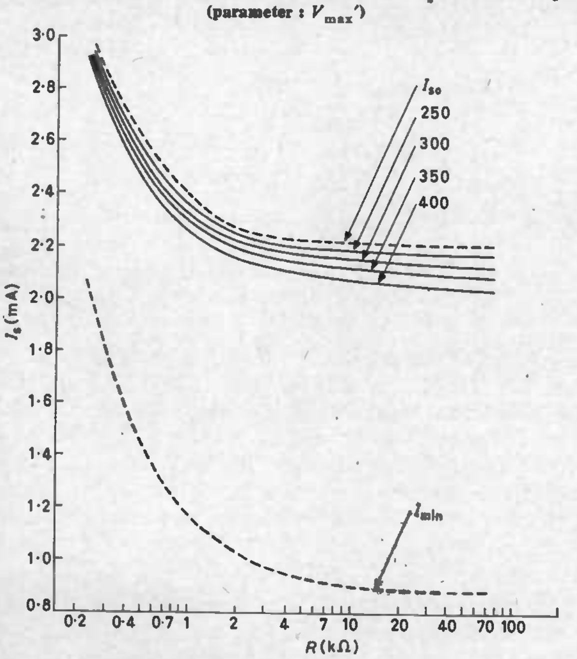
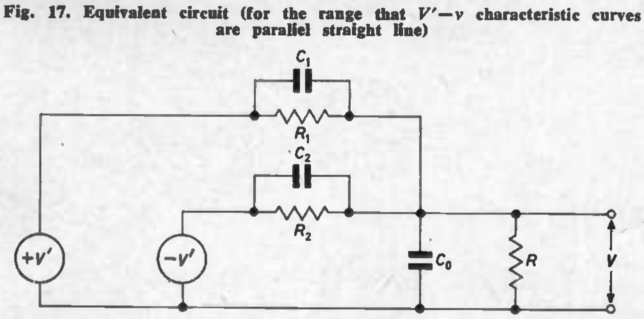



Fig. 16. Characteristic curve between load and $I_s$ (parameter: $V_{max}'$)

Fig. 17. Equivalent circuit (for the range that $V'-v$ characteristic curves are parallel straight line)



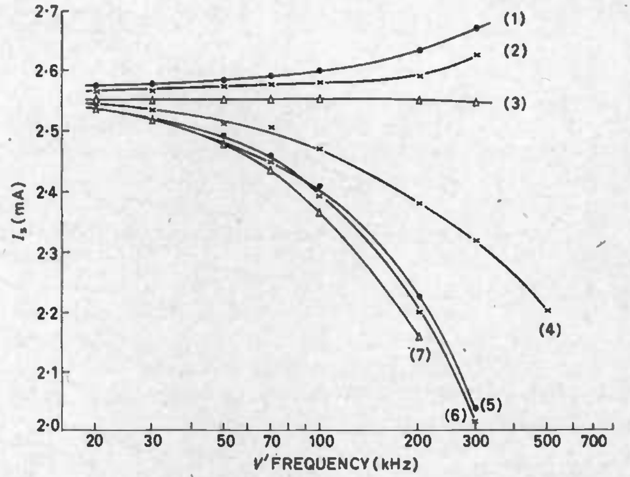




Fig. 18. Characteristic curve between drive frequency and $I_s$ (parameter: $C$)

Physical and Chemical Research and Dr. K. Kanda of Research Reactor Institute of Kyoto University for valuable discussions.

He wishes to acknowledge his gratitude to Professor Dr. Y. Toriyama of his department for constant support.

## Appendix

Characteristics of SONY type-1T1101 and SONY type-1T1104

| | | SONY type-1T1101 | SONY type-1T1104 |
|---|---|---|---|
| $I_{F(max)}$ (mA) | | 40 | 50 |
| $I_{R(max)}$ (mA) | | 50 | 60 |
| $P_{D(max)}$ (mW) | | 25 | 25 |
| $T_s$ (°C) | min. | −55 | −55 |
| | max. | 100 | 85 |
| $T_j$ (°C) | min. | −55 | −55 |
| | max. | 100 | 85 |
| $I_p$ (mA) | min. | 1.95 | 5* |
| | typ. | 2·0 | 6 |
| | max. | 2·05 | 7 |
| $I_p/I_v$ | min. | 7 | 4·5 |
| | typ. | 8 | |
| $V_p$ typ. (mV) | | 70 | 70 |
| $V_v$ typ. (mV) | | 340 | 340 |
| $V_f$ typ. (mV) | | 480 | 480 |
| $-R$ typ. (Ω) | | 60 | 25 |
| $C$ (pF) | typ. | 6 | 15 |
| | max. | 10 | 30 |
| $R_s$ (Ω) | typ. | 1·5 | 0·8 |
| | max. | 2·0 | 1·5 |
| $L_s$ typ. (nH) | | 0·4 | 0·4 |
| $f_o$ typ. (GHz) | | 3 | 2 |
| $f_r$ typ. (GHz) | | 3 | 2 |

* *For the value of $I_p$, 5mA is employed in the calculations.*

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# A Junction Transistor Electrometer Circuit

By T. K. Cowell*, B.Sc.(Eng.), C.Eng., M.I.E.E.

*The maximum input resistance obtainable with conventional junction transistor circuits is limited by collector/base leakage resistance and reduced effective current gain at the necessarily low operating currents.*

*The basic circuit configuration described—a collector guard-ringed complementary emitter-follower pair—overcomes these limitations, and with cascaded stages yields input resistances in excess of one million megohms.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

TWO development projects with which the author has been concerned recently—a battery operated pH meter[1] and an analogue (voltage) store—have involved a requirement for a circuit possessing a very high input resistance. For the pH meter, the circuit required consisted essentially of a d.c. voltmeter having a full-scale sensitivity of slightly less than one volt. In the case of the store (consisting of a capacitor and unity gain buffer amplifier) it was required to store voltages in the range of $\mp 10$V with respect to earth, with a discharge time-constant of not less than about twenty minutes. The particular kind of pH electrode used in the first application, and the value of storage capacitor employed in the second both dictated a minimum value of input resistance of about $5 \times 10^9 \Omega$ for the associated amplifier.

The obvious advantages to be obtained by using semiconductors rather than thermionic valves, and (at the time of writing) the relatively high price of m.o.s.t. and field effect devices prompted the author to investigate the use of a circuit based upon cascaded emitter-followers.

The lack of difficulty with which the performance requirements outlined above were achieved suggested that considerably higher values of input resistance might be obtained, and subsequently led to the design of a circuit possessing an input resistance in excess of one million megohms.

### Cascaded Emitter Followers

Before discussing circuit arrangements, it will be worth while noting certain features of the basic emitter-follower circuit which are germane to the arguments following.

In Fig. 1(a) a reasonable approximation to the value for $R_{in}$ (assuming large signal conditions) would be:

$$R_{in} = V_{in}/I_{in} = r_b + (1 + \beta) R_e \quad ...... \quad (1)$$

By considering this input resistance as the emitter load for a preceding stage, and so on for several stages, it can be seen that theoretically at least, a resultant input resistance several orders of magnitude greater than $R_e$ would be obtained. Such a circuit arrangement for three stages is illustrated in Fig. 3(a).

However, returning to the single stage, the simplified equivalent circuit of Fig. 1(b) suggests a somewhat lower value of input resistance than that given by equation (1), due to the presence of collector resistance $r_c$ effectively in parallel with $R_e$. Furthermore, in practice, transistor collector-base leakage resistance acts to reduce the input resistance still further, and leakage current flowing between collector and base may result in failure of the circuit to operate properly under some conditions.

Before discussing circuit arrangements overcoming these problems, something must be said about the choice of an appropriate type of transistor. Clearly, the early stages of the cascade arrangement mentioned above will operate at very small collector currents: it follows that the transistors used must have a useful current gain under such conditions, and that leakage currents should be very small. Such requirements suggest the silicon planar epitaxial transistor as the most suitable, and experimental work was confined to this type.

### Transistor Current Gain at Ultra-Low Currents

Because of the lack of manufacturer's data about the operation of transistors at collector currents below about 10$\mu$A, a simple arrangement was devised for measuring effective current gains over a range of collector currents extending well below this value—down to about 20pA in fact. The method is described in Appendix (1).

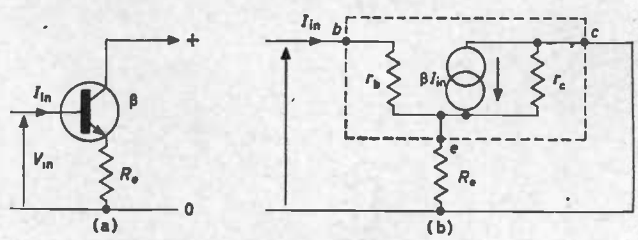


**Fig. 1(a). Basic emitter follower**
**1(b). Simplified equivalent circuit of Fig. 1(a)**

Fig. 2 shows typical current gain/emitter current relations for Texas Instruments types 2S502 and 2N3702 transistors measured using the technique outlined.

Fig. 3 shows a basic circuit arrangement theoretically yielding an input resistance of the order required. Ignoring for the present the effects of collector-base leakage, an approximate value for input resistance can be calculated assuming that the curve of Fig. 2 for the 2S502 applies to the transistors.

Assuming for convenience an input of 10V, and ignoring the emitter-base voltages, so that $V_{in} = V_o$,

Then for $R_e = 20\text{k}\Omega$

$$I_4 = \frac{10}{2 . 10^4} = 0{\cdot}5\text{mA}$$

$$I_3 = \frac{0{\cdot}5}{1 + \beta_3}\text{ mA} = 2{\cdot}4\mu\text{A} \quad (\beta_3 = 205)$$

$$I_2 = \frac{2{\cdot}4}{1 + \beta_2}\mu\text{A} = 22\text{nA} \quad (\beta_2 = 108)$$

$$I_1 = \frac{22}{1 + \beta_1}\text{ nA} = 580\text{pA} \quad (\beta_1 = 37)$$

$$R_{in} = V_{in}/I_1 \; \frac{10}{5{\cdot}8 . 10^{-10}} = 17\,000\text{M}\Omega$$

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## Improved Circuit Arrangement

In practice, the arrangement of Fig. 3 would have limited use; the circuit would operate only for positive input signals, and assuming a nominal half-volt for each emitter-base junction voltage, for inputs appreciably greater than about 1½V. Moreover, the output voltage would be offset from the input by the three emitter-base junction voltages—an appreciably temperature dependent quantity.

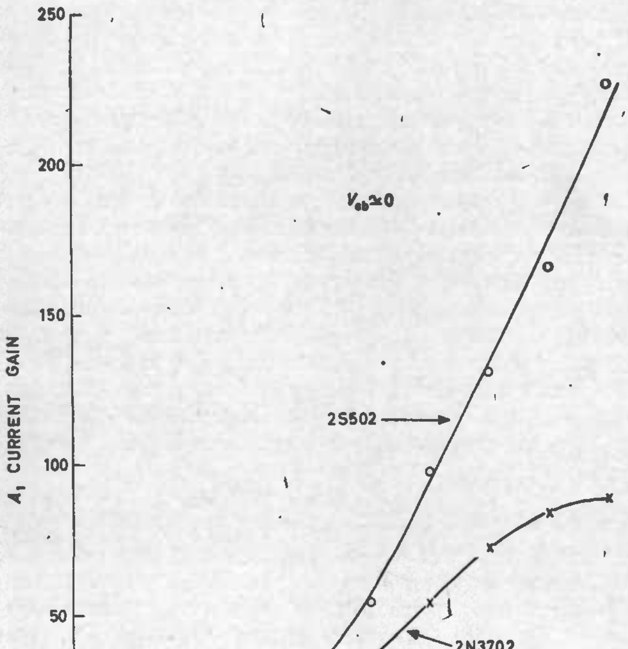
A better arrangement, illustrated in Fig. 4(a) is obtained by adopting a two-battery circuit and introducing complementary emitter-follower stages. This results in first order cancellation of the temperature dependent offset voltage between input and output, and also overcomes the previous restrictions on signal amplitude and polarity.

Additionally, the arrangement leads to a means for effecting an important improvement in performance.

This is achieved by connecting the collectors of the npn transistors to the final stage emitter, which since the circuit has virtually unity gain, means that the first stage collector behaves as a guard ring with respect to the base electrode. Such a circuit arrangement has the following advantages. First, and most important, collector-base leakage current in the first stage is virtually eliminated, and second, a higher effective current gain is realized since the shunting effect of $r_o$ on the effective emitter load resistance is eliminated (see Fig. 1(b)). A further incidental advantage obtained is that the limitation on input signal amplitude imposed by the maximum collector-base voltage rating of the npn transistors is avoided. This is in practice quite important, as transistors having useful current gain at ultra-low collector currents do not in general have high collector-base voltage ratings.

The basic two-battery circuit does, however, result in an

**Fig. 2. Current gain versus emitter current for Texas Instruments types 2S502 and 2N3702 transistors**



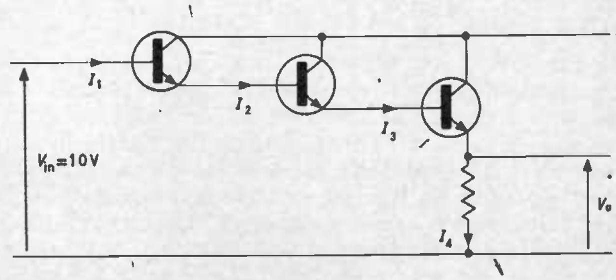




**Fig. 3. Simple cascaded emitter follower circuit**

input offset current, since with zero signal voltage, there are still currents flowing in the transistors. To overcome this, a means to supply the first stage standing base current must be introduced. Fig. 4(b) shows the basic two-battery circuit modified for guard-ring operation, and including arrangements for supplying the first stage base bias current.

## Design Procedure

To see what is involved in designing a circuit of the type shown in Fig. 4(b) for some specific application, it is convenient to consider the circuit in a more generalized form illustrated in Fig. 5. Here, box $A_1$ may represent either a single npn transistor or two or more in cascade, having an overall effective current gain of $\beta_1$: similarly for box $A_2$, where the transistor or transistors are pnp type, effective overall gain $\beta_2$.

The design task can now be stated in terms of this generalized arrangement, and amounts essentially to the following: given values for load resistance $R_L$, required input resistance $R_{in}$ and supply and maximum signal voltages $\mp V_B$ and $\mp V_S$ respectively, it is required to determine values for $R_o$, $R_1$ and $R_2$, and the current gains $\beta_1$ and $\beta_2$. (Strictly, the supply voltages may be considered as independent variables, but are here assumed to be fixed by other considerations.)

As is frequently the case with transistor circuits, a rigorous approach to circuit synthesis involves quite lengthy calculations. However, by making some simplifying assumptions, equations may be obtained expressing the required values as simple functions of supply and maximum signal voltages, and load and required input resistances. Because of the dependence of transistor current gains on operating currents, it is necessary also to estimate the latter in establishing the combination of transistors required to achieve any particular overall current gain, $\beta_1\beta_2$. Formulae for $R_o$, $R_1$, $R_2$, $\beta_1$, $\beta_2$ and the emitter currents are derived in Appendix (2).

It should be noted, however, that a circuit designed strictly in accordance with the formulae would not function properly, since the expressions for $R_2$ and $R_1$ imply zero emitter currents for the associated transistors at respectively maximum positive and negative signal amplitudes. (In the case of $R_1$, the situation is worse, since the voltage across this component is less than that assumed in the derivation by the sum of the emitter-base potentials of the pnp or npn transistors).

Satisfactory performance may be obtained by making appropriate allowance for the above factors; for example by designing for a maximum signal voltage in excess of that actually required. Although this represents a somewhat empirical approach, the author feels that the tedium of a rigorous design would seldom be justified—particularly in view of the very limited control the designer has over the overall current gain of whatever combination of transistors is used in any particular application.

## Practical Circuits

In Fig. 6 is shown the circuit of a simple experimental high impedance probe for an oscilloscope. Employing two transistors only, the arrangement has a measured input resistance of 1 000MΩ, and operates for signals in the range ∓5V. For simplicity, a fixed resistor supplies the first stage base bias current, so there is a residual input offset current of a few nanoamperes. The small input/output voltage offset of approximately 90mV due to the difference in emitter-base voltage of the two transistors is accommodated by the oscilloscope shift control.

Fig. 7 illustrates a practical 1-0-1 voltmeter embodying the circuit arrangements so far discussed, similar to that used for the pH meter application mentioned previously. In this circuit, the need for two separate batteries is avoided by the use of a potential divider to provide an 'earth' line. The measured input resistance of the circuit is approximately 14 000MΩ. Preliminary adjustments involve first setting 'voltage-zero' with the input short-circuit, then adjusting the 'current-zero' with the input open-circuit.

## Improving Performance

In the circuit of Fig. 7, because the latter (pnp) stages operate at higher currents than the first and second stages, the emitter-base potentials of $VT_3$ and $VT_4$ are in general higher than those for $VT_1$ and $VT_2$. This results in the 'output' terminal voltage (i.e. the potential at the emitter of $VT_4$) being offset from the input by a small positive amount (200mV or so). This is of no consequence in the circuit illustrated, since it is automatically compensated for in the zero-setting operation. For some applications, however, it is a disadvantage: for example, in a buffer amplifier application where the input voltage must be reproduced at the output terminal relative to earth. It is also of consequence in circuits using a greater number of transistors to obtain substantially higher values of input resistance. This is because the offset voltage is also the collector-base potential of the first stage, and results in residual collector-base leakage current in that stage.

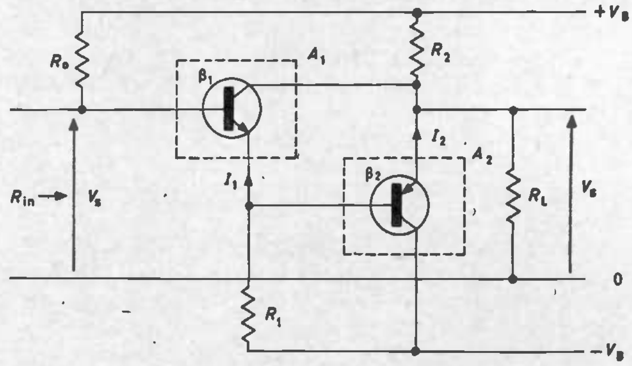



**Fig. 5. Generalized circuit arrangement**

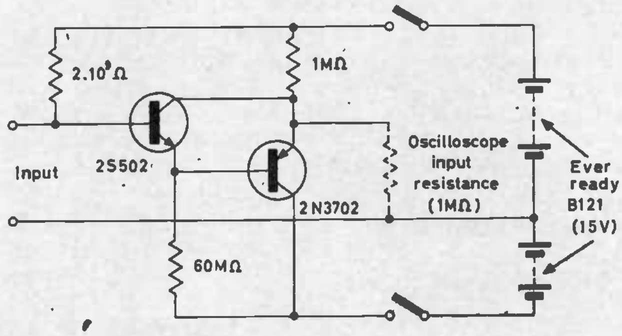
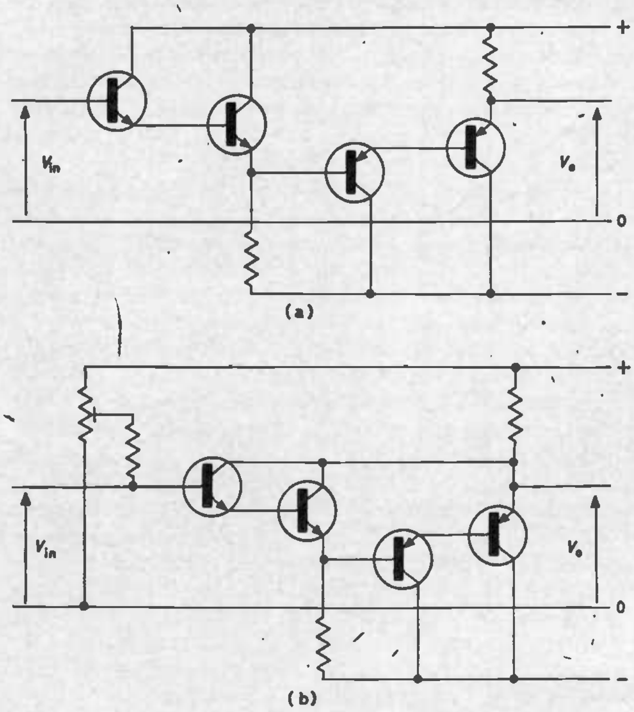



**Fig. 6. Experimental oscilloscope probe unit with input resistance of 1 000MΩ**

Although the offset may be reduced by selecting transistors, this is not in general a practicable solution. The use of a potential divider between output and negative line may be acceptable in some applications, but results in increased output resistance and a reduction of overall (no-load) voltage gain. This latter is significant in applications where very high input resistance is required, since with a voltage gain appreciably less than unity, effectively zero first stage collector-base potential will not be maintained as the input voltage varies.

Fig. 8 illustrates in basic form an arrangement overcoming the foregoing difficulties. The potentiometer provides at its wiper a voltage adjustable between that of the emitters of the last and last-but-one stages. This permits zero input/output offset to be obtained without introducing a reduction in voltage gain. In practice a value of potentiometer resistance high enough not to cause excessive emitter current in the penultimate stage raises the circuit output resistance appreciably, and it is desirable to have an emitter-follower stage following the potentiometer.

**Fig. 4(a) Cascade circuit using complementary stages**
**Fig. 4(b) Guard-ring connexion for collectors of npn stages**




## Million Megohm Buffer Amplifier

Fig. 9 illustrates a practical circuit incorporating the above arrangement, which because of the net offset voltage introduced by the first six stages, requires two additional npn stages in order that zero input/output offset may be obtained. Setting-up the circuit is essentially the same as in the previous case; adjustment of $RV_2$ for zero input/output offset with the input shorted, then setting $RV_1$ with the input open-circuit. Because of the difficulty involved in screening the very high impedance input stages, it was found helpful in practice to temporarily connect a small capacitor (100pF or so) between input and ground terminals, and then to adjust $RV_1$ for zero rate of change of output.

The measured input resistance of the circuit was found to be not less than $1{\cdot}4 \times 10^{12}\Omega$ for signals in the range $\mp 10$V; the open-circuit voltage gain was approximately 0·97.

**Discussion**

The measured input resistance of experimental circuits constructed by the author have been found to be somewhat higher than that predicted using the approximate design formulae derived and the transistor data of Fig. 2. This is due to the npn transistors having higher effective current gains than determined by the method of measurement of Appendix (1), where the transistor collector is not guard-ring connected with respect to the base, as is the case in practice.

Circuits of the type described are clearly more complex than those using field effect or m.o.s.t. devices to obtain a high input impedance. However, in some cases they may offer certain operating advantages. In the author's experience, such arrangements are less susceptible to damage by transient high voltage inputs than are m.o.s. transistors. Furthermore, although a full investigation has not been made, results so far indicate that the effect of temperature variation on input offset current is less than that for a field effect device.

One point which came to light during the experimental work was that for the transistors used, the temperature coefficient of emitter-base voltage increased appreciably at very low emitter currents. In the case of 2S502 it rose from typically $-2{\cdot}5$mV/°C at $I_e = 100\mu$A, to $-3{\cdot}4$mV/°C at $I_e = 100$pA (at 20°C). The author has not seen this effect mentioned in the literature previously.

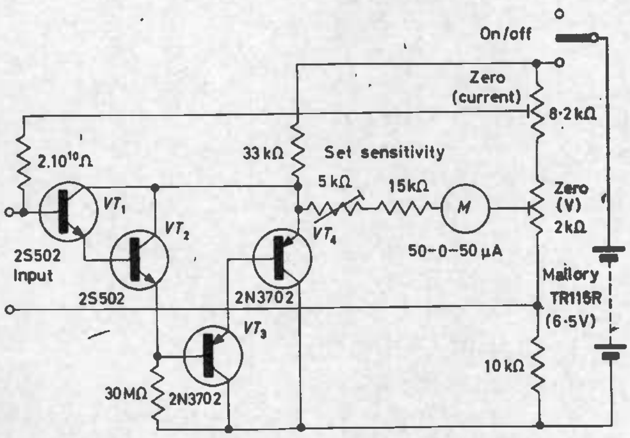



Fig. 7. Practical 1-0-1 voltmeter with input resistance of 14 000MΩ

As a result of this variation in $\Delta V_{eb}/\Delta T$ it is necessary to use a greater number of pnp stages than npn stages if good temperature compensation is important. For example, by adding a further pnp stage to the circuit of Fig. 8 and suitably adjusting the potentiometer, effectively zero overall temperature coefficient of input/output voltage may be obtained. An experimental circuit using this technique constructed by the author was readily adjustable for a drift of less than $30\mu$V per hour under average laboratory conditions. It is not possible, however, simultaneously to obtain zero input/output voltage offset and zero overall temperature coefficient with the circuit arrangements described.

Although the maximum value of input resistance so far obtained is somewhat lower than that for more conventional electrometer circuits, it is adequate in many applications where these would otherwise be used. In this connexion, it is perhaps worth observing that a voltage store circuit comprising a (perfect) $1\mu$F capacitor and the buffer circuit of Fig. 9 would have a discharge time-constant of about a fortnight.

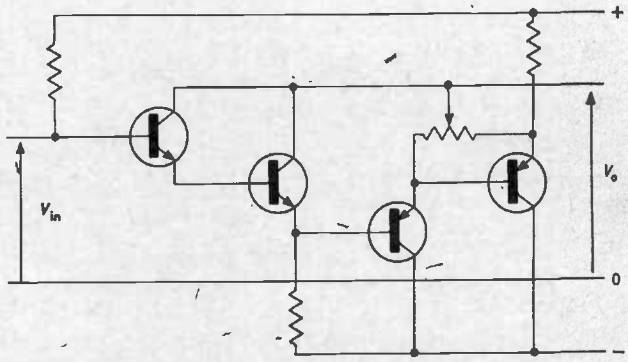



Fig. 8. Arrangement to obtain zero input/output offset voltage

## Appendix

### (1) Transistor Tests

This procedure was devised for evaluating the effective large signal current gain:

$$A_l = I_e/I_b \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

for transistors in grounded collector connexion. $A_l$ is determined for values of $I_e$ in the range $10^{-3}$ to $2 \,.\, 10^{-11}$A. Measurements are made under the condition $V_{ob} \simeq 0$, to reduce errors due to collector-base leakage current.

*Theory*

Fig. 10 shows the circuit arrangement. With $S$ initially closed:

$$I_e = \frac{E - V_{ce}}{R} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

With $S$ opened $I_b$ flows from the capacitor, and in time $\Delta t$ seconds, the capacitor voltage will change from zero to $\Delta V$ volts. If the value of $C$ is chosen so that $\Delta V$ is small compared with $(E - V_{ce})$, the emitter current will remain sensibly constant during $\Delta t$. Thus, $I_b$ and $V_{be}$ will also remain constant during this period, and:

$$I_b = C\,(\Delta V/\Delta t) = C\,(\Delta V_e/\Delta t) \quad \ldots\ldots \quad (4)$$

From equations (2), (3) and (4):

$$A_l = \frac{E - V_{ce}}{CR} \,.\, (\Delta t/\Delta V_e) \quad \ldots\ldots\ldots\ldots \quad (5)$$

*Procedure*

Insert value of $R$ such that with $S$ closed, approximate required emitter current flows (equation (3)). Back off d.v.m. to zero using $RV$. Check $V_{ce}$, adjust $E$ for correct $I_e$ (equation (3)).

Estimate $A_l$, select $C$ to give $\Delta t$ between say 20 and 60 seconds for $\Delta V_e = 10$mV (equation (5)).

Open $S$ measure $\Delta t$ with stop watch for $\Delta V_e = 10$mV, and calculate $A_l$ (equation (5)).

*Example: test on type 2S502*

Set $E = 10$V. For $I_e = 10^{-9}$A, set $R = 10^{10}\Omega$. Check $V_{ce}$ (380mV), trim $E$ to 10·38V. Assuming $A_l = 30$, use $C = 0{\cdot}1\mu$F. Obtain $\Delta t = 45$sec for $\Delta V_e = 10$mV. Calculate $A_l = 45$.

By choice of appropriate values of $C$ and $R$, values for $A_l$ were determined for emitter currents in the range $10^{-4}$ to $2.10^{-11}$A (see Fig 2) for Texas Instruments types 2S502 and 2N3702 transistors.

Polystyrene capacitors were used for values of $C$ of $0{\cdot}1\mu$F and smaller.

### (2) Derivation of Formulae

With reference to Fig. 5, to simplify the working and the derived relationships, the following assumptions are made:

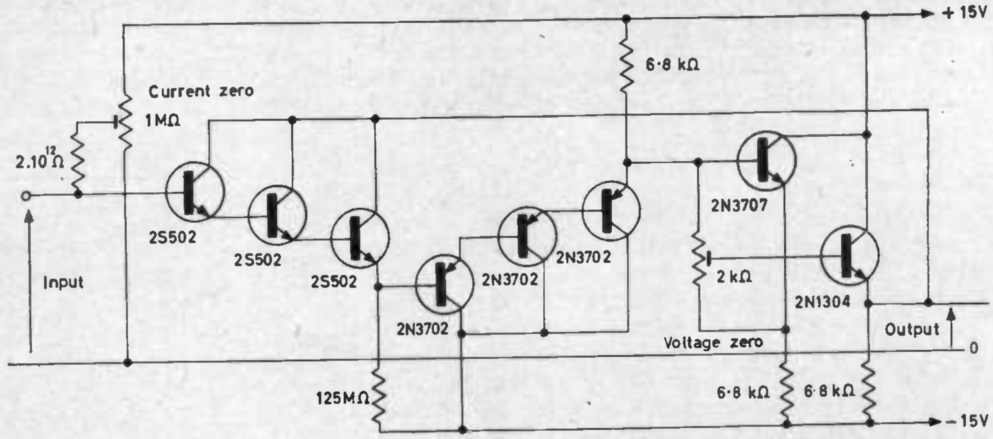



Fig. 9. Experimental million-megohm buffer amplifier

(a) Each transistor or cascaded combination may be treated as an ideal current amplifier having zero input resistance.

(b) The positive and negative supply voltages ($V_B$) are equal.

(c) The maximum positive and negative signal voltage excursions ($V_s$) are equal in magnitude.

(d) Collector current of the npn transistor(s) is always negligible compared with the total current flowing in $R_2$.

(e) The transistor emitter/base potential differences may be neglected in calculations involving the d.c. conditions (i.e. the sum of the voltages across $R_o$ and $R_1$ or $R_2$ and $R_1$ is assumed equal to the sum of the supply voltages).

(f) The circuit has overall unity gain.

(g) Leakage currents are negligible.

(h) Transistor current gains do not vary with signal voltage.

(i) Each stage has an input resistance equal to $\beta$ times the effective common load resistance.

Note: To avoid confusion with signs, all voltages appearing in the equations are regarded as positive quantities.

*Value for $R_2$*

Limiting (maximum) value for $R_2$ is defined by d.c. conditions when signal voltage is maximum positive, and emitter current $I_2$ tends to zero, i.e.:

$$\frac{V_B - V_s}{R_2} = V_s/R_L$$

from which:

$$R_2 = \left(\frac{V_B - V_s}{V_s}\right) R_L \quad \ldots\ldots\ldots\ldots (6)$$

*Value for $R_1$*

Limiting (maximum) value for $R_1$ is defined by d.c. conditions when signal is maximum negative, and emitter current $I_1$ tends to zero, i.e.:

$$\frac{V_B - V_s}{V_s} = 1/\beta_2\left(V_s/R_L + \frac{V_B + V_s}{R_2}\right)$$

which yields, after eliminating $R_2$ by substituting from equation (6):

$$R_1 = \frac{(V_B - V_s)^2}{2V_BV_s} \,.\, \beta_2 R_L \quad \ldots\ldots\ldots (7)$$

*Value for $R_o$*

Correct value for $R_o$ is such that it will supply the standing base current required by the first transistor under zero signal conditions, i.e.:

$$V_B/R_o = 1/\beta_1\left((V_B/R_1) - (1/B_2)\,.\,(V_B/R_2)\right)$$

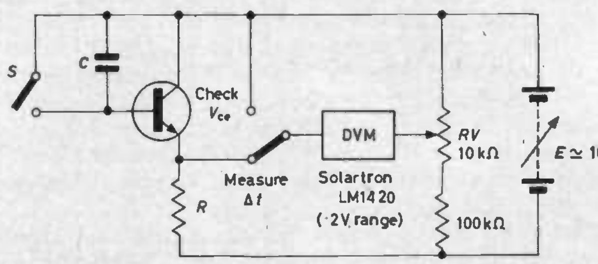



Fig. 10. Circuit for transistor tests

which yields, after eliminating $R_2$ and $R_1$ by substitution from equations (6) and (7):

$$R_o = \frac{(V_B - V_s)^2}{V_s(V_B + V_s)} \,.\, \beta_1\beta_2 R_L \quad \ldots\ldots\ldots (8)$$

*Required Current Gain*

The total current gain required is defined by the need to achieve the required input resistance, $R_{in}$. Writing $G_{in} = 1/R_{in}$, $G_o = 1/R_o$, etc.:

$$G_{in} = G_o + (1/\beta_1)\,G_1 + (1/\beta_1\beta_2)\,(G_2 + G_L)$$

Inverting equations (6), (7), and (8) to obtain expressions for $G_o$, $G_1$ and $G_2$ and substituting into the above yields:

$$G_{in} = \left(\frac{V_B + V_s}{V_B - V_s}\right)^2 G_L/\beta_1\beta_2$$

from which:

$$\beta_1\beta_2 = R_{in}/R_L\left(\frac{V_B + V_s}{V_B - V_s}\right)^2 \quad \ldots\ldots\ldots\ldots (9)$$

*Standing Emitter Currents*

For the pnp transistor (or the final transistor of a cascaded combination), when $V_s = 0$:

$$I_2 = I_{2s} = V_B/R_2 \quad \ldots\ldots\ldots\ldots (10)$$

Similarly for the npn transistor (or the final transistor of a cascaded combination):

$$I_1 = I_{1s} = V_B/R_1 - (1/\beta_2)\,I_{2s}$$
$$= V_B\left((1/R_1) - (1/\beta_2R_2)\right) \quad \ldots\ldots (11)$$

Standing emitter currents for transistors other than the final one of a cascaded cimbination may be obtained directly knowing the standing current and current gain of the following transistor.


**Acknowledgment**

The author would like to acknowledge assistance with the experimental work given by members of the staff of the Medical Electronics Department, St. Thomas' Hospital.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# Solar Noise as a Means of Providing Accurate Vertical Polar Diagrams for Radars

**(Part 2)**

By M. J. B. Scanlan*, B.Sc., A.R.C.S.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 342)

IN Part 1 of this article, it was shown how the sun can be used as a signal source to measure the vertical polar diagrams of aerials (especially surveillance radar aerials) over a wide range of gain and with good relative accuracy. However, the sun as a source has several disadvantages, as follows:

(1) It is too large (its angular diameter is about 30′ of arc) so that sharp variations in the v.p.d. will be somewhat smoothed. For comparison, an aircraft taken as 100ft, subtends something like 40″ of arc at 100 miles. Hence there may be a gap in the v.p.d. not shown by sun measurement, but large enough to lose an aircraft. This is especially true at higher frequencies, or with aerials mounted high (in terms of wavelengths) above a reflecting surface which they illuminate.

(2) The brightness distribution across the disk is non-uniform and variable. This could be troublesome if the aerial beamwidth is of the same order as the sun's angular diameter.

(3) The effective disk temperature is not known, so that absolute gain measurements are only possible using a gain horn (see Fig. 6). Use of gain horn, of course, effectively measures the disk temperature.

(4) The sun, even considered over a whole year, rises and sets over only a restricted range of azimuths, so that the v.p.d. can never be measured north or south of the station. If the ground contours vary considerably with azimuth, and if north-south is an important vector operationally, tests must be by flight trials.

(5) In high latitudes, and especially in winter, the elevation of the sun is limited; thus in lattitude 52°, and in midwinter results are only possible up to 14½° of elevation.

For these reasons, and as soon as the solar noise technique was established and reduced to routine, attention was turned to other stars which might make it possible to overcome some or all of these difficulties. Radio stars are generally small in angular size compared with a radar beam, or with any likely fine structure in the v.p.d., which meets difficulty (1) above; because of the small size, any non-uniformity of brightness is unimportant, while the output is constant and known, so that difficulties (2) and (3) are also met. The fourth point, however, that solar noise is only available over a relatively small range of azimuths is not much helped by considering other stars, since it turns out that of the three or four stars detectable with normal radar aerials and normal receivers (room temperature parametric amplifiers), the largest Cassiopeia A never descends below 20° in latitude 50°N: the next largest Cygnus A, comes down to about 3° above the northern horizon, and is therefore potentially useful: Taurus A, the third star, behaves like the midsummer sun, but is too weak for comfortable measurements with simple systems, while the fourth Virgo A, is probably only just detectable and rises and sets like the winter sun. Since larger aerials or special receivers may make all these stars useful, their positional data is given in Table 1, while Fig. 9 gives a plot of their flux against frequency. This flux, multiplied by half of the effective reception area of the aerial, will be the power received. It is to be noted that the flux given is the total flux and only half this quantity will be received on any one polarization: hence the factor of one half in the previous sentence.

**TABLE 1**
**Positions of the Sources**

| | R.A. | Declination |
|---|---|---|
| Cassiopeia A | 23h 21min | +58·4° |
| Cygnus A | 19h 58min | +40·6° |
| Taurus A | 5h 31min | +22·0° |
| Virgo A | 12h 28min | +12·7° |

*While these figures are accurate enough for use with wide beam radars, i.e. for the purposes of this article, they vary slowly with time, and more accurate figures are required when using very narrow beams, e.g. with very large satellite communications aerials. The position of the stars is calculated just as for the sun, except that the right ascension (instead of the equation of time) gives the (sidereal) time at which the star transits the Greenwich meridian. See Fig. 12.*

Attention was soon concentrated on Cygnus A, with three objectives:

(1) To enable the main features, at least, of v.p.d.'s to be checked on a bearing never covered by the sun, i.e. near north.

(2) To check the effect, if any, of the finite size of the sun on the measured v.p.d. especially the very important dip in the ground reflection pattern between 3° and 4° on the SA.120 aerial.

(3) To give an absolute measure of gain, or more accurately, of effective receiving area, by measuring the flux received. There are three factors involved in this measurement, namely, the flux of noise power from the star, the overall noise temperature of the receiving system, including the aerial, and the receiving area or effective gain, of the aerial. Hence it was proposed to use a standard figure for the flux, taken from the literature, to measure the overall noise temperature as accurately as possible, and so calculate the gain. Once this is known, it can be taken as a constant, being unlikely to change during the operational life of the aerial, and the technique can be used in reverse to a measure of overall noise temperature which is more likely to change and which is quite difficult to measure absolutely by conventional means. In effect, every aerial would have its "Cygnus factor" i.e. the ratio of star noise to set noise, and as long as this is maintained the radar performance is almost assured. Note here that the Cygnus factor is in fact a measure of $G/T$, gain upon overall

* The Marconi Company Ltd.

noise temperature, which is a commonly used figure of merit for large aerials, as used for instance for satellite communications. It is unfortunate that $G$ and $T$ occur in the radar equation as $G^2/T$, so that a measure of the $G/T$ ratio is not of itself an adequate measure of performance. The factor $G^2/T$ comes of course from the fact that primary radar is a two-way process, and $G$ enters in both ways, but $T$ only in reception: in contrast satellite communications or radio astronomy are one-way systems, so that $G/T$ describes the aerial completely.

The first experiments on Cygnus were done in 1958 on the L-band aerial whose v.p.d. is shown in Fig. 5. The receiver as before was a crystal mixer (noise factor about 6·5dB), and a signal large enough to be measurable, and therefore to give another measure of the gain, was confidently expected. However, the site of this aerial had been carefully chosen with other considerations in mind, and 'viewing' to the north of the station was obstructed by rising ground, and by towers, buildings and trees: moreover, the sharp (cosecant squared) fall off of gain above about 2°, together with the refraction correction of 0·7° at the lowest angle of elevation of the star, meant a severe downward revision of the expected signal. Signals were received (Fig. 10) but only weakly and erratically.

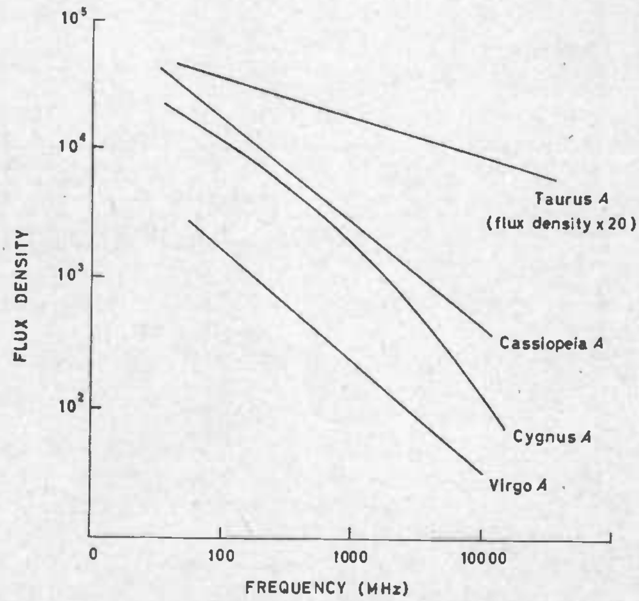



**Fig. 9. Flux density against frequency for the four most powerful sources.**
*The unit of flux density is $10^{-26}$ $Wm^{-2}$ $Hz^{-1}$. Based on Conway, Kellerman and Long, M.N.R.A.S., 125, 3, 261-84, 1963*

**Fig. 10. L-band signals from Cygnus A.**
*The double response arises from the use of a crystal mixer with a squinting aerial (linear array)*

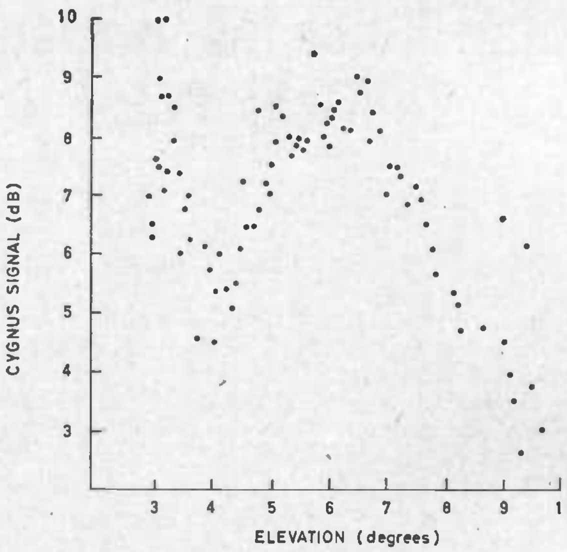



**Fig. 11. Signal strength from Cygnus A against elevation—tilt 4° nominal.**

There is little doubt that the use of a parametric amplifier, together with the more sensitive recording systems now in use, would give a signal useful for gain measurements.

In view of these disappointing results, no further work was done on radio stars until, in about 1963, parametric amplifiers at 600MHz became available. The initial problem was to compare, under operational conditions, the performance of the triode receiver (noise factor 7 to 8dB) and an electron beam parametric amplifier (Adler tube) since some controversy* had arisen[3] over the performance of coherently pumped Adler tubes as radar receivers. The question at issue was whether coherent pumping, i.e. pumping at exactly twice the signal frequency, so that signal and idler coincide, gave an improvement in radar range over non-coherent pumping, and if so, of what amount.

As a preliminary to checking radar performance, measurements were made on Cygnus, with the two receivers in turn connected to an SA120 aerial (see page 280). This experiment showed a difference of about 7dB in the Cygnus factor of the two receivers: this difference is therefore the difference in noise temperatures, since the same aerial was used in both cases. This experiment was difficult, since the Cygnus factor with the triode receiver was only about 2 per cent: nevertheless, it was much quicker, easier and probably as accurate as the inconclusive attempt to measure the change in radar performance which followed[4]. Here was an early example, under difficult conditions, of the advantages of star measurements for radar performance comparison.

When parametric amplifiers using varactor diodes and high idler frequencies became available, the work was resumed, there being no doubt under these circumstances about the equivalence of radar and star noise performance. In the first experiments, an attempt was made to measure the operational v.p.d., in order to check it near north and to investigate the shape measured with a very small source: did the sun in fact smooth out the ground reflection pattern? Results were obtained

** This controversy, incidentally, has had one long term effect, in that some people seem still to be uneasy about checking radar aerial and receiver performance with noise rather than with radar signals, since a coherently pumped Adler tube certainly gave different answers in the two cases. However, the question does not arise with non-degenerate parametric amplifiers, i.e. those in which the signal and idler frequencies are well separated: in such cases, performance with noise or radar signals is equivalent.*

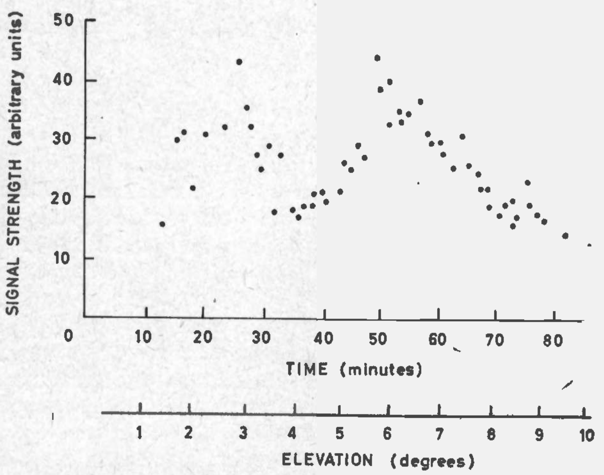



**Fig. 12. Signal strength from Taurus A, in arbitrary units, against time and approximate angle of elevation**

(Fig. 11) from about 3° in elevation to about 10°, showing the dip and the second main lobe, but they were not entirely satisfactory for two reasons, one instrumental and the other fundamental. Instrumentally, the results suffered from the fact that the receiver gain fluctuated considerably, so that calibration was needed (but not used) after every star pass. This difficulty was traced not to the parametric amplifier, as might have been expected, but to the effect of unstabilized heater voltages on the gain of the i.f. amplifier. The fundamental difficulty was that at low angles, and especially near north, 'twinkling' of the star signal occurred due to atmospheric effects, so that the received signal fluctuated so widely as to make some points fall far off the curve. The two points at about 3·2° on Fig. 11 are cases in point; here the pen recorder hit its stop twice running, while the preceding and succeeding points were only about half scale. For these reasons, the results shown in Fig. 11 do not conform closely to the expected lobe structure, although the general pattern is fairly clear. It was assumed that the sun is so large a source that twinkling is largely smoothed out, although some evidence of it was obtained in some sun runs (see Part 1).

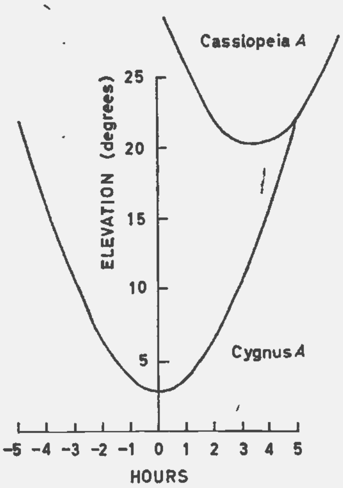



**Fig. 13. Elevation of Cas A and Cyg A near north.**
*Zero of time is when Cygnus is north, i.e. when sidereal time is 7 hours 58 minutes*

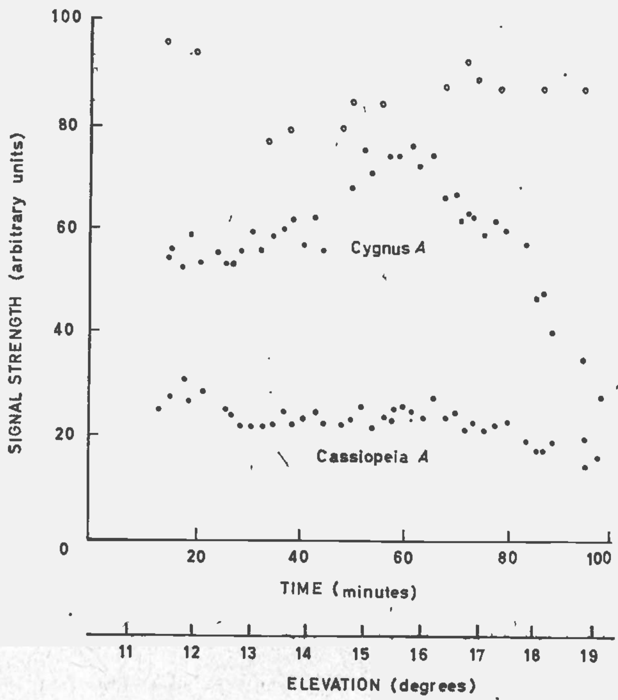
There was a further possibility, i.e. that twinkling occurred so severely on Cygnus and Cassiopeia because these sources are seen in the northern sky, which is known to be affected by auroral effects: was it possible that another star, seen away from the north, would give a more satisfactory ground reflection pattern? Accordingly, in November 1964, a run was done on Taurus, the third most powerful source in the sky, which rises and sets like the midsummer sun (but not, of course, at the same time, in general), followed next day by a run using Cygnus and Cassiopeia. The Taurus run, plotted in Fig. 12, again showed severe fluctuations, partly instrumental, but largely caused by twinkling: as before the general lobe and gap pattern was evident, but some points are wildly scattered.

It thus became clear that stars, other than the sun, are unsatisfactory sources for this purpose, at least at small angles of elevation: no doubt, by taking many more points on successive days, a fairly satisfactory picture could be obtained statistically, but this procedure was rejected as being almost as laborious as flight trials, in which twinkling, which will of course occur with high long range targets, is very much the lesser evil.

There still remained the possibility of measuring the free space gain, i.e. the gain with the aerial tilted so high that no ground reflection occurs. This position of the aerials would give less gain, of course, than the operational, low tilt, position, but it might have two advantages over that position. Firstly, at a high enough tilt, there would be no twinkling, and secondly, the aerial would now see a relatively cold sky, rather than a warm ground and atmosphere, so that the overall noise temperature would be lower, thus helping to offset the reduced gain. If the free space gain could be measured in this way, the gain in the operational tilt would follow, since it had already been shown that to convert from free space to operational gain was relatively easy using the sun.

The first measurements on free space gain were made in March 1965, using maximum tilt (15° nominal) and Cassiopeia and Cygnus as sources. The co-ordinates of these stars have already been given, and their elevations when they are near north are plotted against time in Fig. 13. The elevations plotted are for the site of these experiments, at Rivenhall, Essex, in latitude 51°52′N, and zero time is taken as that at which Cygnus crosses the meridian to the north, i.e. it is sidereal time 7 hours 12 minutes. Note that with an aerial tilted to 15°, Cygnus is visible descending and ascending, while Cassiopeia is seen only at or near minimum elevation; hence both stars can be seen by one sweep of the aerial after zero hour, but only Cygnus before it.

**Fig. 14. Signal strength from Cas A and Cyg A.**
*Tilt 15°. Circles are calibration points*



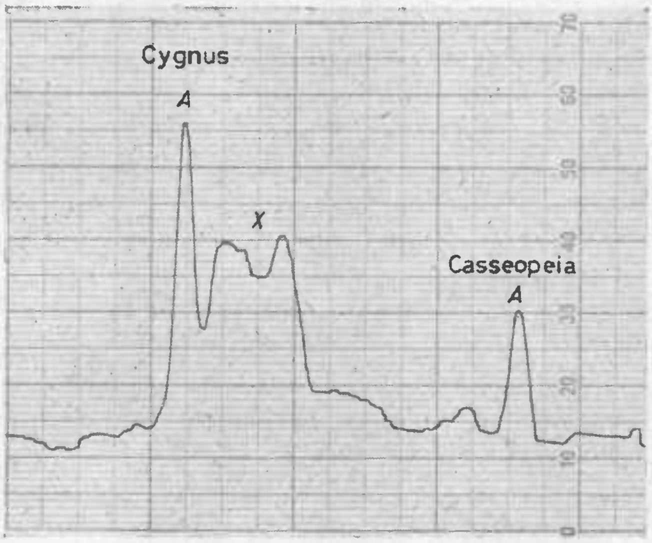



**Fig. 15. A typical pen record of one pass of the aerial.**
*From left to right, the stars are Cyg A, Cyg X (double humped), and Cas A*

In the first free space experiments measurements were taken both before (Cygnus only) and after (both stars) zero hour, there being a gap of about six hours between the two experiments. The results of this run are plotted in Fig. 14, in which the amplitude of the star signals is plotted against time and angle of elevation. The record is much better than before, in that there is now little sign of twinkling, but the receiver gain fluctuated during the run in a discontinuous way, as is shown by the spread of the calibration markers. Hence it would be difficult to use these results to obtain an accurate Cygnus factor or aerial gain. The chart recording of a single passage of the aerial through the stars is shown in Fig. 15: Cassiopeia gives a response against a clear background, but the response from Cygnus A is only just resolved from the neighbouring 'star' Cygnus X, which is in fact a large and complex hot area rather than a single star.

The next attempt, in October 1965, was modified in two respects. Firstly, the mains were carefully stabilized, to hold the i.f. gain steady, it having been shown that the major cause of gain instability previously was the variation of heater voltages in the i.f. amplifiers, all other voltages being stabilized independently. Secondly, it was convenient at this time of year to measure the free space pattern using the sun immediately before repeating the measurements on the stars. The free space pattern measured on the sun was intended to give an accurate curve, with which the star observations could be compared. However, two minor snags marred the experiment; firstly the receiver noise factor was not very good, since the amplifier was badly aligned, and secondly, the observations were given up too early, so that the star record stops just past the peak of the v.p.d. It is recognized in retrospect that the observers on this occasion were unduly influenced by non-scientific factors in giving up just after noon! The results of this run are plotted in Fig. 16, in the form of a v.p.d. taken on the sun, together with points measured on Cygnus, all plotted in decibels against angle of elevation. There is seen to be reasonable agreement between the shapes of the two curves, especially when the star signals are largest at the peak of the beam. Careful measurements were also made of the noise factor of the receiver (2·9dB, including the coaxial cable from the duplexer) and of the aerial temperature, by measuring the ratio between the noise on the aerial, and with the receiver terminated in a room temperature load. This ratio was 2dB, about twice that recorded with the aerial at operational tilt: this reduction in aerial noise at 15° tilt goes some way, at least, to off-setting the reduced aerial gain.

In the last experiment to date, carried out on 21 October, 1966, the noise factor of the receiver, complete with a shorter coaxial cable than before, was 2·0dB, and the noise ratio, receiver plus room temperature load to receiver plus aerial, was 2·8dB. The i.f. attenuator was altered to allow 0·1dB steps in attenuation: these steps and the 0·5dB steps previously used were carefully checked at d.c. to be well within 1 per cent. The procedure was, as before, to measure the v.p.d. on the sun, and to compare the shape measured on the stars with this. The results are shown in Fig. 17, from which it is seen that the fit between the sun and the Cygnus curves is excellent, although there is clearly some instability in the Cygnus results between 16 and 17°.

**Experimental Results**

In this section the Cygnus factor for the receiving system, and the gain of the aerial will be calculated from the last results available, and similar results will be given from Cassiopeia and from previous experiments without detailed calculation.

The flux from the stars is taken to be (Fig. 9)

Cas A $4{\cdot}8 \times 10^{-23}\,\mathrm{Wm^{-2}\,Hz^{-1}}$

Cyg A $3{\cdot}0 \times 10^{-23}\,\mathrm{Wm^{-2}\,Hz^{-1}}$

These figures must be divided by two to give the horizontally polarized flux.

On 21 October, 1966, the noise factor was 2dB, and the load/aerial noise ratio was 2·8dB.

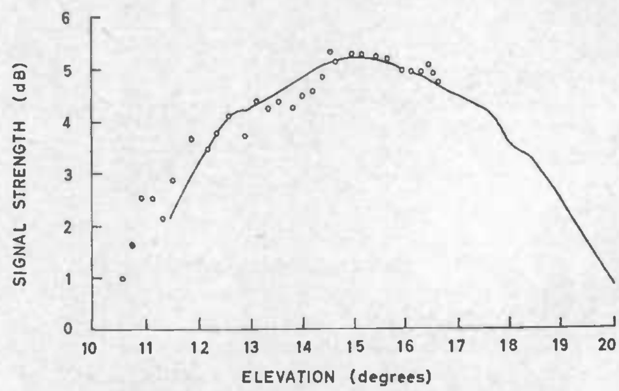
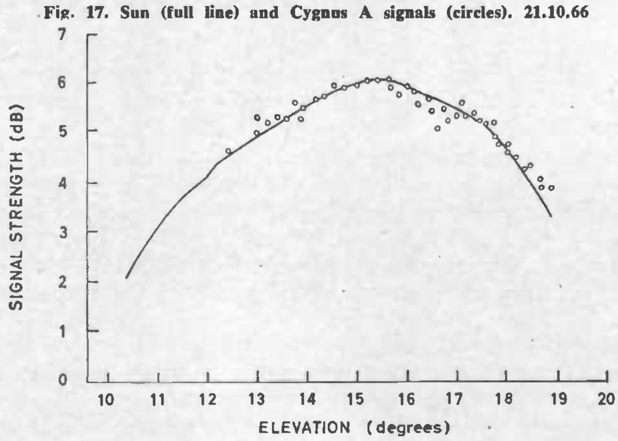



**Fig. 16. Sun (full line) and Cygnus (circles) signals.**
*Vertical levels adjusted for best fit near peak. Tilt 15°. 1.10.65*

**Fig. 17. Sun (full line) and Cygnus A signals (circles). 21.10.66**



Hence the receiver noise temperature, $T_R$, was 170° and the aerial temperature, $T_A$, given by $\frac{290 + 170}{T_A + 170} = 2{\cdot}8\text{dB} = 1{\cdot}91$, was 72°.

Hence the overall noise temperature, $T_{OA}$, was 242°.

On that day, the maximum Cygnus signal exactly equalled the 0·5dB calibration marker, i.e. the Cygnus factor, given by $a - 1$, was 12·2 per cent. The total power received from Cygnus is given by $(a - 1)$ times the receiver noise power, i.e. power from Cygnus = $0{\cdot}122 \times 242 \times 1{\cdot}4 \times 10^{-23}\text{W Hz}^{-1}$ and since the horizontally polarized flux from Cygnus is $1{\cdot}5 \times 10^{-23}\text{ Wm}^{-2}\text{ Hz}^{-1}$, the effective aerial area, $A'$, is $\frac{0{\cdot}122 \times 242 \times 1{\cdot}4}{1{\cdot}5} = 27{\cdot}6\text{ m}^2$ and the gain, $G$, given by $4\pi A'/\lambda^2$, is 31·2dB.

Working in the same way, a table may be prepared of the results to date (Table 2).

TABLE 2

| DATE | FREQ. (MHz) | OVERALL NOISE TEMP. | CYGNUS FACTOR (per cent) | GAIN FROM CYGNUS (dB) | CAS A FACTOR (per cent) | GAIN FROM CAS A (dB) |
|---|---|---|---|---|---|---|
| 1.10.65 | 597 | 336° | 10·5 | 32·3 | 3·7 | 30·6 |
| 21.10.66 | 585 | 242° | 12·2 | 31·2 | 4·2 | 29·7 |

**Discussion of the Results**

In Figs. 16 and 17 (especially), there is good agreement between the shape measured on the sun (full lines), and the points measured on Cygnus. Thus in Fig. 17, the greatest difference between the points and the line is 15 per cent (0·6dB), and most of the 40 points are very much closer than this. The error in the measurement of the ratio of the peak Cygnus signal to receiver noise (i.e. the measurement of 'Cygnus factor') is thus thought not to exceed 2 per cent. The error in the Cassiopeia factor is similarly estimated at about 5 per cent, assuming the receiver to be perfectly linear: the error is larger because the deflexions are smaller. These estimates of error are supported by the above results, in that the ratio Cygnus factor to Cassiopeia factor is the same within about 2 per cent on the two days.

Turning now to the measurement of gain, it will be noted that two additional factors must be known, the overall noise temperature of the receiving system and the flux density of the stars used. Both figures are difficult to estimate, since the overall noise temperature depends on two measurements (of receiver and aerial noise temperatures) about which it is notoriously difficult to be certain, largely because mismatch errors are so prevalent and indeterminate, while the flux density from the stars depends on data such as that in Fig. 9, which, even if accurate, is difficult to read. The overall noise temperature measurement is thought to be accurate, at least with such care as was taken before the last series of measurements, to ±0·25dB; the second difficulty remains, and is best dealt with by treating the gains so measured as related to the flux densities assumed, to be modified if better flux density figures becomes available. With this proviso, one would expect an overall error in gain of the order of ±0·3 or 0·4dB.

It is disappointing to find that the gains actually measured have a much wider scatter than this, a total spread of nearly 3dB between four measurements. Closer inspection reveals systemized differences: thus the measurements using Cas A are lower, by 1·7 and 1·5dB, than those using Cygnus. Here the explanation is uncertain: the flux figures taken may be wrong, the Cas A measurements, taken at over 20°, must be corrected to the peak of the beam at 15° using the known shape of the beam, and there is a possibility, at least, that the Cygnus A figures are inflated by flux from Cygnus X, since the two stars are not fully resolved, see Fig. 15. Again, the gains measured on 21.10.66 are lower, by 1·1 and 0·9dB, than those measured on 1.10.65: this difference is due in part (0·2dB or so) to the frequency difference, in part to the fact that 585MHz is on the edge of the design band of the aerial, and in part, it is suspected, to errors in overall noise temperature, especially on 1.10.65. The frequency difference between the two days was inadvertent and unfortunate.

**Conclusions**

The work discussed in Part 1 (the measurement of the shape of a vertical polar diagram) can safely be counted successful, despite its limitations: repeatable wide range and accurate measurements were made of the v.p.d.'s of radar aerials, so that flight trials could be relegated to their proper purpose, to measure aircraft echoing area, and to check the size, rather than the shape, of the radar cover against a particular target. Moreover, now that accurate measurements of shape are possible, the theory of ground reflection effects has received a considerable impetus[5]. The theory can now predict the cover to be expected on a given site, and these predictions can be checked, at least on some bearings. If the predictions agree where solar noise measurements are possible, one has greater confidence in them on all bearings. This new possibility of theoretical predictions, checked by measurements on a few bearings, removed one of the great difficulties of solar noise measurements, i.e. that they are available only on certain bearings near east and west.

The results discussed in Part 2 show that the 'Cygnus factor' of a receiving system can be measured with good accuracy (±0·1dB) and the 'Cas A factor' to lower, but probable acceptable, accuracy (±0·2dB). Hence one has a useful overall check on the long term performance of a radar, to be carried out at say three monthly intervals, or as required. However, the Part 2 measurements have failed to give any accurate measurements at low angles of elevation, because of 'twinkling', or to give consistent, repeatable and accurate free space gain measurements. Since the star measurements themselves are accurate enough, as demonstrated above, the errors must arise from the other data required (the flux densities and the overall noise temperature) or from the interpretation of the data (e.g. some of the flux attributed to Cygnus A comes in fact from Cygnus X).

It is hoped to resolve these difficulties in future experiments by using not only Cassiopeia and Cygnus but also Taurus A. This star gives a flux density of about 40 per cent of that from Cygnus A, and it passes through 15° elevation, i.e. it can be measured in the peak of the beam. The maximum signal from Taurus will thus be comparable with that from Cassiopeia, which although a more powerful source is only seen at about −5dB on the peak of the beam. Taurus, like Cassiopeia, is well clear of confusing sources, and measurements on it should help to resolve the spread of results existing at present. There is also a requirement for more accurate and trustworthy measurements of overall noise temperature (the difficulties of this measurement are being increasingly recognized[6]) and of receiver linearity, at the very low levels of input signals given by Cassiopeia and Taurus.

It will no doubt have been realized that these experiments have been carried out as and when permitted by other duties and by availability of the equipment. It is hoped that it will soon be possible to complete the experiments, and to demonstrate that absolute gain measurements are possible with the right technique.

**Acknowledgments**

In the course of a long series of experiments, such as these recorded here, the author has been helped by very many of his colleagues: particular thanks are due to Miss A. P. Darwent and Messrs. F. J. Gregory, A. K. Kwasieborski and F. Whybrow, for their support, moral and physical, and to the Station Staffs at Bushy Hill and Rivenhall Airfield, for their patience and co-operation. Dr. D. H. Shinn has freely given the benefit of his knowledge of astronomy and of radio astronomy in particular. The author is indebted to the Director of Research of the Marconi Company Limited for permission to publish.

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# A High Duty Cycle High Stability Monostable

By R. C. French*, C.Eng., A.M.I.E.R.E.

*A monostable circuit is described which is capable of duty cycles up to 99 per cent and which produce a pulse whose length is stable to 1 per cent with a variation in supply voltage of 10 per cent A change in pulse length of only 1·5 per cent occurs when the duty cycle is varied between 0 and 99 per cent.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

IN a high duty cycle monostable the duration of the quasi-stable state is a high proportion of the time between successive triggers. If the circuit is to be used reliably at high duty cycles, the period of the quasi-stable state must be constant. Should this period increase, the circuit could be in the quasi-stable state when a further trigger occurred so that the trigger would be ignored. By similar reasoning the trigger must also be stable, or if variable or random then never at a rate greater than the maximum acceptable to the monosable. The quasi-stable period should also be independent of the trigger rate. In the conventional monostable, shown in Fig. 1, the maximum duty cycle is about 70 per cent and the quasi-stable period is very dependent on supply voltage and pulse repetition frequency (p.r.f.).

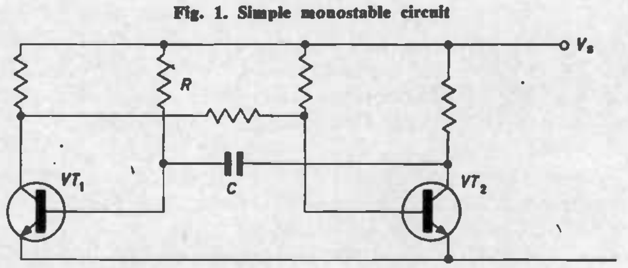
In the conventional monostable the capacitor $C$ is charged up to nearly $V_s$ while the circuit is in the stable state with $VT_1$ on and $VT_2$ off. When triggered into the quasi-stable state $VT_1$ is reverse biased until the voltage across the capacitor $C$ has decayed to nearly zero, and the stable state is resumed. However, although $VT_1$ is now on again and $VT_2$ off, the capacitor $C$ has still to charge up to nearly $V_s$ before the original conditions are restored. If the circuit is triggered again before the charging of the capacitor is complete, either the trigger will be ignored or the period of the resulting quasi-stable state will be shorter than before.

In order to overcome the defects of the simple circuit of Fig. 1 the new monostable was developed.

## Circuit

The circuit of the monostable is given in Fig. 2. The transistors $VT_1$ and $VT_2$ form a conventional bistable pair. In the stable state $VT_1$ is on and $VT_2$ is off so that the collector voltage of $VT_2$ is at the voltage $V_1$. $V_1$ is determined by the divider chain consisting of $R_2$, $MR_2$ and $R_{10}$, $R_{11}$. The transistor $VT_3$ is conducting and its emitter is at $(V_1 - V_{be3} + V_{D2})$ and $VT_4$ is reverse biased. When the circuit is triggered by a positive pulse to the base of $VT_2$ the bistable changes state so that $VT_1$ is now off and $VT_2$ on. The collector voltage of $VT_2$ drops to about 1V

**Fig. 1. Simple monostable circuit**




* *Mullard Ltd.*

and cuts off $MR_2$ and reverse biases $VT_3$. The voltage $V_1$ is maintained because $MR_1$ is now conducting. With $VT_3$ reverse biased the capacitor $C$ now charges through $R_9$ until the voltage at the base of $VT_4$ goes negative relative to the voltage at the junction of $R_{10}$, $R_{11}$ and $VT_4$ conducts. When the current flowing through $R_6$ due to $VT_4$ is sufficient the bistable resets to $VT_1$ on and $VT_2$ off. $VT_2$ collector goes positive and heavily forward biases the emitter-follower $VT_3$ which rapidly discharges the capacitor.

Transistor $VT_3$ suffers from reverse base emitter voltage during the unstable state which, without either $R_8$ or a protective diode in series with its base or emitter, would result in Zenner breakdown. The use of a protective diode slows up the discharge of the timing capacitor and reduces the maximum duty cycle. A further difficulty is with breakdown between collector and emitter of $VT_3$ with the large current that flows during the discharge of the capacitor. The BSY39 is satisfactory but other transistors have been found inadequate.

The period of the quasi-stable state may be determined externally by applying a positive pulse through a diode to the base of $VT_1$ before the normal circuit action takes place. In a conventional monostable this is only possible towards the end of the quasi-stable period.

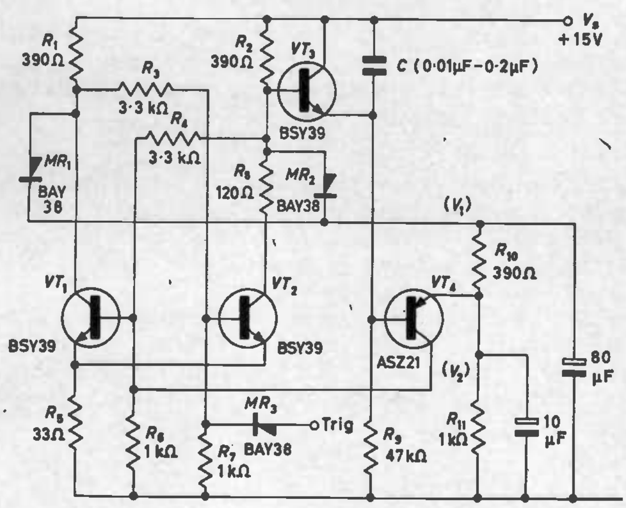
The circuit may be used as a self-gating triggered ramp generator. If the linearity of the ramp is inadequate it can be improved by increasing the voltage dropped across $R_9$ relative to the amplitude of the ramp, either by increasing the supply voltage, or by returning $R_9$ to a negative supply. If the latter were done, however, the pulse length stability would be no better than the stability of the negative supply voltage.

The pulse length can be varied by adjusting $R_9$ or $C$. Values of $C$ were used between $0{\cdot}01\mu$F and $0{\cdot}2\mu$F giving pulse lengths between 200$\mu$sec and 4msec with $R_9$ set to 47k$\Omega$. The value of $R_9$ can be varied between 1k$\Omega$ (below which $VT_3$ passes too much current in the stable state) and 47k$\Omega$ (above which $VT_4$ will not pass sufficient current to reset the bistable).

## Theory

By arranging for one side of the capacitor to be at earth (or supply) and using an emitter-follower the timing capacitor can be restored to its normal voltage rapidly enough to make duty cycles as high as 99·5 per cent quite

Fig. 2. The improved circuit



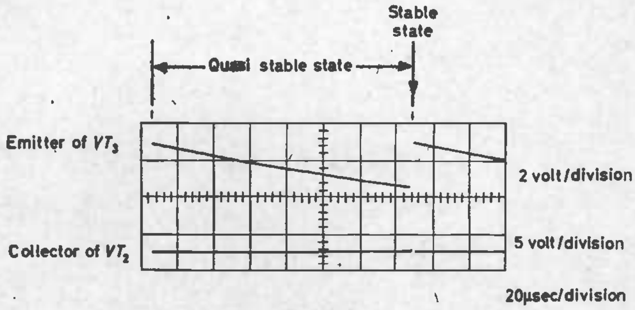




Fig. 3. Circuit waveforms

possible. The high stability of the pulse length against change in supply voltage is achieved in the following way. The voltage across the capacitor varies between $(V_1 - V_{be3} + V_{D2})$ in the stable state to $(V_2 - V_{be4})$ at the end of the quasi-stable state.

Now $$V_1 \simeq \frac{R_{10} + R_{11}}{R_2 + R_{10} + R_{11}} \cdot V_s = k_1 V_s$$

$$V_2 = \frac{R_{11}}{R_{10} + R_{11}} \cdot V_1 = k_2 V_s$$

where $k$ = constant.

The change in voltage across the capacitor during the quasi-stable state is

$$V = (V_1 - V_{be3} + V_{D2}) - (V_2 - V_{be4})$$

but $V_{be3} \simeq V_{D2}$ and $V_{be4} \ll V_2$

so that $V \simeq k_1 V_s - k_2 V_s$

$= V_s (k_1 - k_2)$

The rate of change of voltage across the capacitor is

$$dv/dt = i/c$$

where initially $i = V_1/R_9$ (taking $V_{D2} = V_{be3}$)

Since the change in voltage across $C$ is small compared to $V_1$ the charge characteristic can be taken as linear so that

$$dv/dt = V_1/R_9C$$
$$= k_2 V_s/R_9C$$

The duration of the quasi-stable state $T_u$ is the voltage change divided by the rate of change

i.e. $$T_u = \frac{V_s (k_1 - k_2) R_9C}{V_s k_2} = k_3 R_9C$$

$T_u$ to a first approximation is therefore independent of the supply voltage.

## Conclusions

With pulse lengths in the range of 200$\mu$sec to 4msec a maximum duty cycle of 99 per cent was achieved. The change in pulse length due to 10 per cent change in supply voltage or due to warm up when switched on, was less than 1 per cent. Changing the duty cycle from 0 to 99 per cent reduced the pulse length by 1·5 per cent.

The waveform at the emitter of $VT_3$ is shown at the top of Fig. 3 together with the waveform at the collector of $VT_2$. The waveform at the collector of $VT_1$ is the inverse of that shown for $VT_2$.


### Acknowledgment

Acknowledgment is made to the Ministry of Aviation who supported this work.

# A Stable 50Hz Invertor for Mains Operated Equipment

By W. T. Maloney*, B.Sc. (Hons)

*An invertor design which provides good frequency stability at mains frequencies is described. The invertor operates from a 12V supply and provides an output of about 40W at 240V. It is shown that with slight modifications reliable operation can be obtained from a 6V supply.*

(Voir page 407 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

MANY types of invertor are available for operating mains equipment from batteries. The purpose of this article is to describe a simple design for applications where medium power and precise frequency control are required. Such an application is the operation of recording equipment in vehicles. The invertor about to be described was used to drive a good quality transistorized tape recorder, with a rated power consumption of 40W at 240V a.c., from a 12V accumulator. It is useful for any apparatus which relies on mains synchronization for proper functioning.

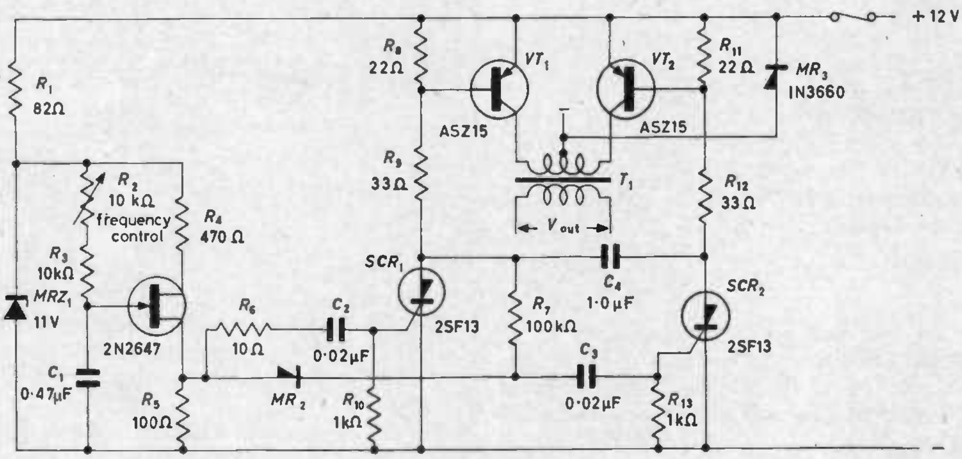



**Fig. 1. Invertor circuit**

## General

The silicon controlled rectifier invertor is probably the most popular type today, particularly for high power applications. This has come about because of the inherent high voltage, high current capability of the s.c.r., without the restriction of a 'safe' area of operation, as with the transistor. However, for low supply voltages, the transistor invertor has certain advantages, one of these being improved efficiency resulting from the lower saturation voltage of the transistor compared to the 'on' voltage of an s.c.r. (typically 1 to 2V).

The s.c.r. invertor can also present difficulties when operated into varying reactive loads. An invertor with non sine wave output driving equipment designed for 50Hz a.c. use may result in saturation towards the end of each half cycle. This reflects a low impedance into the primary winding of the invertor transformer, and in the case of a parallel s.c.r. invertor results in a large current pulse appearing at the anode of the conducting s.c.r. If the commutating circuit has been designed for reasonable efficiency with normal loads, the circuit may fail to commutate and a large current will flow through both s.c.r.'s, limited only by resistance of transformer and inductance windings.

In an equivalent transistor system, the situation mentioned above would cause the transistors to tend to come out of saturation, with resulting higher power dissipation for this part of the cycle. However average power would only slightly increase, and the circuit would still function. This is a particularly important aspect, when an invertor is required for general-purpose use.

## Practical Circuit

The circuit of the invertor is shown in Fig. 1. Frequency control is achieved by adjusting $R_2$, which varies the pulse rate from the free running unijunction transistor oscillator. Supply to the oscillator is held constant from an 11V Zener diode $MRZ_1$.

Pulses from the base 1 of the unijunction transistor, at 10msec intervals, are fed to the gates of $SCR_1$ and $SCR_2$. These are arranged in a conventional commutating circuit, with resistive load. The size of the commutating capacitor is given by

$$C \geqslant \frac{1{\cdot}4 t_{off} I}{E} \text{ microfarad.}$$

* *Department of Civil Aviation, Australia.*

where $t_{off}$ = turn off time of the s.c.r. in microseconds
$I$ = maximum load current in amperes at time of commutation
$E$ = minimum d.c. supply voltage.

A capacitor of $1\mu F$ is sufficient in this case, as there is no possibility of overloads. The commutating capacitor should be no larger than necessary, as charging current for the capacitor is drawn through the 'off' transistor, so decreasing the efficiency.

Diode $MR_2$, resistor $R_7$ and capacitor $C_3$ provide a gating circuit to ensure correct starting of the invertor. When the supply is first switched on $MR_2$ becomes back biased by the supply potential, and so the first pulse is directed to the gate of $SCR_1$. This back bias is removed once $SCR_1$ fires, so enabling the second pulse to trigger $SCR_2$. Commutation occurs because the commutating time-constant is larger than the gate pulse time-constant.

The two s.c.r.'s provide a highly stable square wave output, used to drive directly the bases of the output power transistors $VT_1$ and $VT_2$. Without filtering, the output voltage will be a square wave, and the transformer should

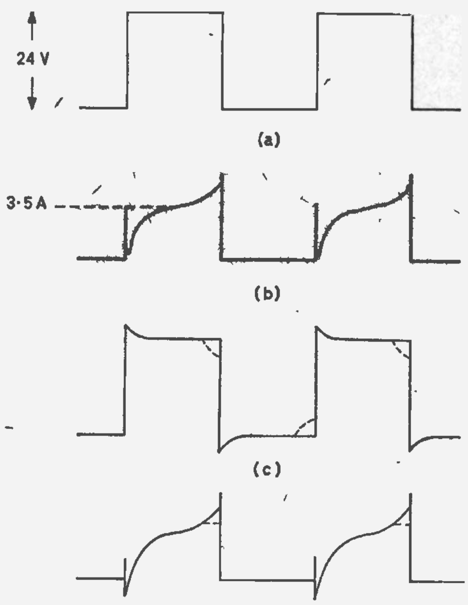



Fig. 2. Voltage and current waveforms

be designed to ensure that saturation does not occur. A square wave output is satisfactory in most cases, but is likely to cause buzz in small induction motors. If a filter is added it is of course necessary to design the transformer to cope with the insertion loss of the filter.

**Invertor Waveforms**

Voltage and current waveforms for one collector are shown in Fig. 2. With a 40W resistive load the waveforms shown in Fig. 2(a) and (b) were observed. The voltage waveform of (a) is also the same as the output voltage and current waveforms. For a 40W reactive load the waveforms vary as in Fig. 2(c) and (d). The dotted line indicates the effect of saturation in the latter part of the cycle, causing high power dissipation in the output transistors. The peak of the current waveform becomes limited as the transistor comes out of saturation. For any particular equipment this effect may be remedied by increasing the base drive, to regain the normal waveform. It must then be decided if the current waveform in the equipment itself is acceptable. It will be noted no voltage transients appear at the collector during the switching interval. To this end the primary of the output transformer was bifilar wound.

**Efficiency**

Power losses in the invertor are mainly due to the base drive, transformer, and drop over the power transistors. A large transformer ensures good efficiency, and in this case a 1¼in core wound with 15 B & S gauge enamelled wire was used. The fixed loss in the base drive circuit is dependent on what maximum power output is required from the system. The graph of efficiency as a function of power output is shown in Fig. 3. It can be seen that the efficiency falls off as the power output exceeds 60W. Above this the drive to the output transistors is insufficient to maintain saturation, and both efficiency and regulation suffer. Power output of over 100W may be obtained by increasing the drive.

**Lower Voltage Operation**

Operation from a 6V supply is now considered. The invertor may be operated quite reliably from a 6V supply provided the circuit is optimized, with respect to triggering conditions for the silicon controlled rectifiers. Manufacturer's recommendations do not advise operation of the unijunction transistor at supply voltages less than

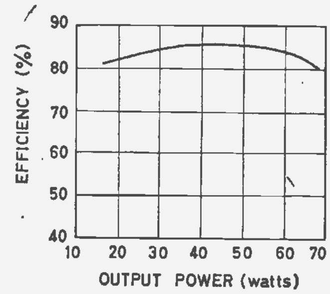



Fig. 3. Efficiency versus output power

10V, due to unacceptable values of signal amplitude. The circuit will, however, function quite reliably at lower voltages—in fact oscillation will occur from slightly above 2V.

Assuming a Zener voltage of 5·8V, base 1 pulses of approximately 2·5V are available. This amplitude will safely trigger $SCR_1$, provided an s.c.r. of reasonable gate sensitivity is used. With the second s.c.r. the residual charge on $C_3$ (maintained by the 'on' voltage of $SCR_1$) must first be overcome, as well as the drop over $MR_2$. A $10k\Omega$ resistor placed from the junction of $R_7$ and $C_3$ to the negative rail, with a germanium diode for $MR_2$, will provide reliable trigger conditions to $SCR_2$. To operate from the lower supply involves an increase in both base drive and commutating capacitor. By using a 1W Zener for $MRZ_1$, with $R_1$ at $82\Omega$, and high sensitivity s.c.r.'s, a combined 6 to 12V invertor in which only the transformer primary winding is switched can be constructed.

**Conclusion**

An invertor with improved stability compared to feedback systems has been described, while still operating efficiently. Simple frequency control combined with reasonable power handling capability provide a versatile unit for general application.

# A Means of Displaying Proportional Plus Rate Information on a C.R.T. in the form of Two Pointers

By M. R. Green*, C.Eng., A.M.I.E.R.E. and K. Lord*

*In this article circuits are described which enable information to be displayed in the form of two pointers on a c.r.t. It in effect serves the function of a mechanical pointer instrument but all mechanical moving parts are eliminated. In the example described, the equipment is intended for airborne use and displays height and vertical speed on the same display.*

(Voir page 408 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

As part of a general investigation into possible new forms of instrument display, consideration has been given to the use of small c.r.t's to present information with a view to eliminating any mechanical moving parts and so improving potential reliability. The c.r.t. presentation has greater flexibility and allows the display of additional information which mechanically might be very difficult or sometimes impossible.

As a general principle it has been assumed that for any given quantity a digital presentation is required but that this in itself is not sufficient and that an additional indication giving a clue to trend and rate in the form of a rotating index (equivalent to the pointer of the mechanical instrument), may also be required. As a practical example experimental work has been carried out on a form of height display; height is displayed digitally by numerator tubes giving thousand and hundreds of feet and in an analogue fashion by a radial line on a c.r.t. rotating once every 1 000ft. Additionally it was thought that a display of vertical speed on the same display would be valuable and this was indicated by another line rotating with the height index but separated from it by an angle proportional to vertical speed, ahead for climb and lagging for dive.

This note gives details of the methods employed to produce the c.r.t. section of such a display. It is assumed that the inputs are d.c. voltages with a common earth.

### General Description

The basis of the system is that a circular trace is produced at the periphery of a c.r.t. face and, at some time interval from a zero datum which is dependent on the input signal voltage, the spot of light is made to travel slowly to the centre of the c.r.t. display face. A radial line is thus drawn on the display. By means of time sharing, two lines may be drawn, one representing the input signal and the other of shorter length its derivative.

With reference to Fig. 1 the action of the circuit is as follows: A suitable reference frequency, in this case 400Hz, is used as a driving signal to produce the rotating spot of light, i.e., circular time-base. This reference frequency is further used to produce a switching reference pulse, which is synchronized with the circular time-base. Thus a reference pulse has been produced which will indicate a defined datum position on the c.r.t. face. The input analogue information is transformed into pulse width modulation by means of the modulator, this being synchronized to the circular time-base by means of the reference pulses. The output of the modulator will be a train of pulses whose width will be dependent on the input signal and whose leading edge will be referenced to the datum of the display. The trailing edges of these pulses

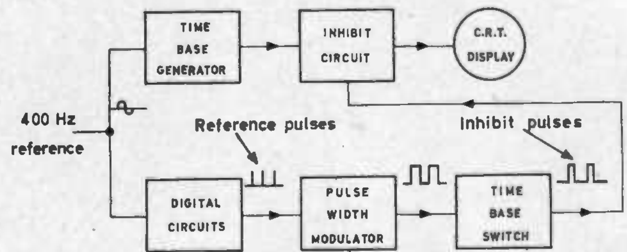



Fig. 1. Basic arrangement of display circuits

will therefore also define a particular point on the display and as the pulse width varies so will this defined position. This trailing edge is now used to trigger the time-base switch which inhibits the circular time-base signal to the c.r.t. When this occurs the rotating spot will return to the centre of the display with an exponential time-constant, thus drawing out a line, the length of line being dependent on the period of inhibition of the time-base. By means of synchronized time sharing several analogue inputs may be displayed, recognition of any signal being made by the length of the line drawn on the display.

Referring to the block schematic Fig. 2 and the illustration of the typical waveforms in Fig. 3, the generation of the circular time-base will be explained in more detail. It is achieved by producing two equal quadrature signals from the 400Hz reference and applying them to the $X$ and $Y$ amplifiers of the c.r.t.; one signal is given a 45° phase lag, the other a 45° phase lead as shown in Fig. 2. These two signals are applied to the 'sample and hold' circuit. This circuit essentially consists of two transistor series switches feeding memory capacitors shunted by resistors. When the transistor switches are closed the voltage across the memory capacitors will follow the amplitude of the two phase shifted signals. When, however, the switches are opened, by means of the inhibition signal, the existing initial voltages will subsequently leak away through the shunt resistors causing the spot on the c.r.t. to decay towards the centre. On closing the switches at the end of the inhibit pulse the capacitors will immediately take up the voltages of the two quadrature signals thereby causing a rapid flyback.

The next part of the circuit to be explained is the

* *Royal Aircraft Establishment.*

generation of the reference and switching pulses. Reference should be made to Figs. 2 and 3, the waveforms at the points denoted by ringed letters in Fig. 2 are shown against the same letter in Fig. 3. The 400Hz reference signal is first squared in $FF_1$ to produce a square wave of suitable form for triggering the divide by four circuits $FF_2$ and $FF_3$. The output of the divide by four circuit is a 100Hz square wave, see Fig. 3(b). One of the outputs of $FF_3$ is further divided in $FF_4$ to provide a 50Hz square wave commutating signal, see Fig. 3(f), for the time sharing processing of the analogue input signals. The other output of $FF_3$ triggers delay multivibrator (monostable) $T_1$ which produces a narrow pulse, see Fig. 3(c). This narrow pulse performs two functions; firstly it provides a reference pulse so that the triggering of delay module $T_2$, which acts as the pulse width modulator, is synchronized to the display and secondly, as the trailing edge of this narrow pulse is used to trigger delay $T_2$, a delay is inherently provided so that sufficient time is allowed for the commutator switches $S_1$ and $S_2$ to settle down before $T_2$ is triggered, see Figs. 3(c), (d), (f). The width of the pulse generated by $T_2$, on being triggered by $T_1$, is dependent on the commutated d.c. signal applied to its d.c. input.

This d.c. signal is the time shared processed analogue input signals, i.e., proportional or proportional plus derivative. Thus the output of $T_2$ will be a train of pulses whose width will be dependent on the amplitude of the analogue signals, see Fig. 3(d). It should be pointed out that even when the analogue input is zero a pulse is produced whose width is approximately equal to two periods of the 400Hz reference signal. When the maximum analogue signal and its positive derivative are present the pulse width from $T_2$ will be approximately equal to $3\frac{1}{2}$ periods of the reference. The trailing edges of these pulses now trigger delay module $T_3$ which again produces pulses whose width is dependent on a d.c. controlling voltage. But in this case the pulse has only two widths, because this delay module dictates the line length drawn out on the c.r.t. face, see Fig. 3(e). Thus as can be seen from the

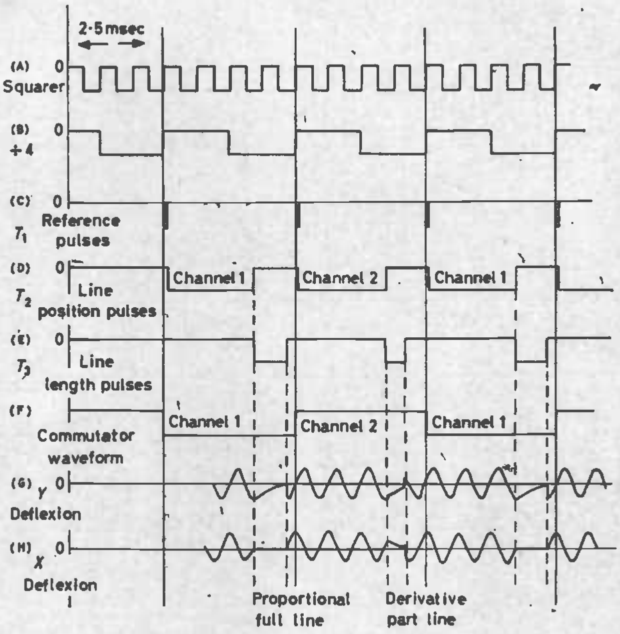
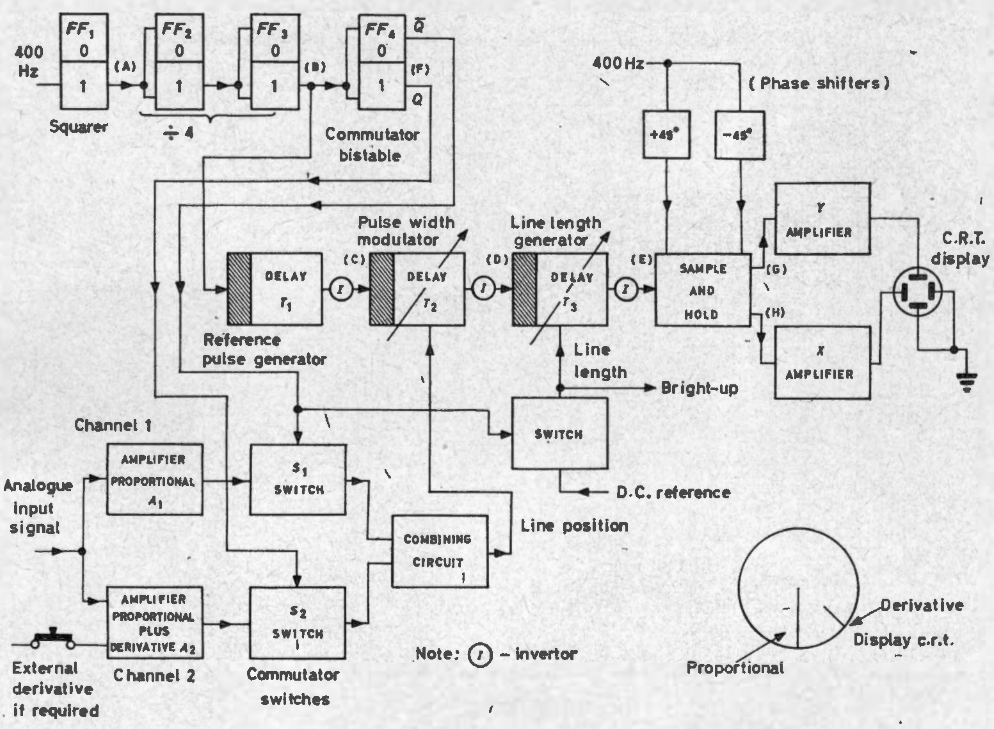



**Fig. 3. Typical waveforms**

above and from Fig. 3 a series of pulses have been produced (see Fig. 3(e)) which are synchronized to the 400Hz reference frequency but vary in time delay dependent on the analogue input signals. Consequently when the output of $T_3$ is used to turn off the transistor switches in the 'sample and hold' circuit the position of the 'pointers' will vary depending on the time relationship between the reference pulses and the switching pulses which in turn are dependent on the input analogue voltages.

The anologue input signals to be displayed are fed to two operational amplifiers $A_1$ and $A_2$, see Fig. 2. The output of amplifier $A_1$ is proportional to the input signal and the output of amplifier $A_2$ is proportional to the input plus its derivative. The reason for both amplifiers producing a proportional signal is that when there is no rate term present both pointers on the c.r.t. display must coincide. The derivative term can either be formed in the amplifier $A_2$ circuit or generated externally and then added to the proportional signal in the amplifier. The outputs of these amplifiers are fed to two series switches $S_1$ and $S_2$, whose states are dependent on the switching (commutator) signal derived from bistable $FF_4$. In fact the switches conduct alternately thus allowing the signals from the operational amplifiers to be time shared and fed to the combining circuit which then feeds the d.c. input of the delay module $T_2$ (pulse width modulator).

**Fig. 2. Arrangement of display circuits**



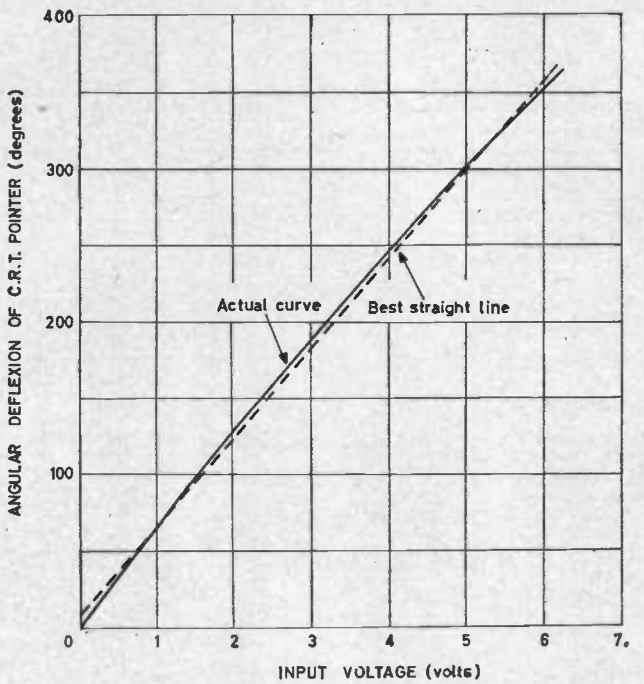



**Fig. 4. Graph of input voltage against angular deflexion of c.r.t. pointer**

The linearity of the pulse width modulator $T_2$ is shown in Fig. 4. A 'best straight line' through the curve shows accuracy limits of $\pm 6°$. To improve the linearity the delay module must be modified as mentioned later.

### Detailed Circuit Descriptions

SIGNAL INPUT AMPLIFIERS (Fig. 5)

Integrated circuit operational amplifiers with a nominal gain of 1 000 are used for the signal channels. The input resistor networks $R_3$ and $R_4$ and $RV_8$ are used for biasing the outputs of the amplifiers for positioning the two pointers on the c.r.t. face with respect to each other. $R_5$ and $R_6$ provide the proportional signal for each amplifier and $R_7C_1$ provides the signal derivatives for $A_2$. If a separate external derivative signal is required, then $R_7C_1$ can be switched out by means of $S_3$ and resistor $R_9$ used instead. The outputs of the two amplifiers are fed to two series transistor switches $VT_1$ and $VT_2$. These switches are controlled by the $Q$ and $\bar{Q}$ outputs of control commutator bistable $FF_4$ so that they are opened and closed alternately. The switches feed a combining circuit consisting of $R_{10}$, $R_{11}$ and a complementary emitter-follower, the emitter-follower having an output impedance low enough to feed the d.c. signal input of the delay module $T_2$.

Diodes $MR_1$ and $MR_2$ across the amplifiers limit any negative excursions of the outputs thus preventing damage to delay module $T_3$.

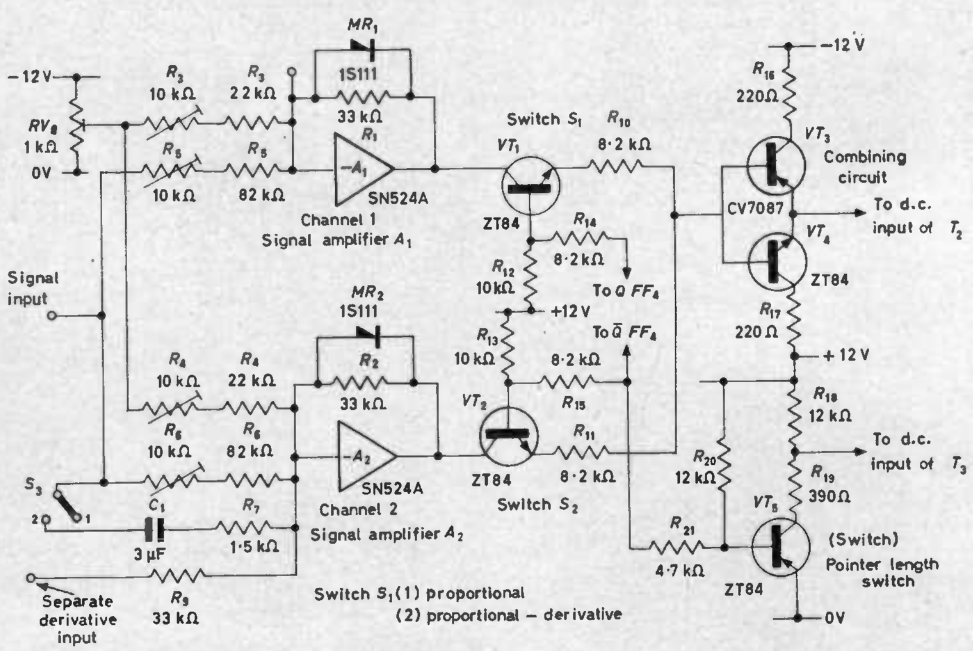
POINTER-LENGTH SWITCHING CIRCUIT (see Fig. 5)

This is a simple transistor switch controlled by the same source as that controlling switch $VT_2$ in such a way that, when transistor switch $VT_2$ is closed and the signal plus its derivative is being displayed, transistor $VT_5$ is on and the d.c. signal input voltage to delay module $T_3$ is reduced thus making the delay module $T_3$ give a shorter output pulse and hence a shorter pointer length.

PULSE WIDTH MODULATOR OR DELAY MULTIVIBRATOR (see Fig. 6)

This circuit is of conventional design with the added facility of being able to vary the output pulse width depending on the magnitude of the d.c. input signal. The output pulse width is primarily governed by the two components $C_4$ and $R_7$ and the amplitude of pulse transferred from $VT_1$ through $C_4$ to the base of transistor $VT_3$. Before the trigger pulse is applied $VT_3$ is on and the capacitor $C_4$ is charged up negatively to the negative 12V supply through $VT_1$ and the base of $VT_3$. On receipt of the positive trigger pulse $VT_3$ is turned off and $VT_2$ is turned on by the cross coupling network $R_8$,$R_9$. Now the left-hand side of the capacitor $C_4$ will now be effectively earthed and a positive voltage will be applied to the base of $VT_3$ keeping it off. The charge on $C_4$ will gradually leak away to the negative supply through $R_7$, and when this voltage is nearly zero $VT_3$ will switch on. If the diodes $MR_9$ and $MR_8$ were omitted the positive voltage applied to the base of $VT_3$ would be approximately +12V. With $MR_9$ connected this positive voltage is 'caught' at a potential set by $RV_1$. Thus if it is assumed that the discharge of $C_4$ is initially linear the waveform at the base of $VT_3$ will be very similar to a right-angled triangle whose apex will be dependent on the catching voltage and consequently as the time-constant remains the same the base of the triangle must also vary in the same proportion as to the height. Thus by varying the catching voltage the time that $VT_3$ remains off may be varied reasonably linearly with an applied d.c. 'catching' voltage, see waveform of base voltage in Fig. 8. The d.c.

**Fig. 5. Signal input amplifiers commutator switches and combining circuit**



input signal which is to be pulse width modulated is applied through $MR_8$ and the effect on the output pulse width is the same as that explained for the diode $MR_9$, providing this signal is less than that supplied to $MR_8$ by $RV_1$. The purpose of $RV_1$ is to give a fine variation of output pulse width during initial setting up. The emitter-follower transistor $VT_1$ speeds up the charging of $C_4$ by providing a low output impedance for the collector of transistor $VT_2$. The output pulse is taken through an invertor amplifier (transistor $VT_4$) which provides a low output impedance and resets the logic levels. As the delay monostable circuit requires a positive going waveform for triggering, invertor amplifiers must be connected between each delay module, as shown in Fig. 1 to produce consecutive delays.

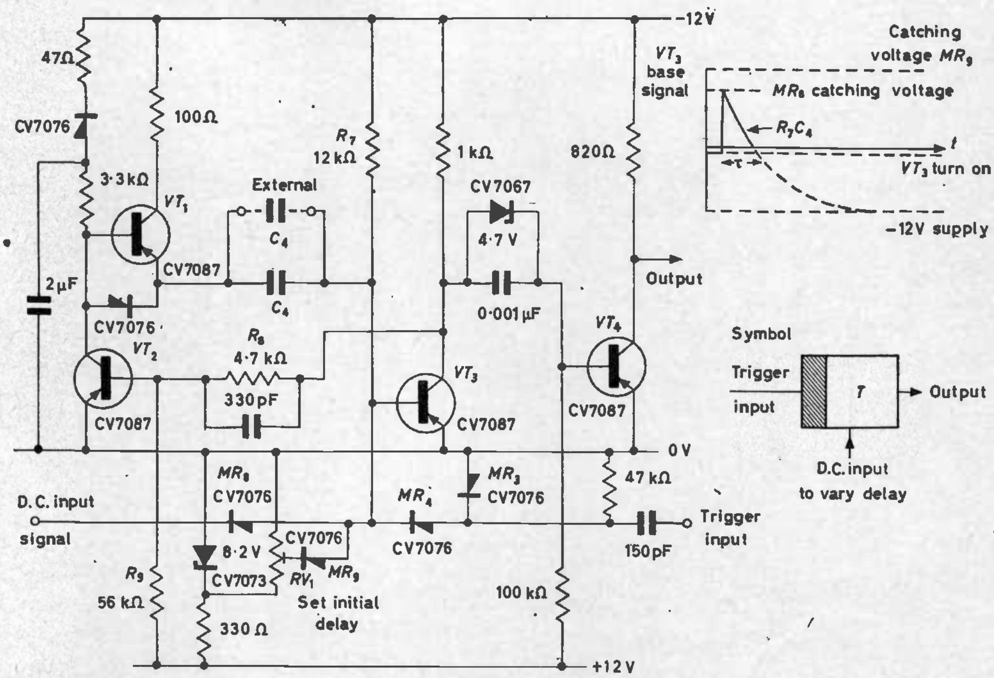


**Fig. 6. Pulse width modulator or delay monostable c.c.t.**

The accuracy of the input voltage to time conversion is mainly governed by the exponential discharge of $C_4$ and for better linearity the charging resistor $R_7$ should be replaced by a constant current source by the inclusion of a few components and an npn transistor.

The approximate transfer function is given below:

$$\tau = CR \ln \frac{V_s + V_{in}}{V_s}$$

where

$\tau$ = the width of the pulse in seconds

$CR$ = time-constant components $C_4$ $R_7$

$V_s$ = the supply voltage of 12V

$V_{in}$ = d.c. input signal.

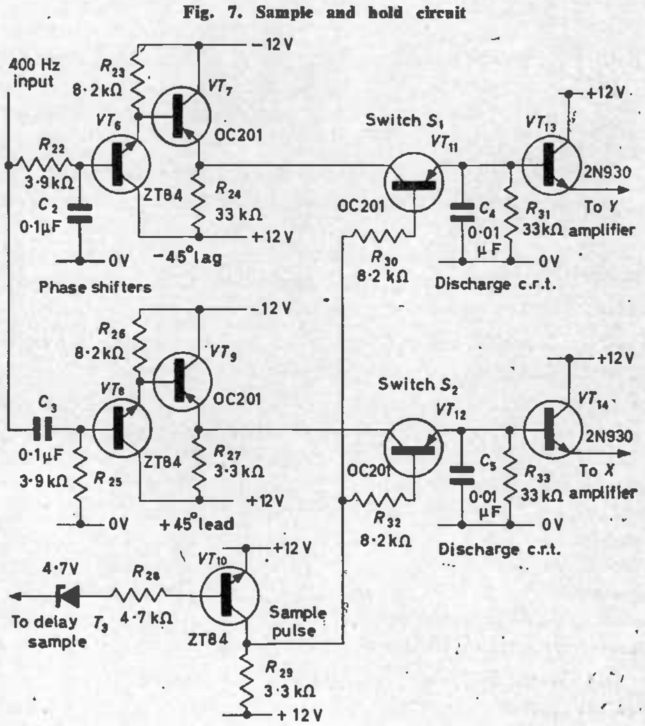
## 'Sample and Hold' Circuits

The 400Hz input signal is fed to two phase shifting networks, $R_{22}$ $C_2$ and $C_3$ $R_{25}$, which produce 45° lead and lag signals respectively. The outputs of these two networks feed complementary emitter-followers of high input impedance and very low d.c. offset voltage. The emitter-followers then feed the discharge networks $C_4$ $R_{31}$ and $C_5$ $R_{33}$ through series transistor switches $VT_{11}$ and $VT_{12}$. These switches are controlled by the sample pulse from the delay module $T_3$ which, when present, turns off the switches. Because of the low source impedance feeding the transistor switches $VT_{11}$ and $VT_{12}$ the discharge capacitors $C_4$ and $C_5$ will follow the 400Hz input signal amplitude. When the transistor switches are opened, however, the instantaneous level of the 400Hz input signal will be stored by the capacitors $C_4$ and $C_5$ and will leak away exponentially through the resistors $R_{31}$ and $R_{33}$. It should be noted that these discharge circuits must be matched, or else the line drawn on the c.r.t. face will not be straight. The rate of discharge will effect the brightness of the line drawn and if a constant brightness is required then the discharge resistors should be replaced by constant current discharge circuits.

## C.R.T. and Deflexion Amplifier (see Fig. 8)

The c.r.t. deflexion amplifiers consist of two identical amplifiers in the cascode circuit configuration. This form of circuit was used to enable transistors with limited $V_{ce}$ rating to be used to produce driving signals to the c.r.t. deflexion plates of 140V peak-to-peak. The resistors $R_{35}$ $R_{36}$ and $R_{40}$ $R_{41}$ ensure that the output signal swing is equally divided between the transistors thus making sure that the maximum $V_{ce}$ can never be greater than 90V. Resistors $R_{38}$ and $R_{43}$ give some measure of gain stability and give each cascode stage a gain of approximately 20. The emitter resistor decoupling capacitors, $C_6$, $C_7$, give high frequency

**Fig. 7. Sample and hold circuit**



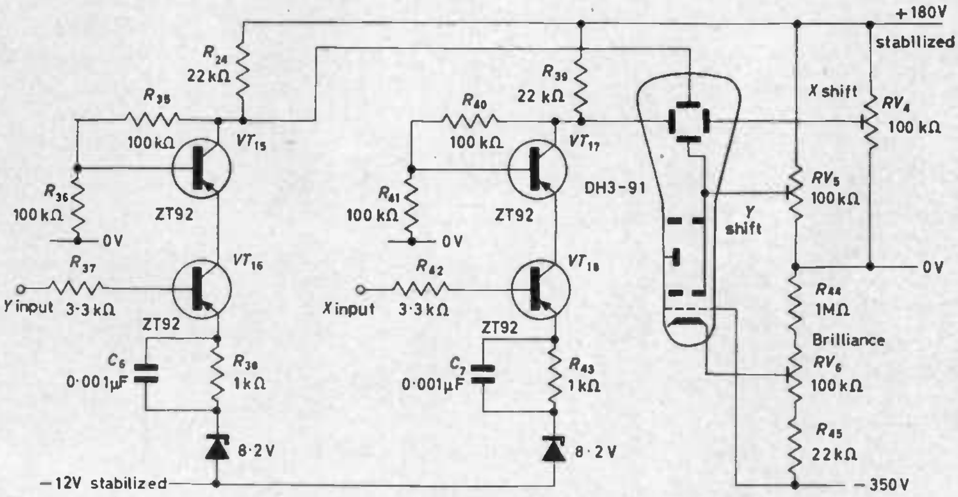



**Fig. 8. C.R.T. and deflexion amplifiers**

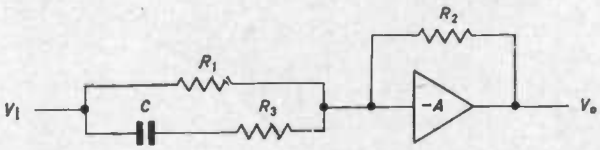



**Fig. 9. Proportional pulse derivative amplifier circuit**

compensation, thus improving the bandwidth of the amplifiers.

The c.r.t. has a one inch diameter face and has a final anode potential of approximately 500V, which gives a reasonable degree of focus and brilliance. The brightness of the c.r.t. trace is varied by altering the potential between grid and cathode by means of $RV_6$. The variable potentiometers $RV_5$ and $RV_4$ are the $X$ and $Y$ shift controls, used for centring the trace.

## Comments

The work carried out has shown that by employing electronic techniques pointer type displays can be made without the use of moving parts. This experimental model has been built to prove the feasibility of such a system and if further development is required a number of points require consideration. These are discussed below:

(a) *Brilliance*. The brightness of the trace is not adequate under high ambient light conditions as may be experienced in aircraft cockpits or instrumentation control rooms. This of course could be improved by using high brilliance tubes with pulsed bright-up and post deflexion acceleration, similar to that used for head-up displays.

(b) *Linearity* can be improved by the use of constant current charging networks in the delay modules.

(c) *Flyback*. The one big disadvantage that is difficult to overcome electronically (if the additional rate is required) is, that when a repetitive ramp function to define the pointer position is applied to the display to produce a rotating pointer, the 'flyback' part of the input waveform produces a 'kick' on the rate pointer, if the rate is derived by differentiating the input ramp signal. This may be overcome by producing the rate term external to the display unit. Another possible method would be to produce a blanking signal when the flyback occurs.

(d) *Power Supply*. The 400Hz supply would not usually be used, as a frequency shift would effect the calibration of the display. The variation in calibration is dependent on pointer position and may vary between 1 and 2°/Hz shift of the 400Hz supply.

Thus it would be preferable to use an internal oscillator with temperature compensation that would offset any variations of the output of the delay modules.

In general the display can be used for a variety of jobs, such as an altimeter display as originally mentioned, though not necessarily aircraft, where the rate information is of benefit such as controlling machinery where a derivative demand using a display is required.

Although the possibility of driving the display from digital information only has not been fully investigated, it is clear that the display electronics could be modified to enable input digital signals to be decoded by means of 'weighted' resistors in the feedback path of an operational amplifier. From this amplifier the proportional and rate terms may be obtained. Alternatively the rate term may be received separately in digital form and then decoded. For the numerator tubes a simple diode decoding matrix would be used.

**Acknowledgment**

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## Appendix

### Analysis of Proportional plus Derivative Circuit

With reference to Fig. 9 the overall transfer function of the circuit will be derived as follows:

$$\overline{V}_o = \frac{R_2}{\dfrac{R_1(R_3 + (1/pC))}{R_1 + R_3 + (1/pC)}} \, . \, V_i$$

where $p$ is the Laplace transform operator

$$\overline{V}_o = (R_2/R_1) \, . \, \frac{(R_1 + R_3 + (1/pC))}{(R_3 + (1/pC))} \, . \, V_i$$

$$\overline{V}_o = (R_2/R_1) \, . \left[ 1 + \frac{R_1}{R_3 + (1/pC)} \right] . \, V_i$$

$$\overline{V}_o = (R_2/R_1) \, . \left[ 1 + \frac{pR_1C}{1 + pCR_3} \right] . \, V_i$$

Let $V_i$ be a ramp input of $Kt$ volts/sec.

Therefore:

$$\overline{V}_i = K/p^2$$

$$\overline{V}_o = (R_2/R_1) \left[ K/p^2 + \frac{KR_1C}{p(1 + pCR_3)} \right] \quad \ldots\ldots (1)$$

Performing an inverse transform:

$$V_o(t) = R_2/R_1 \, [Kt + KT(1 - \exp(-t/T_1))] \quad \ldots. (2)$$

where $T_1 = CR_3$ and $T = CR_1$.

As can be seen in equation (2), the output $V_o$ consists of two terms: a ramp $(R_2/R_1)Kt$ volts/sec plus a derivative term $(R_2/R_1)KT$ volts $(= CR_2K)$ which has a lag time-constant of $T_1$ seconds.

# A Simple Digital-Analogue Convertor with Reciprocal Read-out

By L. Davison*, B.Sc., A.Inst.P., and R. Wilson*, B.Eng., A.M.I.Min.E.

*The convertor described provides a convenient means of direct velocity read-out from digital counters whose contents represent transit times. In addition to saving time when large numbers of results are involved, the convertor eliminates the possibility of errors due to misreading of counters, read-out lamp failures and computational mistakes.*

(Voir page 408 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

EXPERIMENTS which are in progress at the Safety in Mines Research Establishment with a coal-dust explosion tube 5m long require the measurement of flame-propagation speeds over 10 equal intervals of tube length. Photocells sited at the ends of the intervals detect the arrival of the flame-front and gate a reference frequency signal to 10 binary counters. Each counter has a capacity of 10 digits, and the minimum resolution required is about 1 per cent.

In a long programme of experiments it was tedious to read all the counters, convert from binary to decimal, and multiply the reciprocal of the count by an appropriate scaling constant in order to obtain the required speed results. Apart from the tedium of the process there was the possibility of visual read-out and conversion errors.

The present digital to analogue convertor is designed to be switch-selected to each counter in turn, with read-out and conversion performed automatically and the result displayed on a milliammeter scaled to read directly in metres per second.

### Basic Circuit

Fig. 1 shows the basic circuit. A binary counter which is to be read out is represented by the chain of blocks in the upper part of the diagram. Each counter stage, when filled, operates its own relay (*A*, *B*, *C*, etc.) and removes the short-circuit from an associated resistor ($R_1$, $R_2$, $R_3$, etc.) in the series chain across the constant-voltage supply *V*. The resistance values are so chosen that the total chain resistance is proportional to the decimal number which is represented by the counter state. The meter, *A*, which measures the chain-current may then be scaled to give direct read-out of flame-speed, as described below.

For a decimal number, *n*, the corresponding flame-speed is

$$u = S/nT \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where *S* is the distance between photocells, and *T* is the period of the reference oscillator. The contribution of the $p^{th}$ binary stage (if filled) to the decimal count is $2^{p-1}$, and the value of the corresponding chain resistor is $2^{p-1}\,r$. Each stage that is filled operates its own relay (*A*, *B*, *C*, etc.) to remove the short-circuit from its corresponding resistor. The total chain resistance is then $\sum_{p=1}^{p=10} 2^{p-1}\,r = nr$, where only the *p* values corresponding to filled stages are used in the summation series. The chain current, *I*, is then given by:

$$I = V/nr \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

which has the same form as equation (1)

* *Safety in Mines Research Establishment, Ministry of Power.*

Thus if $S/T = kV/r$, where *k* is a scale factor,

$$u = kI \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

and meter *A* can be scaled to read directly.

### Alternative Circuit

Fig. 2 is a schematic circuit which is the dual of that in Fig. 1, and it is clear that this could also be used to achieve the desired result. In this circuit the binary stage switches, $2^0$, $2^1$, .... $2^{p-1}$, etc., would be closed for filled stages, and the corresponding conductances brought in. Thus the total conductance across the constant-current generator would be,

$$\sum_{p=1}^{p=10} 2^{p-1}\, g = ng.$$

The voltmeter would then read $v = I/ng$, which again is the same form as equation (1) with voltage corresponding to speed.

This arrangement would lend itself more readily to solid-state switching since the stage switches share a common supply rail. However, the design would have to take account of the voltage-drop across solid-state switches; also constant-voltage sources of high accuracy are in general

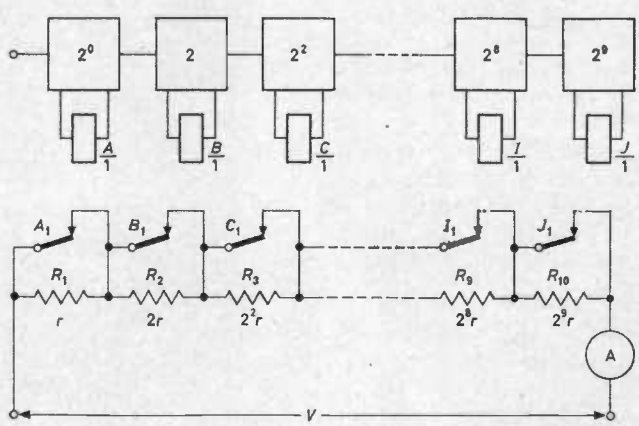
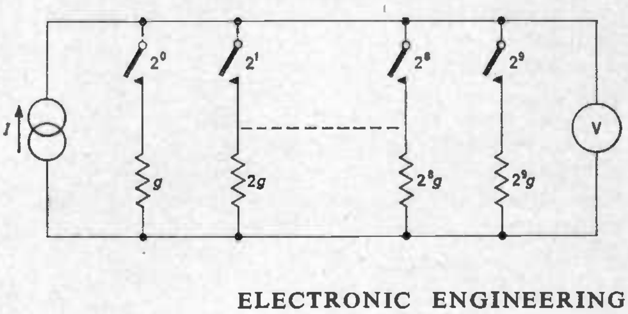



Fig. 1. Basic circuit

Fig. 2. Alternative circuit



more convenient to provide than constant-current sources.

Since a number of sensitive relays suitable for operating from the counter bistables were in any case readily available, the circuit of Fig. 1 was chosen.

## Design

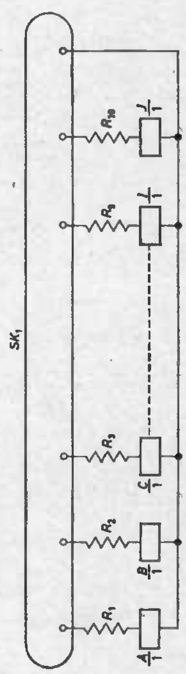
### Meter Resistance Error

Equation (3) takes no account of meter resistance. If, however, the chain voltage $V$ is increased to compensate for this at meter f.s.d., and if the meter f.s.d. voltage drop is $v$, then it may readily be seen that the error introduced, as a fraction of indicated speed value, is always less than $v/V$, and the error rises towards this maximum as zero deflexion is approached.

At f.s.d., the meter circuit used has a drop of 0·1V. For nominal chain supply of $V = 30$V, the maximum error introduced by fixed meter resistance is therefore 0·1/30, or ⅓ per cent. Since there is a range of switched shunts to provide different meter sensitivities, readings will not normally be taken at less than about 50 per cent f.s.d. and the actual worst error in practice will be approximately half the above value. It is recognized that this setting-up procedure yields an error always of the same sign, and in principle the maximum error could be reduced by setting up at something less than f.s.d., but since the precision of setting would require to be of the order of 1/12 per cent the distinction is academic, and setting at f.s.d. is convenient. For the sake of round figures in the chain resistance values the output of the constant-voltage supply is designed to be 30·1V; 30V for the chain and 0·1V for the f.s.d. meter drop.

### Resistance Chain

The current-sensitivity of the read-out meter dictates the value of $r$ in the resistor chain since it is preferable in the interests of reading accuracy for the deflexion obtainable for minimum chain current to be not much less than 50 per cent f.s.d. The 1mA f.s.d. meter offers reasonable mechanical robustness with a requirement for only moderately small chain currents. This latter feature offers the double advantage of keeping the chain wattage low and of relaxing the performance required from the constant-voltage supply.



If the minimum current is taken as 0·5mA (i.e. 50 per cent deflexion) for a count of 1 000, then $0{\cdot}5 \times 10^{-3} = V/1\,000r$,

and hence with $V = 30$, $r = 60\Omega$ .. (4)

### Meter Shunts

The count obtained for a given flame-speed is proportional to the counter reference frequency. Before any experiment, the anticipated flame-speed is estimated roughly, within range of 10 : 1. A reference frequency is then selected, from one of five available, in order to achieve a count within the limits of 100 and 1 000. Given this, it would then only be necessary to provide a simple shunt system to increase meter readability for high counts, provided the application of a scale factor according to frequency were accepted. The object of the instrument is however, to minimize

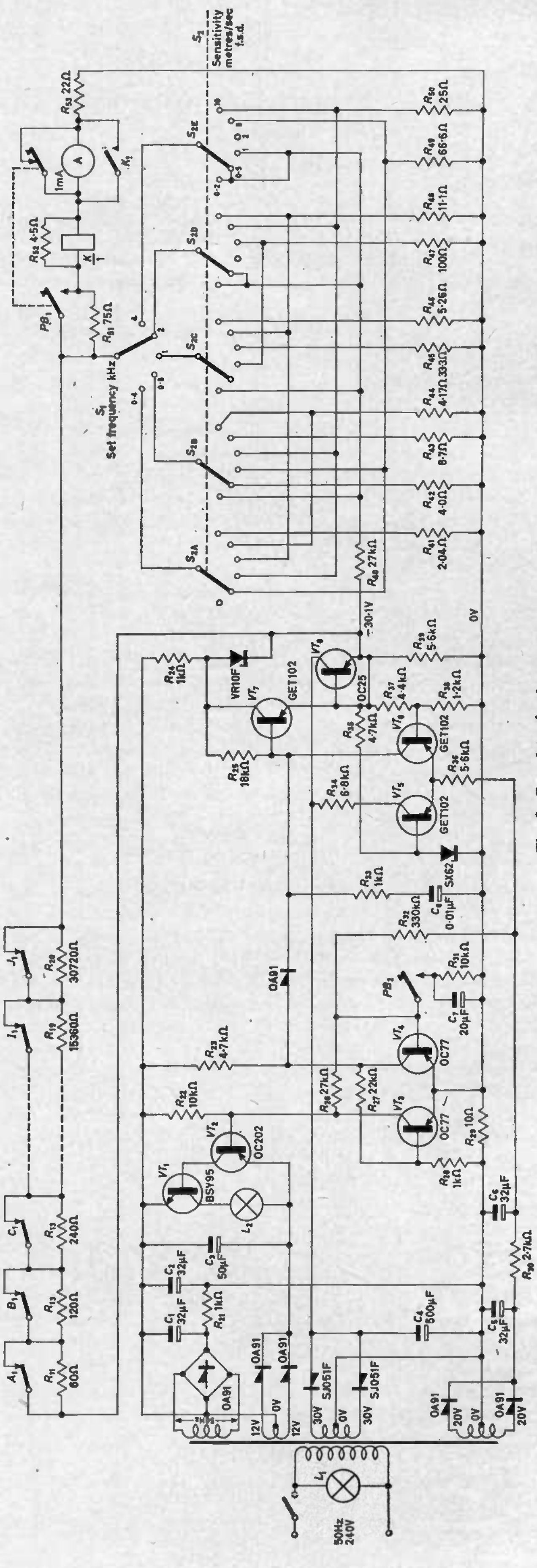


Fig. 3. Complete circuit

manipulation of readings in the conversion process, and it was therefore decided to provide the facility of direct read-out of speed from the meter by means of a switch ($S_1$ in Fig. 3) which is first set to select a group of sensitivity-multiplying shunts appropriate to the reference frequency used. An appropriate shunt within a set is then selected by $S_2$ to improve the readability, when necessary, of the meter indication. The corresponding f.s.d. speed is in that case shown by the pointer on the operating knob of $S_2$. The shunts permit a reading of not less than 40 per cent f.s.d. to be obtained for any speed yielding a count within the prescribed limits of 100 and 1 000.

For $r = 60\Omega$, and $V = 30$V, the count of 100 leads to a maximum chain current of $30/(60 \times 100) = 0{\cdot}005$A, and for a photocell separation of 0·25m, the corresponding speed is given by

$$u' = (0{\cdot}25f/100) = (f/400)\text{m/sec} \quad \ldots\ldots \quad (5)$$

where $f$ is the reference frequency (Hz).

Thus $u'$ is the f.s.d. speed in metres per second corresponding to the maximum chain current of 5mA. Since the meter circuit has a resistance of $100\Omega$, the shunt required for this case is $25\Omega$.

The shunt required for any other f.s.d. speed, $u$, may then be calculated as follows.

The chain current for $u$ is given by,

$$I = (u/u')I' \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

Putting $I' = 5$ (mA) and substituting for $u'$ from equation (5) in equation (6),

$$I = \frac{u}{(f/400)}(5) = 2\,000\,u/f \quad \ldots. \quad (7)$$

Also, since the f.s.d. meter drop is $1(\text{mA}) \times 100\,(\Omega) = 100$mV, and the shunt, $R$, associated with an f.s.d. current of $I$ is required to carry $(I - 1)$ mA, its resistance must be,

$$R = \frac{100}{I-1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

so that substituting for $I$ from equation (7) in equation (8),

$$R = \frac{100}{(2\,000\,u/f) - 1}\,.\,\Omega \quad \ldots\ldots\ldots\ldots \quad (9)$$

The available frequencies are 4, 2, 1, 0·8, 0·4 (kHz) and the f.s.d. speed ranges chosen are 10, 5, 2, 1, 0·5, 0·2 (m/sec). It is clear that if $u/f < 1/2\,000$, the denominator of equation (9) becomes negative, and $R$ cannot be chosen. This merely expresses the fact that there would be insufficient chain current to yield f.s.d. in the unshunted meter for counts which exceeded the specified maximum of 500, and such ranges as turn out to be inadmissible in the above set would therefore clearly be redundant. It would, however, be confusing for the user to have to remember, or consult a table to find, which ranges were permissible according to the reference frequency he was using. Arrangements are therefore made which automatically deflect the meter over full-scale from a separate current source when a non-permitted range is selected. An incorrect reading cannot therefore be taken, and since switching to non-permitted ranges corresponds to increasing meter sensitivity, no impression of discontinuity is conveyed to the observer.

In calculating the values of the shunts required from equation (9), it soon becomes evident that there are many duplications among the different frequency sets. Only 10 shunts are in fact necessary to cover the whole range of frequencies and sensitivities.

## Complete Circuit

The complete circuit is shown in Fig. 3.

A socket $SK_1$ connects relays $A$ to $J$ to a 10-pole 10-way switch (not shown) which routes the digital to analogue convertor to read out from each of the counters in turn. The basic chain consists of 1W resistors $R_{11}$ to $R_{20}$.

The circuit between the mains transformer, and $R_{20}$, including transistors $VT_5$ to $VT_8$, is a constant voltage circuit of conventional design providing, at the emitter of $VT_8$, the stabilized 30·1V for the resistor chain. Overload current protection is provided by the trip circuit $VT_3$, $VT_4$, triggered by over-voltage across $R_{20}$. This is necessary in the event of read-out being attempted from a counter with a very low content. When the trip level of 75mA chain current is exceeded, the bistable transition-signal at $VT_4$ collector drives $VT_7$ and hence $VT_8$ towards cut-off, thus reducing the chain voltage to a negligible value. At the same time, overload-indicator lamp $L_2$ is brought on via $VT_1$, $VT_2$. The trip circuit is reset by push-button $PB_2$.

Milliammeter $A$ is the speed indicator, calibrated 0 to 10 in 100 scale divisions. It is therefore direct-reading on the 10m/sec range. F.S.D. speeds for all ranges are indicated by a pointer-knob on $S_2$. $S_1$ is the reference frequency selector switch.

Meter protection is afforded by relay $K$, whose contacts $K_1$ short out the meter for currents exceeding 1·5mA. The need for protection could arise either because of high chain-currents or because the range switch has been inadvertently left at too high a sensitivity setting.

The relay is more tolerant of heavy overloads than is the meter, but its speed of response is of the same order. It is therefore necessary to ensure that the overload current is presented to the relay before the meter is brought into circuit. This is achieved by the 'press to read' button $PB_1$. In the event of over-current when the convertor is routed to a counter, the current flows through the meter limb via $R_{51}$ with the meter shorted out. Relay contacts $K_1$ then close so that the meter is still short-circuited even when $PB_1$ is pressed.

When $PB_1$ is operated under normal current conditions, one pole shorts out $R_{51}$ and the other removes the meter short-circuit. Since $R_{51}$ has the same resistance as the meter movement the total limb resistance is preserved. Without this feature, spurious lock-outs would occur because the meter movement represents the bulk of the limb resistance and the protective short would lead to significantly high limb-currents than would exist when the short was removed.

$R_{41}$ to $R_{50}$ are the meter shunts, and the over-deflexion current for the non-permitted ranges flows through $R_{40}$ from the 30·1V chain supply.

## Conclusion

Although based on a simple principle, the digital to analogue convertor described offers considerable advantages in terms of convenience and error-avoidance. It also eliminates the need for visual read-out facilities on the associated counters unless these are required for check or other purposes. This can represent a significant economy in components in multi-counter units.

Calibration showed the accuracy of the convertor to be well within that of the meter, which was within 1 per cent f.s.d.

**Acknowledgment**

The authors would like to acknowledge the assistance of Mr. B. Russell in the design and construction of the unit. 

# A Twin-T Filter Design having an Adjustable Centre Frequency

By K. G. Beauchamp*, C.Eng.

*A major difficulty in twin-T filter design where a range of frequencies has to be covered, is to provide a simple means of adjustment for frequency without, at the same time, drastically reducing the Q factor.*

*In the design described here variation of a single parameter gives control over frequency with only a small effect on the designed Q factor.*

(Voir page 408 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

TWIN-T filters are widely used to provide a bandpass or 'notch' filter of high $Q$ factor and using the minimum of components.

In order to maintain this high $Q$ at a particular frequency the critical adjustment of at least two elements of the filter becomes necessary. Where a range of frequencies is to be covered then three filter elements have to be varied simultaneously.

These difficulties can be obviated if an asymmetrical twin-T filter is used in which only a singe resistive element needs to be varied to secure a change in resonant frequency. A range of 3:1 in frequency is practically realizable.

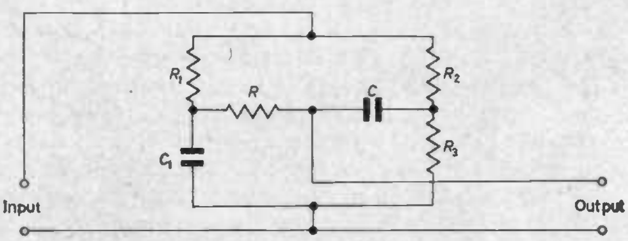
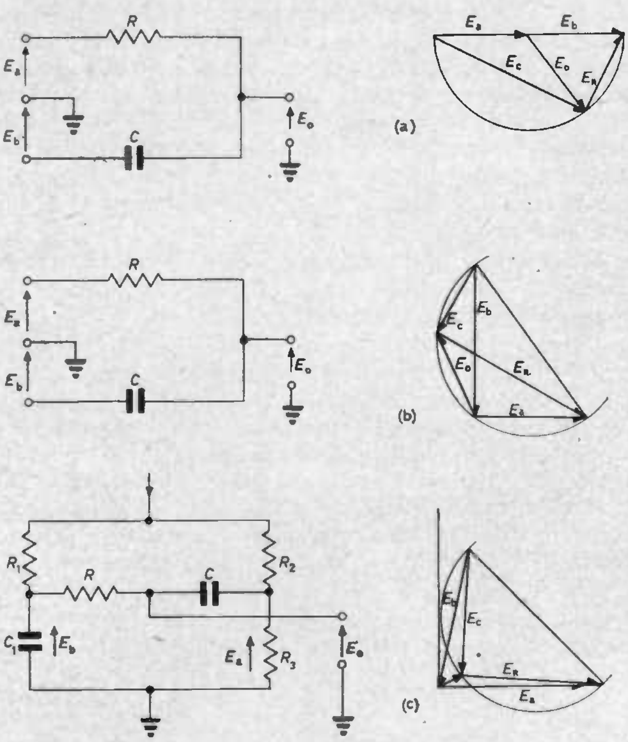



**Fig. 1. Phase discriminating network**

**Fig. 2. Derivation of the vector locus for the network**




## Derivation

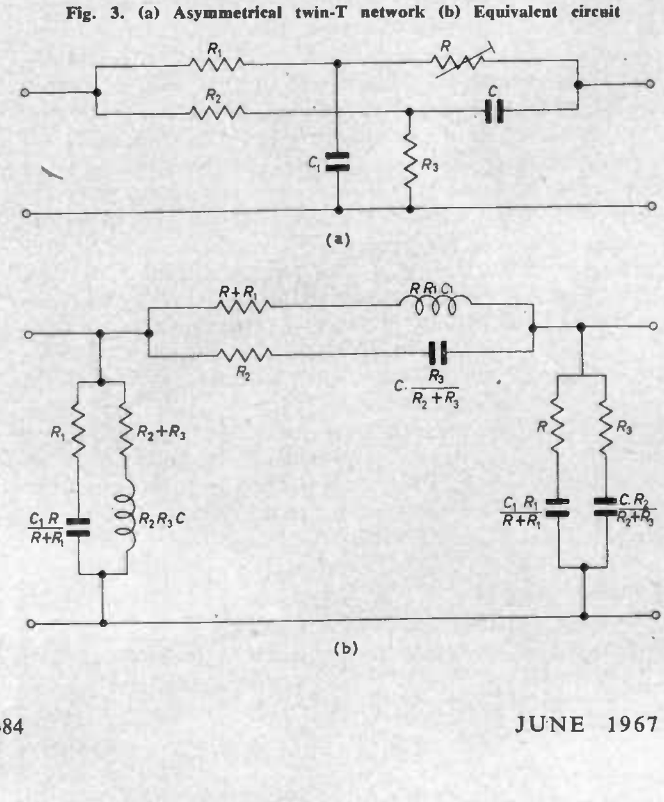
The circuit may be derived from the network shown in Fig. 1 and a simple description follows with reference to the vector diagrams given in Fig. 2.

Consider the simple phase-shift circuit of Fig. 2(a). This is a constant $-\alpha$ (all pass) network in which the locus of the $E_o$ vector describes a semicircular path as the frequency is varied.

Now if the input voltages $E_a$ and $E_b$ are given a relative phase difference of $\pi/2$ rad as in Fig. 2(b), then the semicircular locus of the $E_o - E_R$ junction passes through the origin and $E_o$ experiences a null at one particular frequency.

A practical network is shown in Fig. 2(c), where the $\pi/2$ phase shift is obtained from $C_1 R_1$ and a potentio-

**Fig. 3. (a) Asymmetrical twin-T network (b) Equivalent circuit**




* *United Kingdom Atomic Energy Authority.*

meter, $R_2 R_3$ included to ensure similar amplitudes for vectors $E_a$ and $E_b$.

The null in this case is not a true one since the phase shift for $E_b$ is slightly less than $\pi/2$ rad. This is not a disadvantage however, when the filter is used with an operational amplifier, since some gain-stabilizing feedback will be available at balance frequency. Variation of balance frequency can be made by adjustment of $C$ or $R$. The latter is more convenient and gives a smaller variation in $Q$ value over the frequency range.

The network of Fig. 1 can be redrawn as a twin-T network having asymmetrical arms (Fig. 3(a)). Converting each of the $T$ networks into its equivalent $\pi$ circuit results in the network shown in Fig. 3(b).

It will be seen that at one frequency $\omega_o/2\pi$, the series arm will behave as a parallel tuned circuit and exhibit a very high impedance. Since the shunt reactance arms of the circuit will have very little effect on the value of this frequency then it is permissible to equate the imaginary components of the series arm to derive the resonant frequency.

i.e.,

$$\omega_0^2 = \frac{R_2 + R_3}{C \, . \, R \, . \, C_1 \, . \, R_1 \, . \, R_3} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \text{(1)}$$

Expressing this in terms of time-constants $T = CR$, $T_1 = C_1 R_1$, and a reduction factor $a = R_3/(R_2+R_3)$ gives:

$$f_o = \frac{1}{2\pi \sqrt{(T \, . \, T_1 \, . \, a)}} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \text{(2)}$$

and the $Q$ factor is given as:

$$Q = \frac{\omega_o R R_1 C_1}{R_1 + R_2 + R} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \text{(3)}$$

Substitution of equation (2) in equation (3) gives:

$$Q = \frac{R}{R + R_1 + R_2} \sqrt{(T_1/T \, . \, a)} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \text{(4)}$$

**Filter Design**

From equation (4) it is seen that the $Q$ value is determined largely by the numerical value of $\sqrt{(T_1/T \, . \, a)}$ and

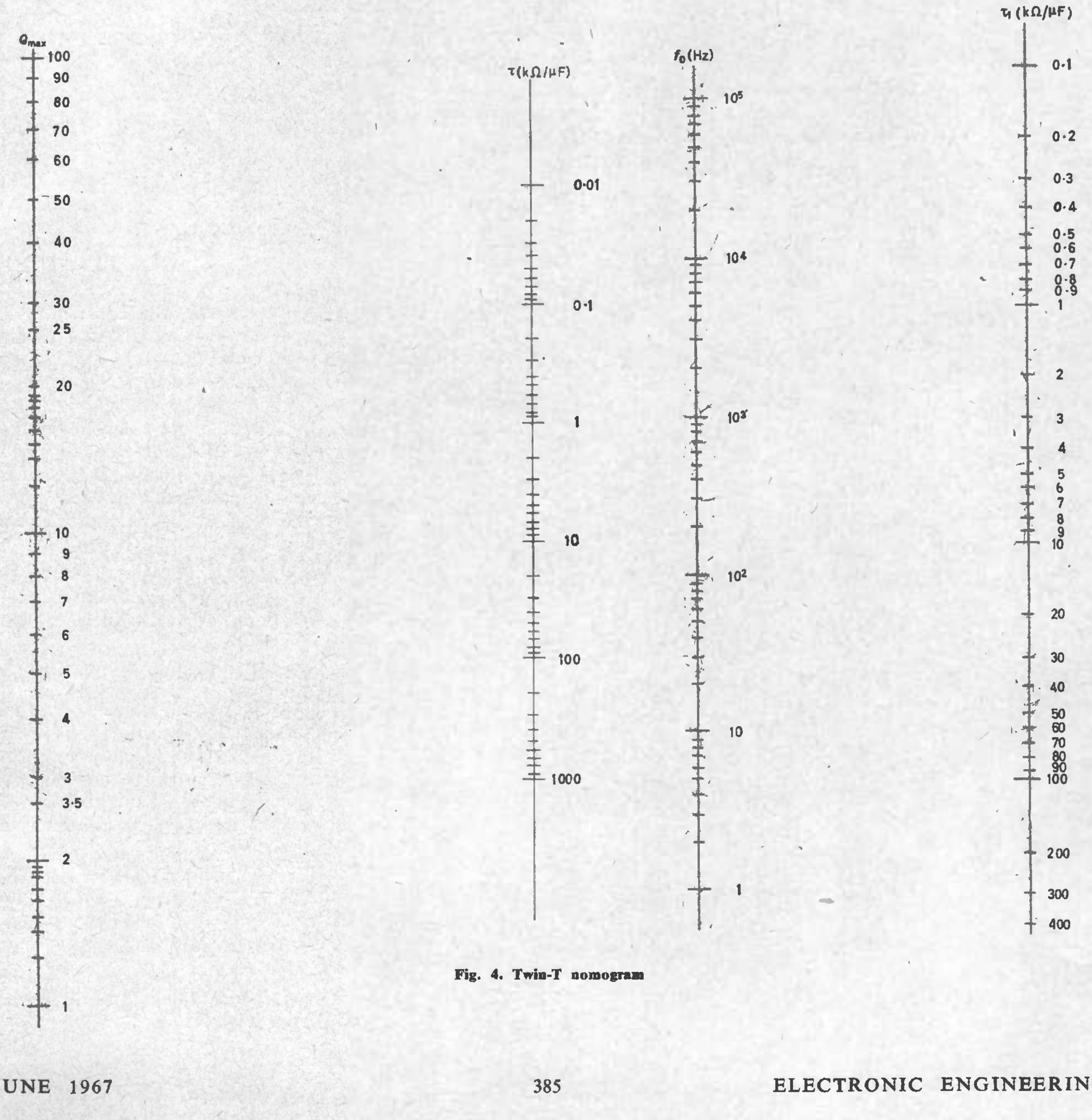



Fig. 4. Twin-T nomogram

if $R \gg R_1 + R_2$ then:

$$Q_{max} \simeq \sqrt{(T_1/T\,.\,a)} \quad \ldots\ldots\ldots\ldots\ (5)$$

which is a useful figure for design purposes.

If a small value is chosen for $a$, say 0·01, then equation (4) can be substituted in equation (2) to obtain a simple relationship between $Q_{max}$, $T$, and $f_o$.

$$T = (16000/f_oQ_{max})\,.\,\mathrm{k}\Omega\,\mu\mathrm{F} \quad \ldots\ldots\ (6)$$

This can be translated into the nomogram form shown in Fig. 4 to facilitate calculation where a number of filters are to be designed.

This nomogram can also give a value for $T_1$ since this is implicit in the relationship given in equation (5).

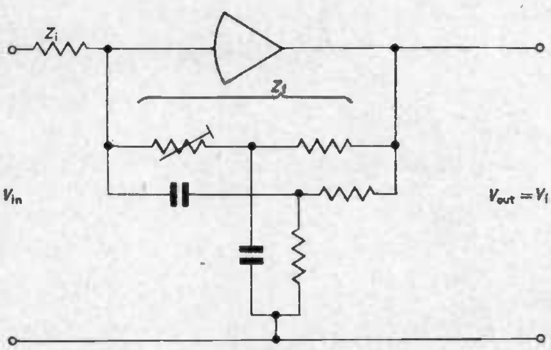
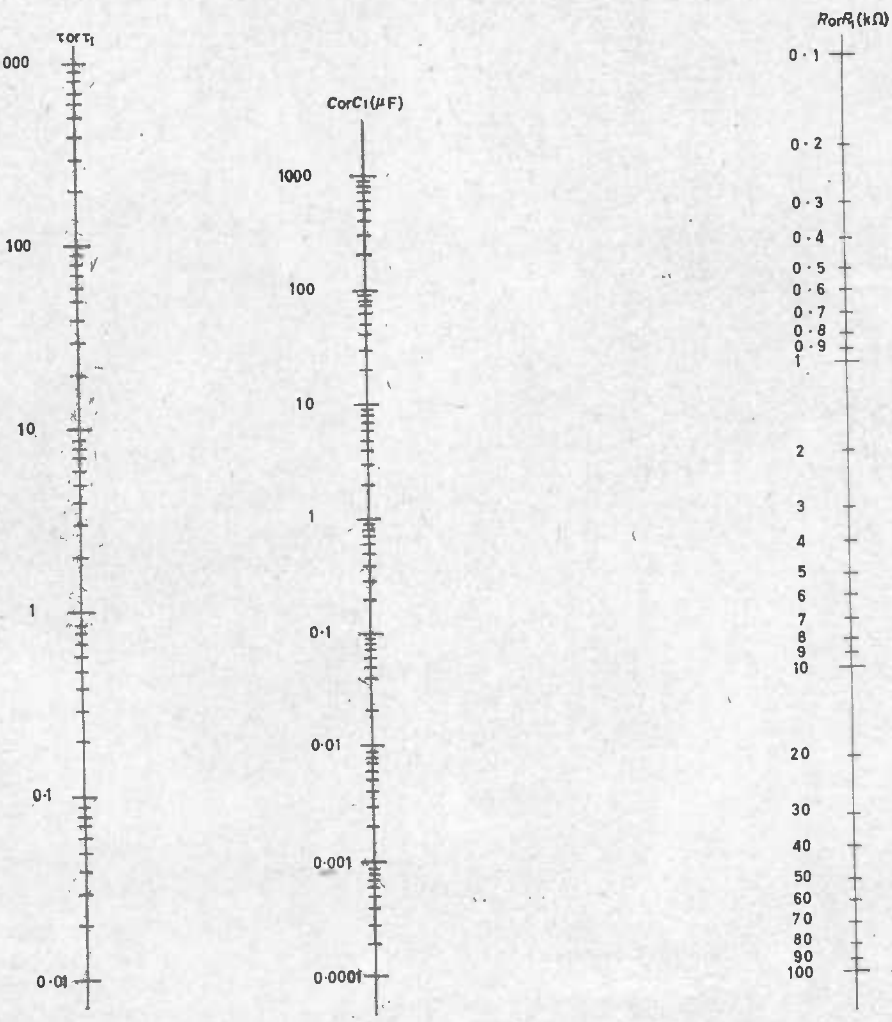



**Fig. 5. Use of twin-T in operational amplifier circuit**

**Fig. 6. Time-constant evaluation**




Design is further simplified if certain assumptions are made regarding the values of the resistive components. Making $R_1 = R_2$ and also, due to the small value of $a$ chosen above, then $R_3 \simeq aR_2 = 0{\cdot}01R_2$.

The value of $R$ is adjusted to vary the resonant frequency of the filter. Its value also has an effect on the stability of the circuit when the filter is used as the control impedance of an operational amplifier (see Fig. 5). It is possible for this circuit to act as a phase-shift oscillator with the second stage of $\pi/2$ phase shift being provided by the value of $R$ and the stray capacitance at the summing junction. This limits the value of $R$ in practical terms to below 100kΩ for $Q$ values up to about 50 in this type of application. Choice of $R_1$ is related to the need to keep the $Q$ factor as constant as possible throughout the desired frequency range. A practical limit to this range is about 3 : 1 which corresponds to a resistance value change of about 10 : 1 (equation (2)).

Under these conditions it can be shown that a choice of $R_1$ equal to the minimum value of $R$ will give similar values of $Q$ at either end of the range.

Thus from equation (4) at minimum $R$ value, $R_{min}$

$$Q_1 = \frac{R_{min}}{3R_{min}}\sqrt{\left(\frac{T_1}{CR_{min}a}\right)}$$

and at maximum $R$ value, $R_{max} = 10R_{min}$

$$Q_2 = \frac{R_{max}}{R_{max} + 2R_{min}}\sqrt{\left(\frac{T_1}{CR_{max}a}\right)}$$

$$= (10/12)\,(1/\sqrt{10})\sqrt{\left(\frac{T_1}{CR_{min}a}\right)}$$

so that the ratio of $Q_1/Q_2$ becomes

$$Q_1/Q_2 = 0{\cdot}33/0{\cdot}27 = 1{\cdot}2$$

Choice of time-constant component values for $T$ and $T_1$ is facilitated by means of the second nomogram of Fig. 6 which relates $T$ to $C$ and $R$ and also $T_1$ to $C_1$ and $R_1$. Having calculated $T$ and $T_1$ using the nomogram of Fig. 4, and chosen a suitable value for $R$ (and hence $R_1$) the values of $C$ and $C_1$ can be determined from Fig. 6.

Choice of $R_1$ also allows $R_2$ and $R_3$ to be obtained from the fixed relations already quoted.

### A Design Example

A filter is to be incorporated in an operational amplifier as shown in Fig. 5. It is to cover the range of 400 to 1 000Hz and a 30kΩ potentiometer is specified for $R$. The desired $Q$ factor is about 20.

The maximum value of $R$ will correspond to the lowest frequency required and from Fig. 4, taking $Q = 20$ and $f_o = 400$Hz a figure is derived of $T = 2$ and $T_1 = 8{\cdot}5$.

Given $R = 30$kΩ and $T = 2$ the capacitance value for $C$ is 0·06μF from Fig. 6.

From Fig. 4 it is found that given $T_1 = 8{\cdot}5$ and the maximum required frequency of $f_o = 1\text{kHz/s}$ then the new value of $T = 0{\cdot}3$, which corresponds to a minimum value of 4·7kΩ for $R$ with $C = 0{\cdot}06\mu\text{F}$. Choosing this as a value for $R_1 = R_3$ gives a value of 2μF for $C_1$ from Fig. 6. This also fixes a value of 47Ω for $R_4$.

A practical filter, designed on this basis could be tuned from 400 to 1 000Hz with potentiometer values of 27kΩ down to 4·3kΩ and the $Q$ factor varied between 16 and 28 over this frequency range.

**Acknowledgment**

The author wishes to thank the Director of the Atomic Weapons Research Establishment for permission to publish this article.

## A New Colour Television Picture Tube

The Compagnie Française de Télévision (CFT), the originators of the SECAM colour television system, have developed a new colour television picture tube which, it is claimed, has an electron transparency of 80-90 per cent, as compared with 15-20 per cent of the conventional shadow mask tube.

In addition this new picture tube operates on a lower modulating voltage and therefore permits an increase of transistorized circuits thus reducing overall cost.

As with the shadow mask tube, the new CFT tube can be used with any of the existing colour television systems and can also, of course, produce a monochrome picture.

The fundamental features of the new CFT tube are that it replaces the mask by a grid of parallel wires, and the colour dots on the shadow mask tube by a series of strips of colour phosphors.

The tube has a rectangular screen with a 19in diagonal.

The luminescent screen is deposited on a glass plate and consists of a series of straight luminescent strips positioned vertically each with a constant width of 0·27mm. These strips are placed edge to edge and three consecutive strips form a group which corresponds to the three primary colours (green, blue and red). There are 480 groups of such parallel strips each 0·81mm wide. The luminescent screen is covered with an aluminium film on which a layer of porous graphite is fixed. This layer of graphite is designed to decrease the factor of secondary emission from the aluminium-fluorescent screen assembly.

The grid is formed by a curtain of 550 parallel wires which are also parallel to the luminescent strips. The wires of 0·1mm diameter are spaced 0·75mm apart.

The electron sources for each of the three primary colours are three electron guns which are angled with respect to each other such that their beams converge along the tube axis close to the grid.

### Principle of Operation

#### Post-focusing and Post-acceleration

A set of cylindrical electrostatic lenses, which focus the beams from the electron guns, is formed by the grid and the screen as the grid is taken to a lower potential than that of the screen. When the beams strike the screen, slightly elliptical spots are formed which have their major axes parallel to the luminescent strips. Owing to the focusing, the width of the spot from each electron gun is made smaller than the width of a colour strip. The screen-grid potential difference introduces post-acceleration which ensures high luminosity.

#### Colour Separation

As the electron guns converge between the grid wires, the beams have separate striking points on the screen. The width of the screen strips, the grid-to-screen distance, the applied voltages and the inclination of the electron guns have been chosen such that the three striking points are located on three strips of different colours. The electrons from the electron gun for a given colour can only strike that colour.

The magnetic fields of four permanent magnets, which can be adjusted positionally, enable the juxtaposition of the spots at the centre of the screen to be set up. Positioning the spots on the axis of the respective luminescent strips is achieved by the action of a constant magnetic field between the exit of the electron guns and the deflexion system.

#### Dynamic Deflexion and Convergence

The main deflexion assembly ensures the general displacement of the three beams over the tube screen.

Proper convergence of the spots over the whole surface of the screen is achieved by the action of the variable magnetic fields applied to the beams emitted from the guns. These fields are generated by a convergence correction system which comprises pole pieces inside the tube and electromagnets outside. Voltage waveforms at the appropriate line and picture scan frequencies are applied to these electromagnets.

#### Colour Purity

The tube has been designed so that the three colours are pure over most of the screen. Colour distortion in the peripheral areas is corrected by the electrostatic field of a peripheral electrode painted on the internal wall of the glass bulb. The effect of the earth's field is corrected by an axial field coil in the same plane as the screen.

The luminous lines which appear on the screen are sufficiently near to each other for it to be difficult to distinguish them. Picture blending is satisfactory with a triplet pitch of 0·81mm.

#### Operational Example

| | |
|---|---|
| Screen voltage (e.h.t.) | 25kV |
| Chromatic purity electrode voltage | 10kV |
| Cone voltage | 8·600kV |
| Focusing grid voltage | 7·1kV |
| Gun focusing electrode voltage | 2·5kV |
| Gun first anode voltage | 400V to 500V |
| Cut-off voltage | −50V |
| Modulation voltage (for a total screen current of 100μA from the three guns) | 15V |
| Brightness measured over a white area of 395×310mm (for a total screen current of 100μA from the three guns) | 15ft.L. |
| Useful screen surface | 1150cm² |

# An Accurate Triangular-Wave Generator with Large Frequency Sweep


By G. Klein* and H. Hagenbeuk*

*A triangular-wave generator is described which combines a wide frequency sweep ($\simeq 10^4$) with a very accurate symmetrical waveform. The amplitude response and the symmetry exhibit variations of less than 0·1 per cent, while the frequency stability is better than 1:10⁴. The circuit can serve as the basis for a single-sweep sine-wave oscillator or an f.m. modulator and voltage (current)-frequency convertor.*

(Voir page 408 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 409)

MANY applications require a generator of symmetrical triangular voltages, the frequency of which can be varied over a wide range by an external voltage or current. The stringency of the requirements made on the accuracy of the amplitude and the symmetry of the triangular waveform depend on the use. For most f.m. applications, these requirements are not particularly rigid. If, on the other hand, a single-sweep sine wave oscillator is to be designed by combining such a triangular-wave generator with an instantaneous triangle-sine convertor, the distortion in the sine-wave voltage will be determined by the symmetry and amplitude stability of the triangle. In order to be most universally applicable a generator was therefore designed to a very stringent specification.

**Principle**

The starting point for such a generator was a circuit that has been in use for a considerable time as an f.m.-modulator and already satisfies rigid requirements in many respects. Its principle is given in Fig. 1. The Schmitt-trigger is used to make the difference between the base-voltages of $VT_1$ and $VT_2$ alternately positive and negative by a few volts so that $VT_2$ carries either a current $I_1$ or no current. By this means the capacitor $C$ is alternately charged with a current $I_2$ and discharged with a current $I_1 - I_2$, provided that $I_1 > I_2$. The amplitude constancy of the triangular waveform is determined by the difference between the changeover levels $V_H$ and $V_L$ of the Schmitt-trigger (about +2 and −6V in the example) which is, in turn, primarily determined by the resistance values and the supply voltages in the Schmitt-trigger, thus allowing considerable accuracy to be attained. The frequency can be varied over a wide range by changing $I_1$ but the triangular waveform is not symmetrical because $I_2$ is maintained constant. For the system to work properly, $I_2$ would have to vary simultaneously and satisfy the requirement $I_1 = 2I_2$, a condition which would be hard to be maintained for large variations of $I_1$. Fig. 2 gives a possible improvment which has been in use for some years†. Transistor $VT_2$ of the balanced pair carries the current $I_1$ in one position of the Schmitt-trigger. Since $VT_1$ then carries no current, the base voltage of $VT_3$ will be equal to the positive supply voltage, so that $I_2$ is then zero. Therefore, in this position, $C$ is discharged by the current $I_1$. In the other position of the Schmitt-trigger, $VT_1$ carries the current $I_1$ and the base voltage of $VT_3$ will be

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† *J. J. Zaalberg van Zelst: private communication.*

about $I_1R$ volts negative with respect to the positive supply voltage. If the emitter resistance of $VT_3$ is correctly chosen, $I_2$ can be made equal to $I_1$. The symmetry of the triangular wave which can be made perfect at a given value of $I_1$, is difficult to maintain if $I_1$ is varied over more than one decade because the base-emitter voltage of $VT_3$ will not remain equal.

In the proposal put forward here, care is taken to keep the current $I_2$ accurately equal to $I_1$, even when the latter varies greatly. Fig. 3 shows how this can be done by making use of a second capacitor $C_2$. Here, switching transistors $VT_1$ and $VT_2$ are symbolically represented by switches $S_1$ and $S_2$. These switches, together with $S_3$ and $S_4$, are controlled by the Schmitt-trigger as before in such a way that $S_1$ and $S_4$ are closed when $S_2$ and $S_3$ are open, and vice versa.

The basic principle of the circuit is the use of the voltage on capacitor $C_2$ to control the current $I_2$ in such a way that the average voltage across $C_2$ is kept constant.

If $T_c$ and $T_d$ are the times during which $S_1$ and $S_4$ ($S_2$, $S_3$) are closed (open) and open (closed) respectively, the following equalities should apply:

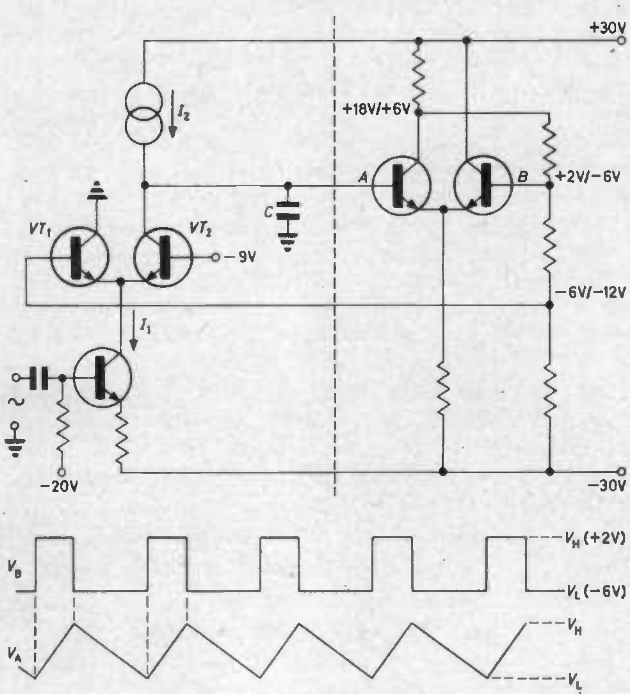
for $C_1$: $I_1T_d = I_2T_c$
and for $C_2$: $I_1T_c = I_2T_d$.

Hence:

$I_1 = I_2$ and $T_c = T_d$ as required.

The extent to which these equalities will apply in practice depends on the accuracy with which the currents fed to $C_1$ and $C_2$ are equal and this is determined by the leakage currents of the switching transistors and the base currents of the input stages of the Schmitt-trigger and the control amplifier. The inequality between the currents can be kept down to about $10^{-9}$A in a transistorized circuit. This means that, if the currents $I_1$ and $I_2$ are no smaller than a few microamperes, the currents satisfy the above equations to within about 0·1 per cent. On the other hand, the currents should not be made greater than about 10mA. This means that where suitable current sources are used, a frequency sweep of $10^4$ can be obtained. Furthermore,

**Fig. 1. Basic circuit for triangular-wave generator**



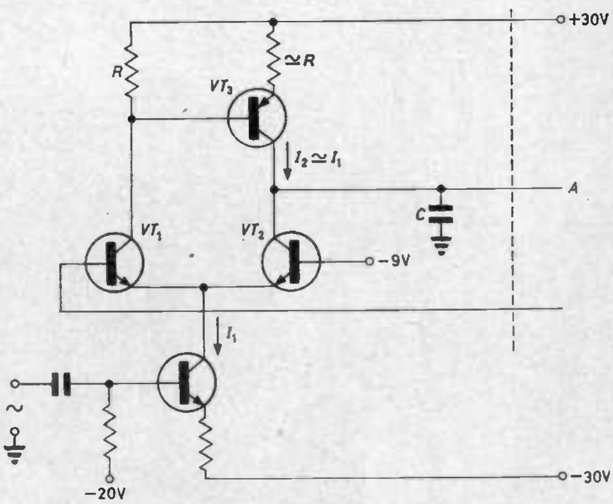




**Fig. 2. Possible regulation of symmetry**

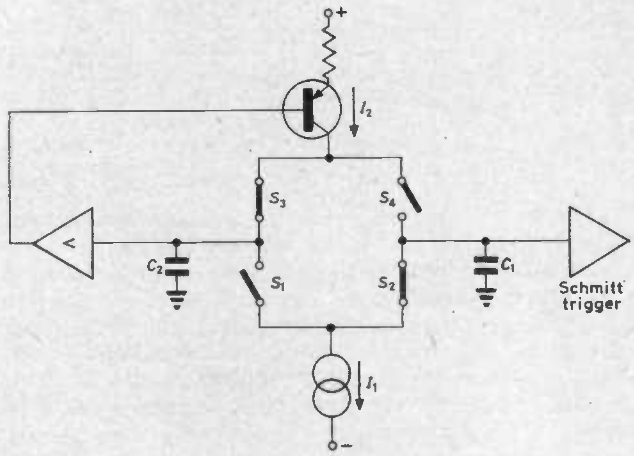



**Fig. 3. Proposed regulation of symmetry**

less than 0·1 per cent asymmetry can be guaranteed. With a reduced frequency sweep, higher minimum values of $I_1$ and $I_2$ may be used and the symmetry will be even better.

The principle as given in Fig. 3 has only one disadvantage: Since $I_2$ is controlled by the ripple voltage on $C_2$ as it charges $C_1$, the rising ramp of the triangular voltage on $C_1$ will exhibit a slight deviation from linearity. This effect is reduced by increasing $C_2$ and/or decreasing the loop gain. This means, however, that the speed with which $I_2$ follows a change in $I_1$ is reduced. A simple calculation shows that for this the following relation applies:

$$\tau = T/8\delta \quad (1)$$

where $T$ = period of the triangular voltage,

$\tau$ = time-constant of the control system of $I_2$,

$\delta$ = maximum relative deviation from linearity in the rising flank of the triangular voltage.

Thus, to limit $\delta$ to 0·1 per cent, the control would require more than 100 periods.

In the case of an external variation of $I_1$, the other current may be varied simultaneously by approximately the same amount. The control system would then only have to correct a possible deviation, and this could normally be allowed to take some time.

A better solution is given in Fig. 4. The ripple in the control voltage can be made zero by making $C_2$ equal to

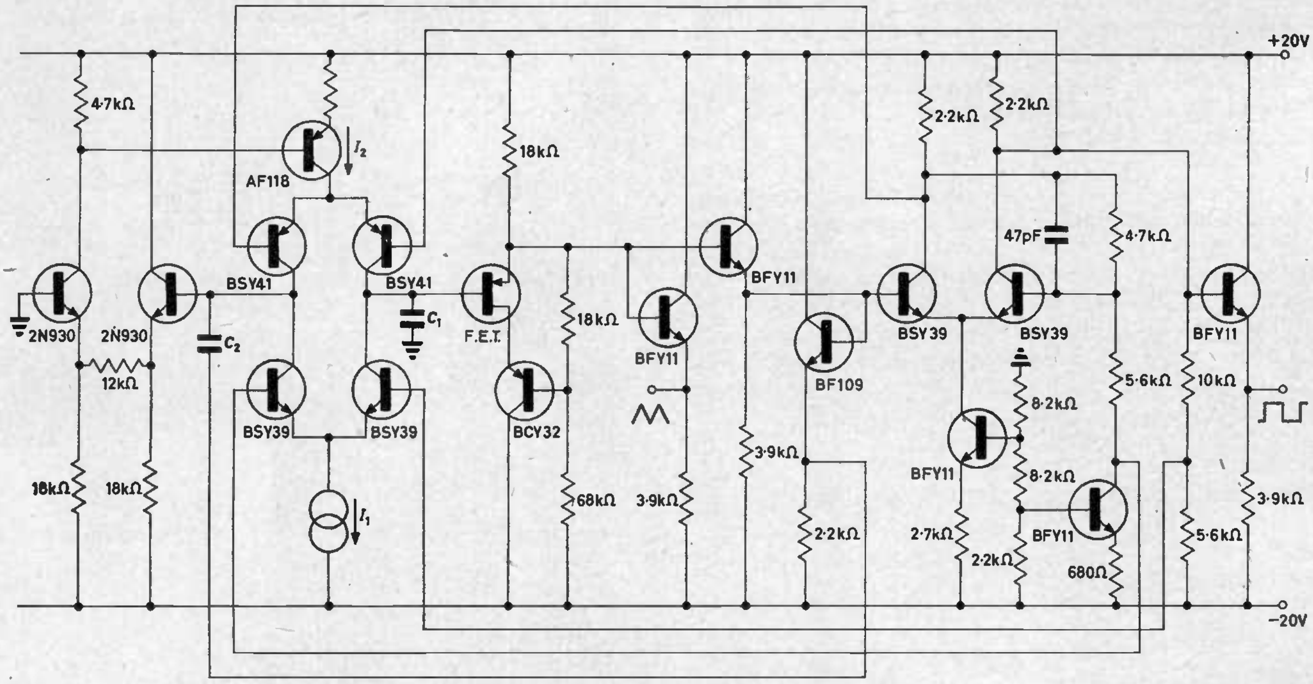



Fig. 5. Practical circuit

$C_1$ and by connecting one side of $C_2$ to a point that follows the voltage on $C_1$. In theory, the non-linearity and the control speed are now no longer related. If the adjustment is perfect, the ripple is zero and no distortion occurs. In practice, however, this perfect adjustment cannot be relied upon as deviations in the values of the capacitors will occur. The ripple voltage can, however, easily be compensated to within 1 per cent. Hence, the denominator in equation (1) is multiplied by 100 and this is adequate for all practical purposes.

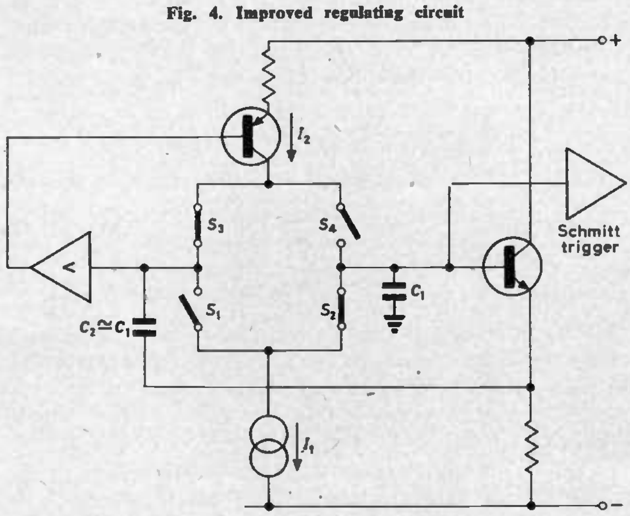
### Practical Circuit

Fig. 5 gives a simplified diagram of a triangular wave-generator incorporating the principles given above. Where necessary, the transistors were arranged as Darlington pairs. Since only currents are switched, the transistors may be replaced, with advantage, by field-effect transistors.

Various current sources may be used for $I_1$, thus providing, for instance, linear or exponential relationships between the frequency and a voltage which may be controlled externally. $I_1$ may also be supplied by two or more current sources connected in parallel. Where the switching circuit is designed with Darlington pairs the asymmetry proved to be much smaller than 0·1% for frequencies below some tens of kilohertz. At higher frequencies an inaccuracy will be introduced by the finite switching time and by parasitic effects of the switching circuit. With the available h.f.-transistors very good symmetry is still possible for frequencies up to some hundreds of kiloherty.

The frequency stability is determined by the constancy of the current source $I_1$ and the switching level of the Schmitt-trigger. At normal ambient temperatures a frequency stability better than $10^{-4}$ was easily obtained for linearly variable current sources. Using sources with exponentially varying currents, the stability is generally less by one order of magnitude. The influence of temperature changes can be kept below 0·01 per cent/°C without difficulty. The amplitude is also determined by the constancy of the switching levels of the Schmitt-trigger.

Fig. 4. Improved regulating circuit




The amplitude constancy with changing frequency is perfect as long as the transit times of the switching circuit are negligible. Due to this effect some increase in amplitude will occur at higher frequencies. It is obvious that the application of high-frequency field effect devices for the switches will give still better results.

The instrument shown in the photograph on page 388 is the first commercially available generator in which the described circuits have been applied. The main characteristics of this function generator—the Philips type PM 5162—are :

*Frequency range*: 0·1Hz to 100kHz, divided in three;

*Ranges*: 0·1 to $10^3$, 1 to $10^4$ and 10 to $10^5$Hz;

*Waveforms*: sine, triangular, square;

*Output voltage*: 3V peak-to-peak into 600Ω.

*Sweep mode*: internal and external with a maximum sweep ratio of 1 : 10 000.

# LETTERS TO THE EDITOR

*(We do not hold ourselves responsible for the opinions of our correspondents)*

## A Novel Voltage Reference Source

DEAR SIR,—We would like to make a few comments on the article by Messrs.

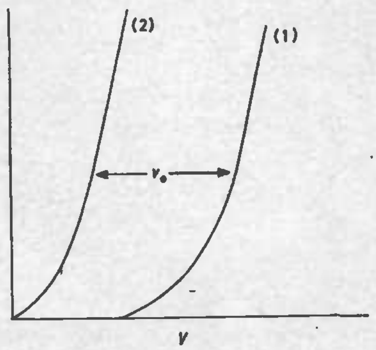


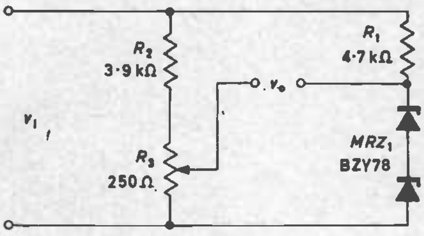

**Fig. 1. Characteristics of two arms of bridge**

Dawson & Taylor on "A Novel Voltage Reference Source," (February issue). The authors refer to "constant current type" and "constant voltage type" devices, but, in fact, the bridge will not operate with such devices. A constant voltage device has a low incremental resistance while a constant current device must have a high incremental resistance but, in fact, (as stated) the incremental resistance of the four arms must be the same. This is in contradiction to the definition of constant voltage and constant current devices.

The bridge operates due to the fact that the characteristics of the two arms of the bridge are as in Fig. 1 i.e. they have the same slope or incremental resistance but are displaced in terms of voltage, the output voltage being due to this displacement. Due to the fact that the characteristics are parallel (for a portion only) the output is independent of the current and hence voltage input.

The transistor arms are not constant current devices and do not operate as suggested. The transistor incremental resistance may be of the order of 30 000Ω, but it is in parallel with the 1·2kΩ resistor and Zener incremental resistance. The latter can be neglected and since the effect of 30 000Ω in parallel with 1·2kΩ is negligible these bridge arms act as constant voltage sources of 4·3V in series with 1·2kΩ (i.e. constant voltage sources substantially) as regards incremental changes. It would thus appear that the transistors could be dispensed with. When batteries are used in place of the Zener diodes the 1·2kΩ resistors are presumably omitted and hence the need for shunt resistors across emitter and collector to reduce the slope resistance to the order of 1kΩ.

**Fig. 2. Diode resistance compensating circuit**




For the complexity of the circuit its performance is not outstanding. The incremental resistance of the Zener diode can largely be compensated by the circuit of Fig. 2. The voltage drop

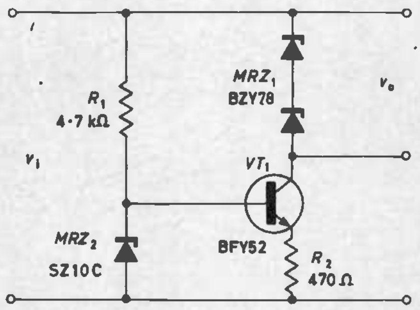



**Fig. 3. Constant current circuit**

across $R_2$ is made such that it is the same as the drop across the incremental resistance of the Zener diodes $MRZ_1$. With a variation of input voltage from 25 to 35V the output is constant within ±7mV. The circuit suffers the disadvantage (and that of the article suffers the same disadvantage) that there is a large change of current in the diode and hence there are changes in diode temperature and the compensation required is different for fast and slow changes.

A better circuit is shown in Fig. 3 which maintains an almost constant current in the Zener diodes $MRZ_1$. Zener diode $MRZ_2$ provides an approximately constant voltage and hence a constant current flows in $R_2$ as the transistor $VT_1$ acts as an emitter-follower. Thus an approximate constant

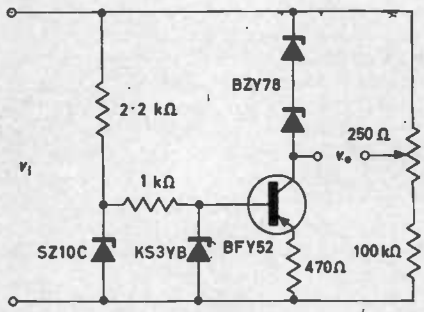



**Fig. 4. Final circuit arrangement**

current flows through the Zener diodes $MRZ_1$. In this case the transistor $VT_1$ acts as a constant current device.

The output of this circuit is also constant within ±7mV for an input change from 25 to 35V. This is reduced to ±2mV by using two Zener diodes in cascade to feed the transistor base.

The final circuit is shown in Fig. 4 which is a combination of the two circuits. This gives an output which is constant within ±0·4mV for an input voltage change from 25 to 50V. There are small changes which occur due to changing temperature of the transistor when slow changes are made and these may be reduced by using a larger transistor. The variations were reduced to ±0·2mV for input voltage changes of 25 to 60V but the circuit has now reached the stability limit due to variations of ambient temperature.

Yours sincerely,


G. N. PATCHETT,
A. R. BAILEY,
University of Bradford.


**The authors reply:**

DEAR SIR,—I would like to reply to some of the points raised by Messrs. Patchett and Bailey concerning the article written by Dawson and myself.

I agree that a constant voltage device has zero slope resistance, while a constant current device has an infinite slope resistance. Practical devices must, however, have finite slope resistances (at least over any significant range) and in this case it is instructive to examine the intercepts of the incremental resistance lines with the voltage and current axes. Thus a "constant current type" device has an intercept on the positive current axis and the negative

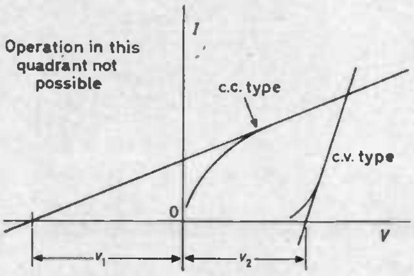


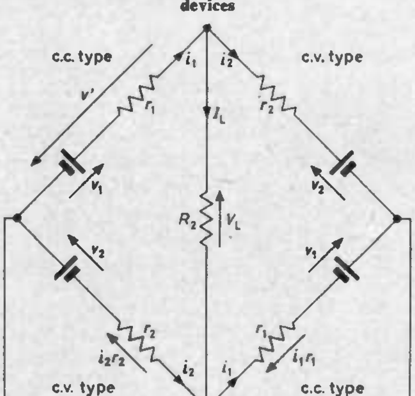

**Fig. 5. Characteristics for constant current and constant voltage devices**

voltage axis, while a "constant voltage type" device has an intercept on the positive voltage axis as shown in Fig. 5. Hence a "c.c. type" device is characterized in its working region by $V = Ir_1 - v_1$ while a "c.v. type" may be represented by $V = Ir_2 + v_2$, where $v_1$ is the magnitude of the negative voltage intercept and $v_2$ is the positive voltage intercept. These classifications correspond to those of "type I" and "type II" devices referred to by Morton[1] when considering non-linear elements for use in stabilizing circuits.

If a bridge network is formed with diagonally opposite pairs of "c.v. type" and "c.c. type" devices, then over the linear ranges of the device characteristics it can be represented by an equivalent circuit of the form shown in Fig. 6. At first sight Fig. 6 might suggest that output voltage $V_L$ could be greater than $V_{IN}$. However, in operation the voltage drop across the incremental resistance $r_1$ due to the current $i_1$ must cause the overall p.d. across the arm to oppose the current and $v'$ is of the polarity indicated. If this were not so the arm would be acting as a source of energy.

Because the incremental resistance of the four arms of the balanced bridge must be the same and because the "c.v. type" elements have a low incremental resistance, while the "c.c. type" elements have a high incremental resistance, to achieve the balanced condition it is necessary to add resistors in series with the former and/or in shunt with the latter. The *I-V* characteristics of the two arms forming one half of the bridge will then be as shown in Fig. 7, i.e. parallel in the operating region. By drawing a load line across these characteristics the output voltage can be determined.

**Fig. 6. Bridge formed from c.v. and c.c. type devices**




With a battery biased transistor as a "c.c. type" element (shunted to reduce the slope resistance) the incremental resistance line makes a negative intercept on the voltage axis and hence the arm operates as a "c.c. type" device in the sense used by the authors. When the transistor is biased with the aid of a Zener diode, this intercept is reduced and in practice may fall near to the origin, but the current passed by the transistor prevents the arm from becoming a constant voltage element as suggested by Professor Patchett.

Voltage stabilizing bridges can operate with one or more of the arms containing a non-linear device, and various combinations of "c.v. type", "c.c. type" and purely resistive arms can be used. Professor Patchett's first circuit (Fig. 2) is an example of a circuit using one

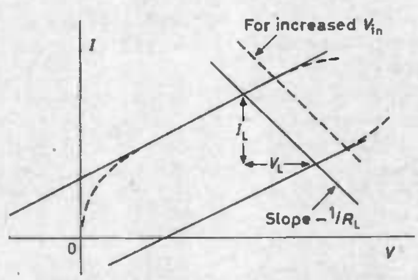



**Fig. 7. I.V. characteristics of the two arms of bridge**

non-linear arm, while in our article a circuit containing two Zener diode arms and two purely resistive arms was mentioned. In this sense, then, the transistor arms are unnecessary; but it was found that stabilization occurred over a much greater range when the transistors were introduced. It is interesting to note that the output voltage of both the battery and Zener diode biased versions of the bridge described by Dawson and myself rises at high input voltages, whereas the output from most bridges containing non-linear elements (including the Zener diode bridges) is of the form shown in Fig. 8. This occurs because the *I-V* characteristics shown in Fig. 3 usually converge and can even intersect at high voltages causing the output voltage to reverse polarity.

Changes in dissipation are bound to cause difficulty in all circuits of this type and it is advisable to under-run all components. In the circuit described the transistors were used in conjunction with heat sinks to maintain a good response to slow changes of input voltage.

As to the performance of the circuit described, ±2mV represented the uncertainty introduced by possible experi-

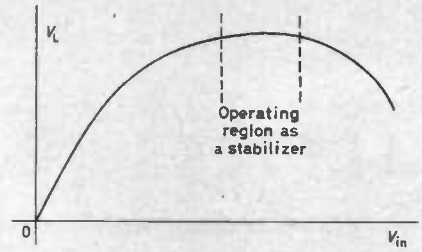



**Fig. 8. Output from bridge with non-linear elements**

mental errors with the measuring apparatus available, and I suspect that the output voltage remained appreciably more constant than this; but I cannot claim that it could rival the performance of Professor Patchett's sophisticated circuit of Fig. 4. However, my interest in this field was limited to a simple extension of the earlier work on stabilizing bridges using more modern devices.

Yours sincerely,

G. C. Taylor,
University of Surrey.


**REFERENCE**

1. Morton, C. *J. Sci. Instrum.* *21*, 15.


## Stabilizing Output of Wien Bridge

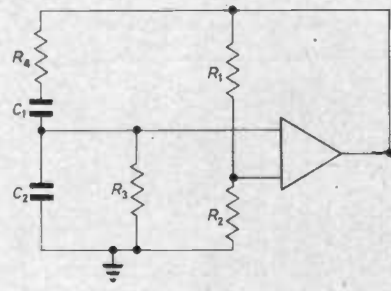
Dear Sir,—Following the letter in your December issue 'Stabilizing Output of Wien Bridge' by R. Krishna the use of a thermistor to stabilize the frequency of a Wien bridge oscillator may prove of interest. A recent application in this company was for an oscillator to be cheaply produced giving a stable output frequency, adjustable between 4 and 5kHz, when operating within a temperature of 10°C to 30°C.

A typical Wien bridge circuit using npn silicon planar transistors[1] was adapted for the purpose and this oscillator, together with its Zener diode regulated supply, exhibited a reduction in preset frequency of 32 parts per 1 000 for an increase in temperature from 15°C to 45°C. Referring to Fig. 1 it can be shown that the frequency of oscillation is given approximately by:

$$\omega^2 = \frac{1}{R_4R_3C_1C_2}$$

and a compensating increase in frequency can therefore be obtained by

**Fig. 1. Arrangement of bridge oscillator**



reducing the $R$ or the $C$ components in this expression.

Oven tests were carried out on the oscillator and when the value of $R_4$ was altered to compensate for drop in frequency due to increase in temperature the relationship shown in Fig. 2 was obtained. Fig. 2 also shows the variation in resistance of an n.t.c. thermistor

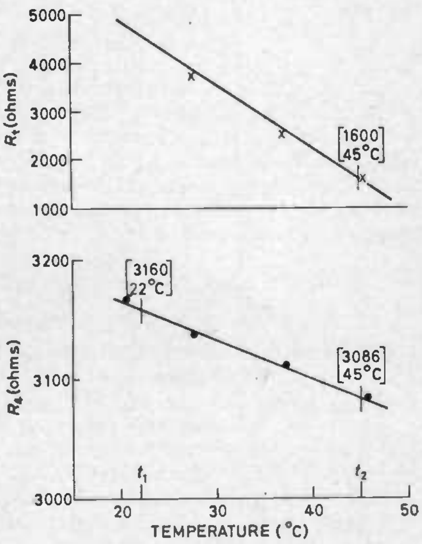



**Fig. 2. Effect of temperature**

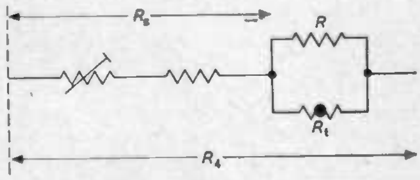



**Fig. 3. Thermistor-resistor network**

(STC Type M53) over the same temperature range, the logarithmic characteristic of the thermistor being represented by a straight line. If $R_4$ is replaced by an n.t.c. thermistor/resistor network then compensation will occur by reduction of the equivalent resistance of the network with increase in temperature thus maintaining the preset frequency.

In the thermistor/resistor network shown in Fig. 3 the value of $R$ to give the required compensation may be calculated from the solution of the simultaneous equations arising from:

$$R_4 - R_s = \frac{R_t R}{R_t + R}$$

for temperatures $t_1$ and $t_2$, where $R_t$ is the resistance of the thermistor and $R_4$ the corresponding resistance required to give a particular frequency at temperature $t_1$ and $t_2$.

This produced a calculated value for $R$ of 634Ω and the use of a 680Ω resistor produced a frequency stability of 1 part per 1 000 over the temperature range 15°C to 45°C, the frequency rising with increase in temperature. The thermistor $R_t$ was fixed to the outside of one of the larger components of the oscillator to make it insensitive to rapid variations in ambient temperature, this arrangement producing a thermal time constant similar to that of the complete oscillator of approximately 10min.

The effects of non-linearities in the oscillator phase shift network and the thermistor have little effect on the overall stability of oscillation within the temperature range given above.

Yours faithfully,

A. J. DAVIS,
Star Paper Mills Ltd.

**REFERENCE**


1. Wide Band Transistor Phase Shift Oscillator. Application Note No. 1, Ferranti Ltd.


**The author replies:**

DEAR SIR,—My communication published in your December issue was concerned with the output amplitude rather than the output frequency.

The idea of using a thermistor to improve frequency stability seems interesting. For a single or over a narrow range of frequencies, obtaining a stable frequency output presents relatively little difficulty. However, it will be very useful if Mr. Davis' idea could be extended to wide-band oscillators to achieve higher stabilities against ambient temperature variation.

Yours faithfully,

R. KRISHNA,
University of Saskatchewan.

## A Modified Cascode Circuit Using Complementary Transistors

DEAR SIR,—Your correspondent J. L. Linsley Hood in the April issue connects a pair of complementary transistors into an r.f. amplifier and calls it a 'cascode', (his Fig. 3). This does not follow the recognized usage of the name which is that of a grounded emitter (cathode) stage driving a grounded base (grid) stage. The circuit he has drawn is that of a long-tailed-pair, i.e. a grounded collector (anode) stage driving a grounded base (grid) stage. This complementary arrangement has an infinite 'tail' and this fact can give an advantage in certain differential amplifiers.

He claims the input transistor has improved h.f. performance. This is marginally so because the impedance in its emitter is not zero. However, this impedance is small, being the input impedance of the following grounded base stage. He could have obtained this advantage in the true cascode by leaving part of the input emitter resistor unbypassed.

It may be noted that the output transistor has a much higher output impedance (because of the high impedance emitter load) in the true cascode than it has in the long-tail-pair arrangement. Also the noise performance of the cascode is superior to that of the long-tail-pair for much the same reason.

L. H. DAWSON,
Chelmsford, Essex.

**The correspondent replies:**

DEAR SIR,—It had not occurred to me that the emitter-emitter connected cascode arrangement could also be regarded as a long-tailed pair which had experienced the same fate as the three blind mice; but I remain unrepentant in the use of the term.

The relative performance of the two arrangements shown in Fig. 2 and Fig. 3 is tabulated below.

| | EMITTER-EMITTER CASCODE | COLLECTOR-EMITTER CASCODE |
|---|---|---|
| *Input impedance* (parameters for $VT_1$) | $\simeq B \,.\, h_{ie} + C_{be}/B + C_{ob}$ | $\simeq h_{ie} + C_{be} + B \,.\, C_{ob}$ |
| *Output impedance* (parameters for $VT_2$) | $\simeq 1/h_{oe} + C_{ob}$ | $\simeq 1/h_{ob} + C_{ob}$ |
| *Current gain* | $\simeq h_{fe}\ (VT_1)$ | $\simeq h_{fe}\ (VT_1)$ |
| $B = 1 + A$ | | |

where $A$ is the effective open-loop gain of $VT_1$ in the circuit,

$$\simeq h_{fe}(VT_1) \times h_{ib}(VT_2)/h_{ie}(VT_1).$$

A typical value for this with silicon planar transistors having an $h_{fe} = 100$ would be of the order of 2 or 3.

The input impedance of the emitter-emitter connected arrangement will be higher than that of the more normal circuit connexion by this amount. The proportionation of the noise components between the two transistors is altered but the overall noise product appears to be similar.

Yours faithfully,

J. L. LINSLEY HOOD
Taunton, Somerset.

## 'Corotrons'

DEAR SIR,—We have read with interest an article in your February issue by Mr. Barlow and Dr. Watson entitled 'Some approaches to photomultiplier power supply design.'

The article states that the current range of Victoreen 'Corotrons' is limited to between 5μA and 100μA.

As Victoreen's U.K. representative, we would like to point out that Victoreen Corotrons, with maximum currents up to 3·0mA and voltages from 350V to 30kV, are now available.

Yours faithfully,

A. JETHA,
Walmore Electronics Ltd.,
London, W.C.2.

# NEW BOOKS

## Analysis of Discrete Physical Systems

By H. E. Koenig. 447 pp. Med. 8vo. McGraw-Hill Pub. Co. Ltd. 1967. Price 110s.

THE title of this book implies that it is concerned with the analysis of systems which are subject to quantization in either time or amplitude. However, the author's interpretation of 'discrete' is the description of systems made up of a finite number of interacting components. By means of the concepts of state space and linear graph theory, they aim to provide a technique for the logical development of mathematical models of complex physical systems.

The first chapter introduces some fundamental concepts of system theory and the mathematical representation of certain basic signals. The following chapters describe the modelling of simple two-terminal and *n*-terminal components. A large number of worked examples are included in the text, drawing primarily from three types of physical processes—electrical, mechanical and hydraulic. The next chapter introduces the concept of a linear graph and uses it to develop the overall description of a series of interconnected elements. Chapters 5 and 6 include many important results of matrix theory in its application to the analysis of linear systems. At this point, the authors appear to have drawn heavily on the work of Gantmacher. Stability, in the sense of Lyapunov, is defined and its application to linear systems is developed in terms of the roots of the system's minimal polynomial. The responses of systems to some simple deterministic inputs are derived in the following chapter and the final chapter appears under the heading 'Analysis of large-scale systems'. Appendices contain further matrix theory, the results of which are used in the text.

Modelling of physical systems is of considerable current interest because one important application is directed towards the overall computer control of complex systems. However, the mathematical description of many systems incorporating thermal or chemical processes results in partial differential equations as well as differential equations because of the inherent distributed parameter nature of parts of the system. This book gives no indication how its modelling techniques could be applied to such a system although infinite dimensional space could be avoided by use of, say, finite difference techniques. Further, the linear graph method appears to become cumbersome for high order multi-variable systems. This is evidenced by the final chapter. An example quoted here to indicate the application of the methods to a 'large-scale system' is the remote position control servomechanism with velocity damping! Clearly, this is included only to illustrate the principles involved, but application of a modelling technique to say, a power station boiler, must be able to cope with up to a hundred algebraic, differential and partial differential equations. Moreover, it is by no means a straightforward task to represent the system in terms of interconnected basic components. Any modelling process must be accompanied by a high degree of physical insight. After a matrix model of a large scale system has been derived, further progress is usually only made possible after a reduction in system order. The currently active field of system model reduction techniques receives no reference in this text.

A number of points arise from a detailed reading of the book. Firstly, the introductory chapter deals with the Fourier series representation of signals yet fails to mention the important minimization of mean square error and non-interaction of coefficients properties of the orthogonal functions. The Gibb's phenomenon is also omitted.

It is pleasing to see that the authors acknowledge the frequently ignored possibility of input derivative terms arising in the state space representation of systems. Such a system is reduced to the simpler form by a change of state variables. However, it would have been wise to point out at this stage that the new state variables have generally lost any physical significance. This is an important factor for design by Lyapunov and in the formulation of a cost function for, say, dynamic programming.

In the eigenvector and eigenvalue analysis no reference is made to multiple zero eigenvalues which would evidently be encountered in the representation of more than a single pure integration. Further, the stability analysis given provides only a yes/no decision on the stability or instability of a system. Techniques for the stabilization of an unstable system or for increasing the damping in an oscillatory system are not given. Analysis is the more powerful if it can be used to indicate the road to design.

Finally, the book analyses the response of systems to simple inputs only. Because the input to a practical system can generally be specified only in statistical terms, some attention to system response to stochastic or spectral inputs would have been expected.

R. EDWARDS.

## Differential and Difference Equations

By L. Brand. 698 pp. Med. 8vo. John Wiley & Sons Ltd. 1966. Price 90s.

THE title of this book is a good one. Both differential equations and difference equations are adequately dealt with and there is a good supply of exercises and applications.

Chapters 1 to 6 give a full, though not original, treatment of differential equations. Incidentally, there is a very useful section on eigenvalues. Full coverage of the *D* operator theory is given so that a comparison can be drawn between *D* and the difference operators used later. Chapter 7 deals with the Laplace transform and applications of its use.

Difference equations are treated in Chapters 8, 9. Chapter 10 is a long chapter on solution in series and one wonders why it did not precede Chapter 8. Chapter 11 is an interesting exposition of an operational calculus with a 'modern' approach developed by Mikusinski, a Polish mathematician. Chapter 12 deals with Existence and Uniqueness of solutions and Chapters 13 and 14 revert to numerical analysis again.

It is not apparent to the reviewer why the subjects are so mixed and perhaps this is the place to comment on the preface. Great emphasis is placed on the analogies between some ordinary differential equations and some difference equations, but although these analogies are indisputable, the text does not possess the originality suggested in the preface. This is, however, not a criticism of what the book does, but only what it apparently intended to do.

This is undoubtedly a very useful book, with clear exposition and comprehensive notations. It has been written to be understood, which cannot be said of all books on advanced mathematics. The author has not been frightened of putting in an apparently trivial step where it was thought necessary. The excellent supply of exercises and applications makes it a book to recommend to undergraduates with examinations in mind, but this is by no means the limit of its market.

D. W. PORTER.

### Signale und Systeme (Signals and Systems)

**By F. H. Lange. 428 pp. Med. 8vo. Friedr. Vieweg & Sohn, Germany. 1966. Price DM.36**

THIS very thorough treatment of the mathematical methods used in modern engineering is designed to supplement lectures given to advanced students (about third year) in communications and control. While the emphasis is on electrical systems the author intends to supply the basic tools for handling all manner of signals and of signal handling systems. This first volume deals with the frequency analyses of the subject. The great merit of the treatment lies in its organization, that is, in the fact that the whole of the subject is presented as a logically developed unit. A number of traditionally known methods of electrical engineering are shown to fit into a larger system thereby attaining fuller meaning and at the same time helping towards the understanding of the less familiar aspects. The field covered can be judged by the six main chapter headings:

- Unmodulated periodic signals
- Modulated periodic systems
- Non-periodic double side band signals
- Non-periodic single side band signals
- Pole/zero representation of linear systems (in the p-plane)
- Statistical and spectral characteristics of random processes.

It must be stressed however that this is a text book for the advanced student who will have to work through it systematically, making full use of the summaries, extensive notes and test questions at the end of chapters. Moreover as lengthy calculations and derivations are consciously avoided according to the introduction the student may have to refer to his mathematics text to understand some of the steps, or to check some of the expressions as there are some printing errors, a very disconcerting feature for those, like the reviewer, who are not too sure of their mathematics. In short, this is a difficult book to read even for those familiar with German, it is not recommended as a reference book but would probably be an admirable guide for the conscientious lecturer preparing a thorough and fascinating course for his more advanced students.

A small point worth mentioning perhaps is the extensive bibliography quoting both Anglo-American and Russian sources.

K. L. SELIG

### Basic Mathematics for Electronics

**By F. L. Juszili, N. Mahler and J. M. Reid. 450 pp. Med. 8vo. Prentice-Hall International. 1966. Price 72s.**

An elementary but systematic approach is presented to the application of algebraic, logarithmic exponential and trigonometric functions in the solution of electric current problems.

No background knowledge of electrical fundamentals is required but sufficient electrical theory is included to allow the reader to gain some understanding of the subject.

### Integrated Electronics

**By K. J. Dean. 132 pp. Crown 8vo. Chapman & Hall Ltd. 1967. Price 28s.**

The principles of operation of integrated circuits together with their limitations and advantages are discussed in this small book.

The applications of these circuits are dealt with and the design methods both for basic circuits and for systems using integrated circuits are considered.

The book contains a glossary of terms, bibliography and two appendices dealing with the types of bistable elements and with the use of Karnaugh maps in logical design.

### Automation and Instrumentation

**Edited by L. D. U. Pellegrini. 722 pp. Crown 4to. Pergamon Press Ltd. 1967. Price £7**

This book contains the proceedings of the 8th International Convention on Automation and Instrumentation held at Milan in November 1964 and s ponsored by F.A.S.T. (Federazione delle Associazioni Scientifiche e Techniche) at which 56 papers were presented.

### Magnetism and Magnetic Materials 1966 Digest

**Edited by C. Warren-Haas and H. S. Jarrett. 273 pp. Med. 8vo. Academic Press Inc. (London) Ltd. 1966. Price 88s.**

The 1966 Digest presents a survey of papers published in the preceding year dealing with this subject, the main source of references being the Index to the Literature of Magnetism.

### Basic Electricity for Electronic Engineers

**By A. W. N. Kerkhofs. 212 pp. Demy 8vo. Philips' Technical Library. 1967. Price 38s.**

The principles of electrical theory as taught in the electrical engineering courses of the Philips Company are provided in this book. Short sections dealing with fuller mathematical treatment are included for the benefit of radio technicians.

### Handbook of Basic Transistor Circuits and Measurements

**By R. D. Thornton. 156 pp. Demy 8vo. John Wiley & Sons Ltd. 1967. Price 34s.**

This is the seventh volume in the series dealing with transistors and other semiconductor devices published for the Semiconductor Electronics Education Committee.

### Electronic Communications

**By R. L. Shrader. 682 pp. Crown 4to. McGraw-Hill Publishing Co. 1967. Price 76s.**

Written for junior college and technical school students, this book contains an electronics and radio communication course including practical information and questions required to pass examinations for Federal Communications Commission radio licences, both commercial and amateur levels.



### Particle Accelerators

**Edited by R. Kollath. 337 pp. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1967. Price 75s.**

This book is a translation of the second edition of the book of the same title published in Germany, and consists of a collection of papers written by specialists in their own field.

It deals with the theoretical principles, design and operation of the various types of particle accelerators.

Linear and cyclic accelerators for electrons as well as heavy particles are described.

A list of some 300 references is included.

### Electronic Computers Made Simple

**By H. Jacobowitz. 330 pp. Demy 8vo. W. H. Allen. Publishers & Co. 1967. Price 10s.**

Intended primarily for self-tuition, this is the first book in the 'Made Simple' series, dealing with analogue and digital computers. Other subjects in this series include mathematics, intermediate algebra and analytic geometry, physics and chemistry.

### Electronic Laboratory Instrument Practice

**By T. D. Towers. 164 pp. Demy 8vo. Iliffe Books Ltd. 1967. Price 35s.**

A practical introduction is given in this book to the application of meters, signal generators, oscilloscopes and bridges used in electronic laboratories.

### Programmierung von Datenverarbeitungsanlagen (Programming of Data Information)

**By H. J. Schneider and D. Jurksch. 144 pp. Post 8vo. Walter de Gruyter & Co. 1967. DM.5.80**

# NEW EQUIPMENT

**A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.**

### DIGITAL ENCODER

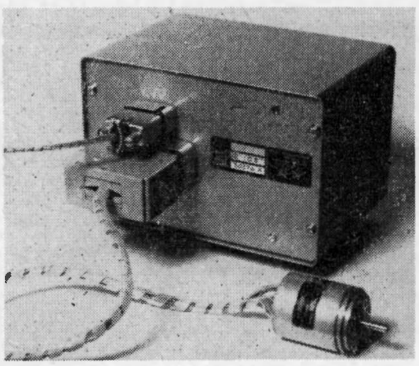
**Moore Reed & Co. Ltd, Walworth, Andover, Hampshire**

*(Illustrated below)*

A 13 digit binary encoder is now available from Moore Reed and Co. Ltd, having the advantages of very low cost, coupled with a 'noise free' output. The electromechanical part of the encoder is housed in a size 18 frame 2½in long. The associated electronics is based on English Electric Norlog elements and uses plug in cards for ready accessibility and ease of installation.

This new product is intended for applications in which space is not at a premium, but in which long life and zero contact noise are of paramount importance. The established 'MR Scan' principle of encoding is used, which was developed for very high reliability.

An added feature of this digitizer is the ability to 'freeze' all outputs at any desired instant, even though the encoder shaft continues to rotate. The same encoder is also available with all electronics 'built in' to the size 18 frame.



*For more information circle* **No. 291**

### ELECTROLUMINESCENT DIODES

**Ferranti Ltd, Gem Mill, Oldham, Lancashire**

A comprehensive range of gallium phosphide (GaP) and gallium arsenide (GaAS) light emitting semiconductor devices is now being manufactured by the Electronic Display Department of Ferranti Ltd.

The basic GaP lamp is a plastic encapsulated device only 0·05in long by 0·03in in diameter, with two 0·005in diameter lead wires extending from the body.

The most important application of these lamps is in the recording of high density information on film with particular reference to aerial reconnaissance photography. The fast rise and fall time of the light output, typically 25 and 250nsec respectively, also enables them to be used for film marking in high speed cameras. The arrays can be manufactured with flexible film-wire connexions terminating in standard multipin plugs. A minimum density of 0·8 above fog level on Ilford HP3 film is achieved by a red lamp (XP10 series) film marking array using a 1msec pulse at 50mA drive current. Due to the low sensitivity of most films in the red region of the spectrum, red lamps are generally only used for contact film marking, where they give a digital record the same size as the block format of the array. Most films however are very much more sensitive in the green region of the spectrum, and the green lamp (XP50 series) can be used for film marking through a lens system. Green lamps can therefore be used to record digital information of a very high density, so making available a greater picture area on each frame of the film.

The red and green lamps can also be used as visual indicators to show the 'state of circuit' in logic circuits, where the voltage and current requirements for driving the lamps are compatible with the power supplies of transistorized circuits.

The intensity of red or green light produced by the lamps when they are forward biased is proportional to the drive current. The maximum dissipation is 50mW under drive conditions of typically 1·9V at 20mA. The red lamps have a peak emission at 7 000Å and the green lamps at 5 650Å. The energy output of the XP10 series extends from 1·7 to 22·5$\mu$W and the luminous intensity from 40 to 500 microlumens. The corresponding figures for the XP50 series are from 0·05 to 1·5$\mu$W and 27 to 900 microlumens. When forward biased the lamps have an indefinite life if pulsed at less than full dissipation, e.g. no degradation of light output occurs after $12 \times 10^6$ 50mA pulses of 1msec duration. If operated at full d.c. dissipation the light output degrades by up to 30 per cent in the first 48 hours and thereafter at typically 15 per cent every 1 000 hours.

The GaAS diodes which are being manufactured emit in the infra-red region at approximately 9 000Å when biased in the forward mode, with the intensity of the output being proportional to the drive current. The XS30L and XS30P series are mounted on TO-46 headers and have lens-can and resin droplet encapsulations respectively. They have a maximum continuous output of 500$\mu$W at typically 1·3V and 100mA. The XS40L and XS40P series are mounted on TO-5 headers and have optical window and resin droplet encapsulations respectively. The maximum continuous output of this series is 5mW at typically 2V and 1A.

The emission of the diodes fitted with lens-cans has an inclusive angle of 100°, and the diodes encapsulated in the resin droplet emit as a point source. GaAS diodes may be used in conjunction with silicon diode detectors, having a peak sensitivity at 9 000Å.

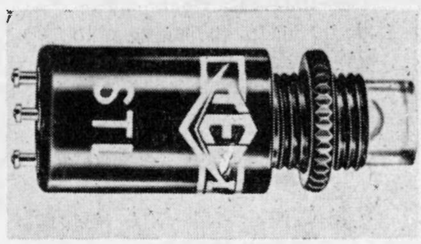
*For more information circle* **No. 292**

### TRANSISTORIZED INDICATOR

**Distributed by: Litton Precision Products, 503 Uxbridge Road, Hayes, Middlesex**

*(Illustrated below)*

A new subminiature 'Mini-Lite', the STL series, has been introduced by Transistor Electronics Corporation. In this rugged, extremely small device, a neon or incandescent lamp is transistor controlled from low-level signals

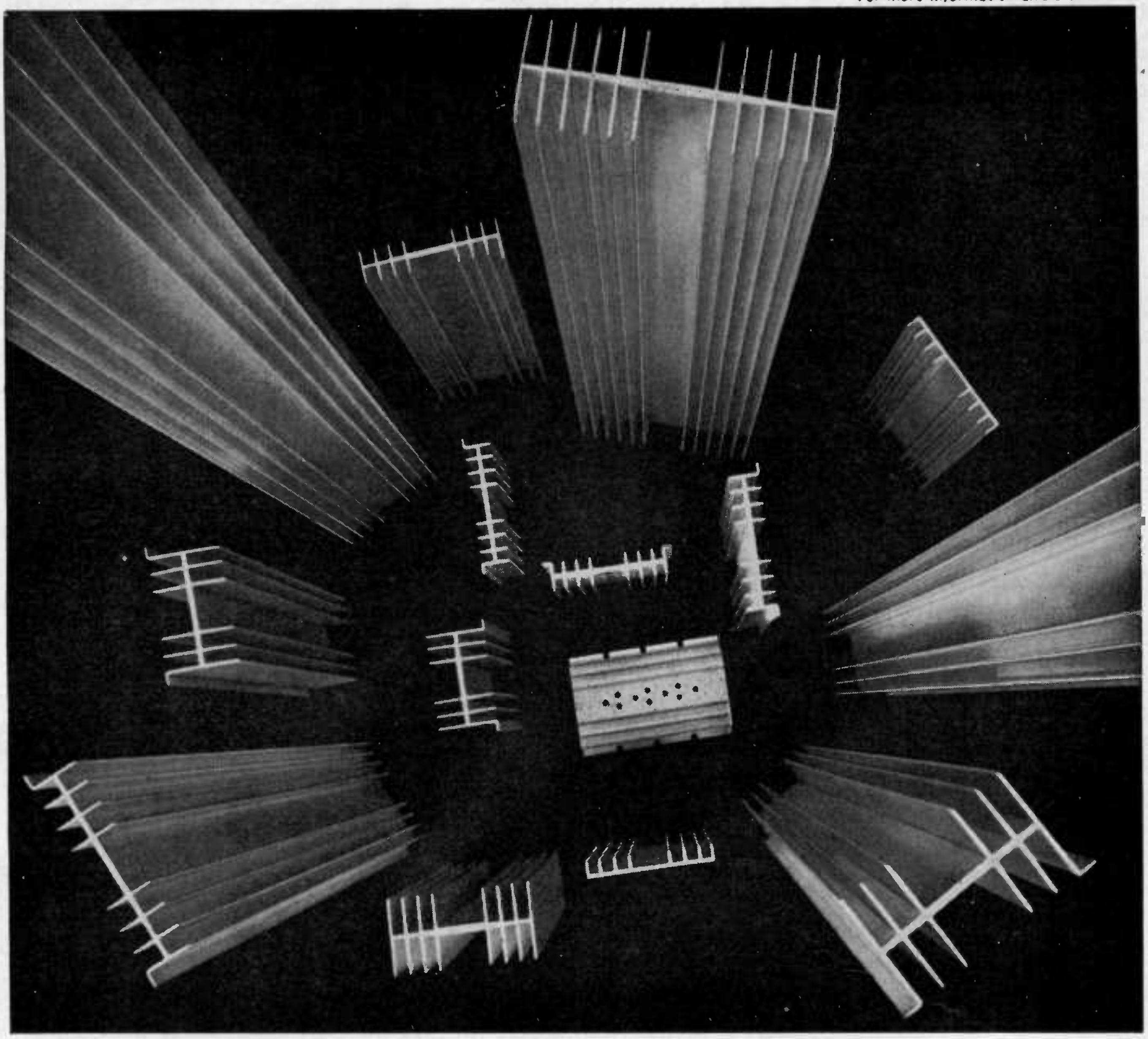



typically present in discrete component or integrated circuit systems.

This long-life, solid-state indicator is designed for use in applications where many indicators are required in a small area. Lamp, transistor, and related circuit are packaged in a 0·36in diameter by 0·6in long body. Overall length is only 1¼in. It mounts from the rear with a knurled nut in a ¼in hole on centres as close as ⅜in. It will fit panel thickness from 1/16in to 3/16in.

The TEC-LITE STL series indicator is available with a permanently wired neon or incandescent lamp that operates on signals as small as 0·5V with supply current of only 0·7mA. Lenses are offered in a choice of 13 colours. One letter, numeral or symbol up to 5/32in or two characters 3/32in high can be hot stamped on the lens face in various colours. Terminals are turret lug, brass with gold finish, isolated from the indicator body which is black or clear anodized aluminium. Operating and storage temperature range is −40°C to +65°C at 95 per cent humidity, maximum.

*For more information circle* **No. 293**



## YARN CLEARER

**Crabtree Instruments Ltd, Green Works, Colne, Lancashire**

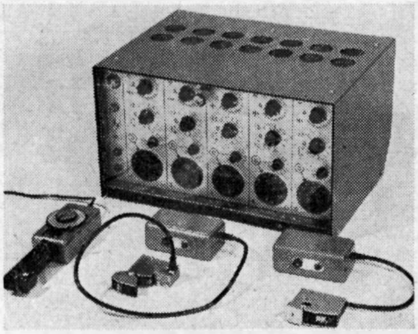
*(Illustrated below)*

An advanced type of yarn clearer available from Crabtree Instruments Ltd is the recently introduced Peyerfil D-Type. This is a yarn clearer capable of controlling the length of the short faults (including torpedo slubs) taken out of the yarn and is also capable of eliminating spinners doubles and similar long thick occurrences in the yarn.

Operating on an optical system, light is diffused by means of a patented optical system over the total surface of the yarn passing through the detector head. This ensures that no slub in any plane can escape detection.

The new D-Type control unit is designed to supply up to 50 clearers in five banks of ten. Each bank may have independent settings. The D-Type control unit has two sets of controls, one being used for setting for diameter and length of short term faults, and the other being used for setting for increase in yarn count, i.e. spinners doubles.

The reference datum for the yarn is obtained by means of a simple but accurate instrument known as the D-Tester which measures the yarn in a similar manner to the clearers.



*For more information circle* **No. 294**

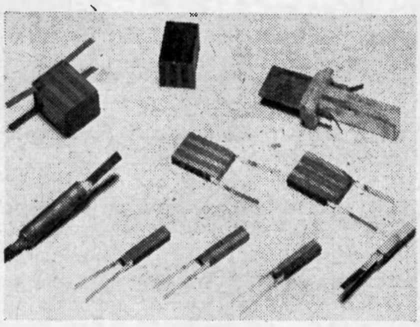
## THERMOELECTRIC MODULES

**The Plessey Co. Ltd, Chemical and Metallurgical Division, Towcester, Northamptonshire**

*(Illustrated above right)*

High output and the ability to operate at high temperature are the major features of thermoelectric devices developed by The Plessey Company's Allen Clark Research Centre, Caswell, and now in production by the company's Chemical and Metallurgical Division at Towcester. All the devices are similar in being made from a semiconductor material—iron disilicide—by a specially devised process that produces a monolithic ceramic mass with no external connexions at the hot junctions and only two terminals on the cold side. The units differ solely in number of junctions and size, according to the output voltage and power required. Output voltage is proportional to the number of junctions and power to the cross-sectional area.

A typical unit is type TEG 3, which has three junctions formed in a monolithic slab of iron disilicide 15 by 19 by 3mm, with flying leads. This gives an open circuit voltage of 600mV with a maximum output of 100mW for a 500°C difference between the hot and the cold junctions. The mean resistivity is $7{\cdot}5 \times 10^{-3}\Omega$cm, giving an internal resistance of $1{\cdot}1\Omega$. A maximum operating temperature of 800°C is recommended.



The method of manufacture, briefly, is to melt iron and silicon in an r.f. furnace under an inert atmosphere, adding to the melt the doping element that produces the semiconducting properties. The interesting point here is that unlike most semiconductors, which need closely controlled impurity and doping levels in the order of a few parts per million, the purity of the iron disilicide is not critical and the doping levels are around 2 to 3 per cent. The dopants are unusual—cobalt for n type, aluminium for p.

The 'n' and 'p' materials are then rendered into a form suitable for making cold compacts. A stack of alternate 'n' and 'p' compacts is then built up with thin sheets of a ceramic insulator between them, allowing contact only where the junctions are to be formed.

A r.f. heating and pressing technique converts the stack into a dense monolithic mass from which individual devices of the required dimensions are cut.

During the process a special heat treatment ensures that the correct phase of the material is formed to give the thermoelectric properties.

One device, the TEG 1, has a single junction giving 25mW at 200mV for 500°C between hot and cold sides. This is particularly intended for flame failure detection on gas burners, and offers fast response and oxidation resistance at high temperatures: 800°C is the recommended maximum, but operation is possible at up to 900°C. Unlike telluride semiconductor sensors, which must be encapsulated, the Plessey devices can work directly in the flame, since the first few seconds' exposure creates a silica glaze that protects the underlying material: life is therefore indefinitely long.

Using the TEG 3, sufficient power is available for completely electrical operation of gas valves, instead of the usual manual set, electrical hold method used with metal couples. The high power and relatively high internal resistance allow the use of fine connecting leads and ordinary solenoids, instead of the heavy braids and special low resistance circuits required by metal thermocouples.

A larger module is available that gives 0·6V and 400mW for 500°C across the faces. Applications envisaged include control of and power supply to larger gas systems, and running small electrical apparatus up to about 5W, e.g. communication repeaters and trickle chargers for batteries.

*For more information circle* **No. 295**

## DECADE CAPACITANCE BOXES

**J. J. Lloyd Instruments Ltd, Brook Avenue, Warsash, Southampton**

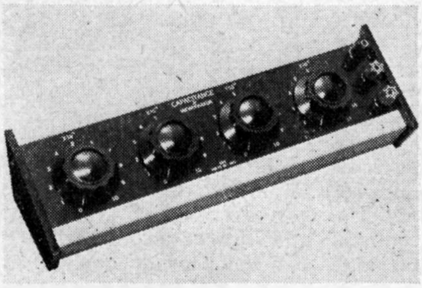
*(Illustrated below)*

Two new precision decade capacitance boxes, suitable for use over a very wide frequency range, are now manufactured by J. J. Lloyd Instruments Ltd. The four decade model has a range from 0 to $1{\cdot}111\mu$F and the three decade from 0 to $0{\cdot}111\mu$F.

Both models are a mixture of carefully aged silvered mica and polystyrene capacitors having a guaranteed minimum accuracy of $\pm\frac{1}{2}$ per cent. Losses at 1MHz are less than 0·0005 per cent and the maximum working voltage is 500V.

The units are housed in screened cases, the dimensions of which are only 13½in by 4in by 3½in for the four and three decade models respectively.

All instruments are individually trimmed so that for any given setting of the decades the capacitance obtained is absolute and there is no need to make an allowance for stray capacitance, in addition to which each box is supplied



with an individual test certificate, giving the actual values of each setting of the bottom decade to an accuracy of 0·1 per cent.

*For more information circle* **No. 296**

## POWER SUPPLY

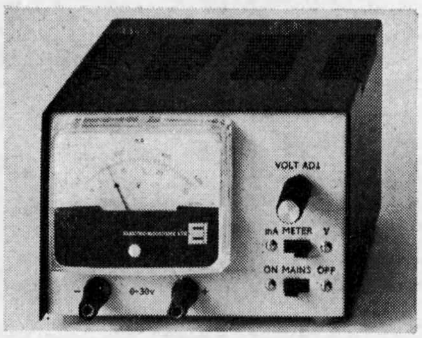
**Electro Inductors Ltd, Grafton Road, Croydon, Surrey**

*(Illustrated on page 398)*

This new, low priced power unit is completely transistorized, and permits the close regulation of voltage necessary for both classroom work and research.

The input voltage may be either 230V or 115V a.c., in either case $\pm10$ per cent, 50 or 60Hz. The output voltage varies by only $\pm0{\cdot}02$ per cent or 10mV, whichever is greater, for $\pm10$ per cent change in line voltage.

The output voltage is continuously adjustable from 0 to 30V d.c. by means of a helical type control; fine adjustment is possible since ten turns cover the full voltage range.




The maximum output current is 600mA, and the regulation is such that over the full current range, and at ambient temperatures up to 45°C, the output voltage varies by only ±0·1 per cent or 10mV, whichever is the greater. The ripple content is less than 1mV. The unit is protected by an automatic current limiting device, giving full short-circuit protection. The output voltage and output current is indicated on a meter with a scale length of 2¾in, having an accuracy of 2 per cent.

*For more information circle* **No. 297**

## HIGH POWER V.H.F. LIMITERS

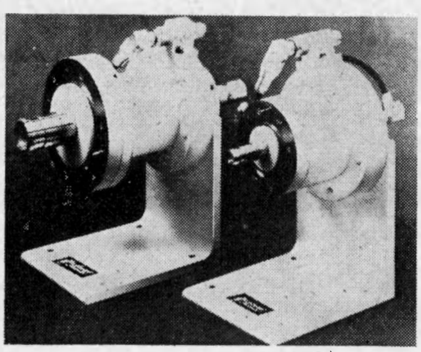
**Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire**

*(Illustrated below)*

A completely new series of passive semiconductor limiters, designed for operation in the v.h.f. range 20 to 200MHz at power levels of 1MW peak, 5kW average, are available from Microwave Associates Ltd. They are intended for use in high-power surface-to-base tracking radars.

Over any 20 per cent bandwidth within this frequency range, these limiters provide complete receiver protection by allowing the passage of microwave energy only below a specified power threshold and limiting the energy above this threshold to a safe level. Their design consists of two stages: the first comprises a number of pin diodes which radially shunt the coaxial transmission line; the second stage which employs varactors to reduce the leakage level from the first. All these diodes are individually replaceable.

The devices have a recovery time of less than 1μsec and an insertion loss of 1dB nominal. Spike leakage is non-existent. They are extremely compact and a pressure seal is provided at the input flange which is specified as either 3⅛in or 1⅝in EIA. Type N connectors are standard but other types may be specified.



*For more information circle* **No. 298**

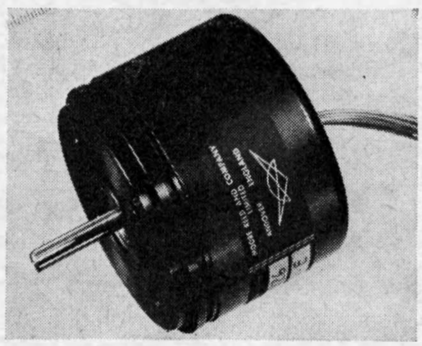
## BINARY CODED DECIMAL ENCODERS

**Moore Reed & Co. Ltd, Walworth, Andover, Hampshire**

*(Illustrated below)*

Moore Reed & Co. Ltd has added a new series of b.c.d. encoders to its range. These encoders in size 23 (2¼in diameter), have an output of two or three decades, each of which is in natural binary consisting of four digits weighted 1, 2, 4 and 8. The total count is obtained in one shaft revolution and the ranges available are 0 to 99, 0 to 199, 0 to 359 and 0 to 399. The brush choosing and gating circuits are built into the encoder, which ensures that the output is unambiguous for any shaft position without the necessity of any external circuits.

The 100 and 200 count versions also have an 'odds parity' output per decade. Thus the sum of binary 'ones' at any instant in any decade is always odd.



The output takes the form of zero volts for a binary nought and 9V for a binary one when the unit is energized with 12V d.c. An 'inhibit' line, controlling the least significant digit, enables the user to 'hold' this in a steady state during readout.

*For more information circle* **No. 299**

## NULL DETECTORS

**K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4**

*(Illustrated above right)*

K.S.M. Electronics Ltd has produced two new instruments to add to its range of products, these are two models of a combined null detector and tuned amplifier. These models are battery driven with exceptionally long battery life.

Both models can be used as sensitive detectors for a.c. bridge work or as low noise amplifiers. The input sensitivity being as low as 1μV for f.s.d. Another application is as a wave analyser at audio frequencies. In the case of the NK3 the tuned frequency range is 15Hz to 20kHz. Another application for both models is their use as a transducer amplifier.

Both models contain frequency selective amplifiers which can also be used as a flat response amplifier. These amplifiers can be used as linear or logarithmic to avoid overload.


*For more information circle* **No. 300**

## GRADIENT PUMP

**Cenco Instrumenten Mij. N.V., Konijnenberg 40, Breda, Netherlands**

*(Illustrated below)*

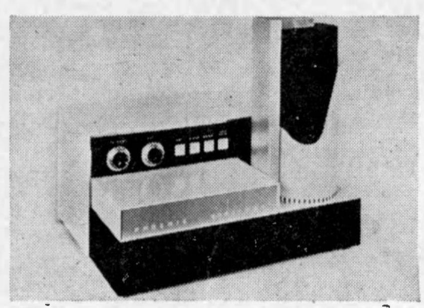
The availability of a new programmed ratio metering pump suitable for chromatographic applications as well as for many other gradient elution systems has been announced by a Cenco subsidiary, Phoenix Precision Instruments Company.

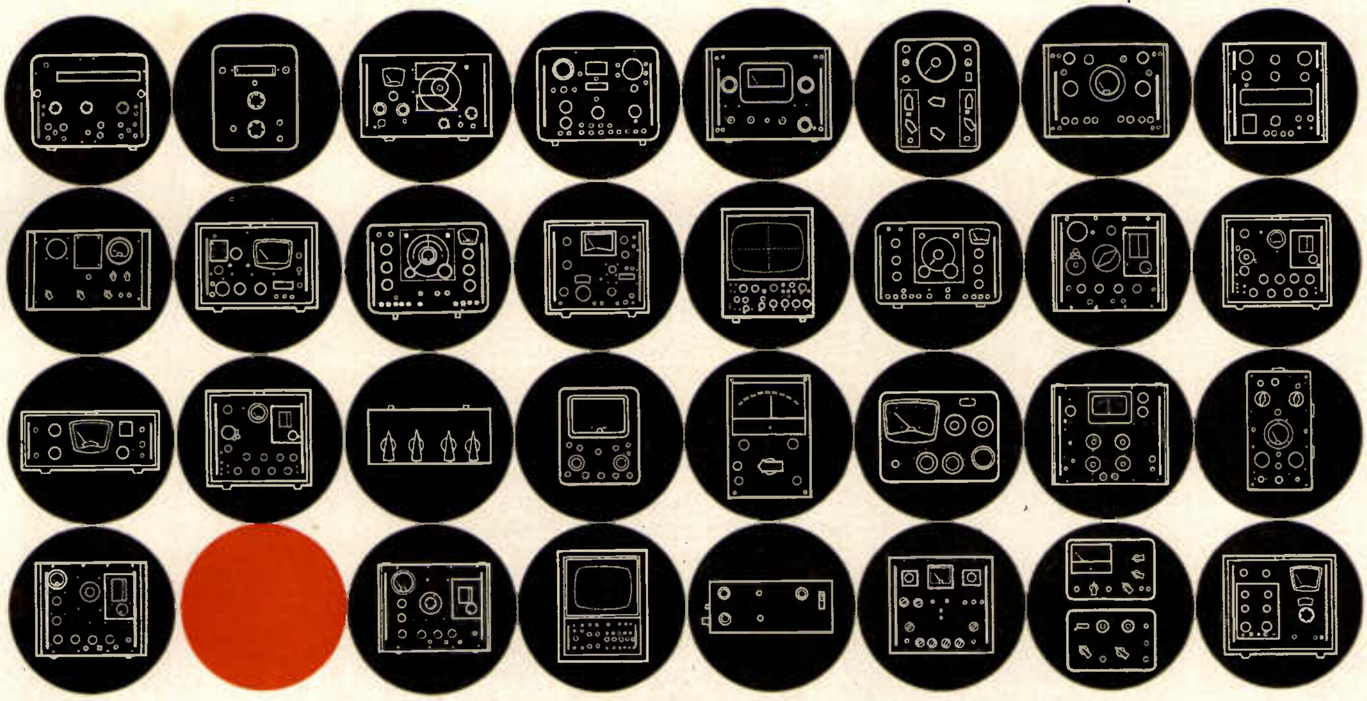
Designated the 'Varipump', this new unit utilizes a photoelectric curve follower which allows the reproduction of any desired gradients. The gradient to be produced is simply drawn on a piece of chart paper, or any other sheet of paper, with a pen or pencil, cut with a pair of scissors, and fastened to the drum of the photoelectric curve follower. The curve follower then follows the pre-cut curve, transmits the signal to the pump unit which reproduces the exact same curve in form of a pH or ionic strength gradient. After use, the curve can easily be removed from the drum and stored for future use. The complexity of gradients is only limited by the art of drawing and cutting of such curves from paper.

The drum, upon which the gradient curve is fastened, can be varied in speed to make a complete revolution in 24, 12 or 2 hours. Due to the continuous loop of the gradient curve, repetitious runs can be made without resetting or refilling of reservoirs.

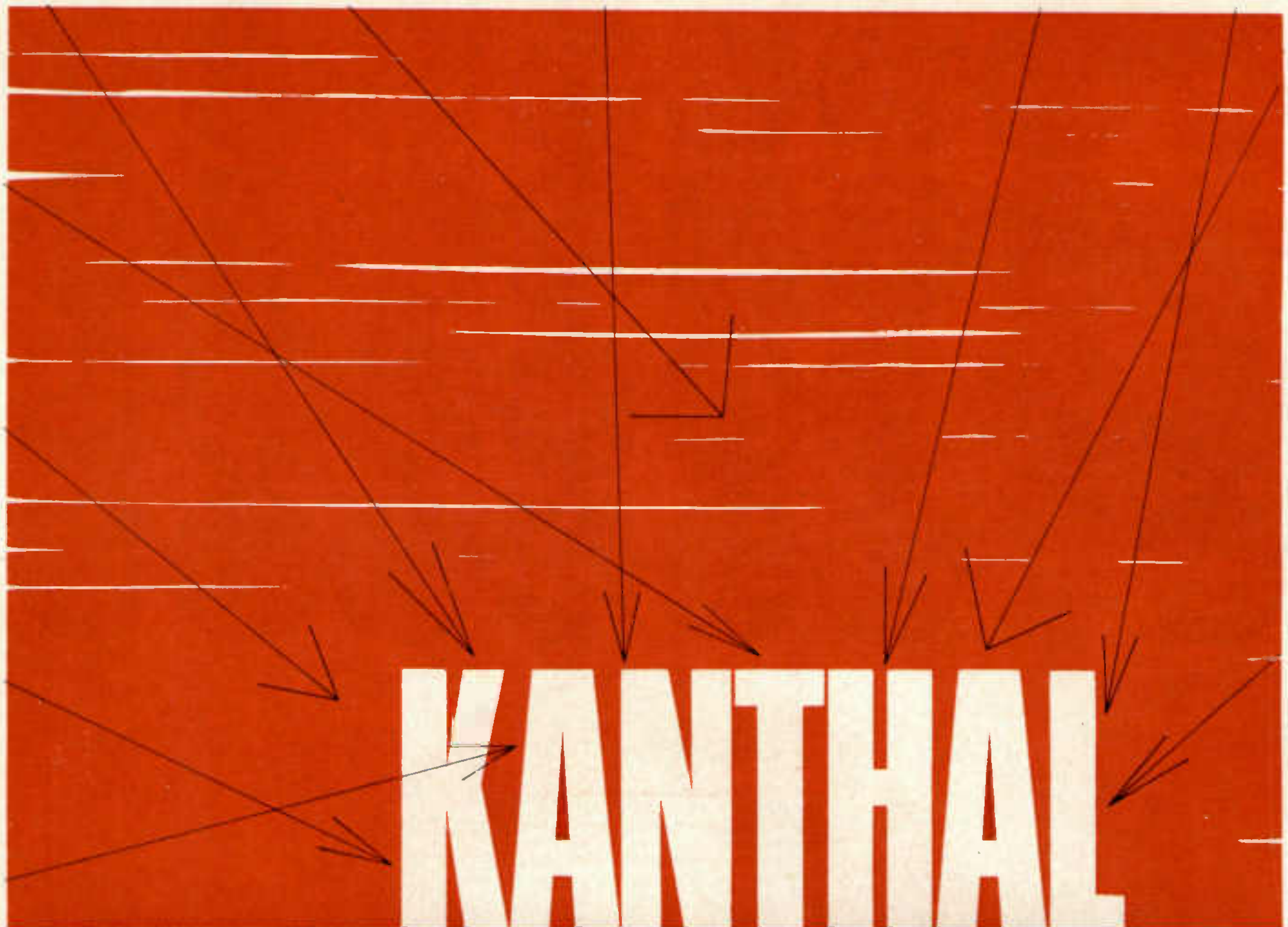
The highly sensitive photoelectric curve follower allows for accurate reproduction of gradients with maximum slope steepness of 89°. The maximum contrast principle of black to white, as is the case in this system, assures positive tracking and prevents the photocell from losing the trace as often occurs on so-called line followers.

Volume delivery of the Varipump is adjustable from 0 to 130ml/h with an accuracy of 0·5 per cent of pre-set volume. The pump unit and all fittings






NEW!
Airmec
MILLIVOLTMETER TYPE 301A




## The world's largest manufacturer of electrical heating & resistance material

Kanthal has a wide temperature range divided into separate alloy groups with individual characteristics. It is used as wire strip and tape for most applications in the field of electrical heating.

KANTHAL A-1, the premier grade, is manufactured as large cold-rolled strip and as wire above 1 mm in diameter (SWG 19 or B&S 18). It is recommended for element temperatures up to 1375°C (2507°F).

KANTHAL A will stand element temperatures up to 1330°C (2426°F) and is manufactured in wire sizes heavier than (SWG 47 or B&S 44).

KANTHAL DSD is manufactured as wire sizes heavier than 0.05 mm (SWG 47 or B&S 44) and as cold-rolled strip and tape. The maximum element temperature recommended for KANTHAL DSD is 1280°C (2336°F).

KANTHAL DSI is manufactured as wire sizes heavier than 1 mm (.04") diameter and as large strip. Maximum element temperature recommended is 1150°C (2102°F).

KANTHAL today is a by-word in the field of electrical heating.

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and lines are fabricated from stainless steel allowing operation with corrosive fluids with discharge pressures up to 1 000 lb/in$^2$.

*For more information circle* **No. 301**

### TRANSVERTORS

**Valradio Ltd, Browells Lane, Feltham, Middlesex**

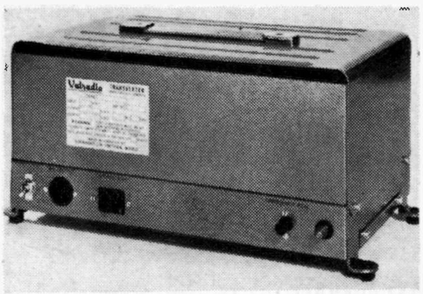
*(Illustrated below)*

The circuit of these units consists of a stable oscillator unit driving heavy duty transistors, the resultant output being fed to a special transformer of the ferro-resonant type, which besides producing a sinewave output also achieves a high degree of voltage regulation against changes of input voltage and load.

Two voltage outputs are provided, i.e., 115V and 230V selected by a simple switch. The units are designed to operate over the voltage range of 11V to 15V for 12V types, 22V to 30V for 24V types.

Frequency output is 50Hz ± ½Hz, or better than 0·005 per cent with a 'Resonator' frequency synchronizer. The frequency stability is such that the frequency will be within ±1Hz when the input of a 12V unit is down to 6V.

The driving oscillator which is of the RC type, incorporates temperature stable components, maintaining a steady frequency over the range of −40°C +45°C.



This range of sinewave transvertors has been specifically designed for waveform and frequency sensitive equipment such as video tape recorders, sound recorders and scientific instruments, etc.

Both 12V and 24V types are available with either 120W or 200W output.

*For more information circle* **No. 302**

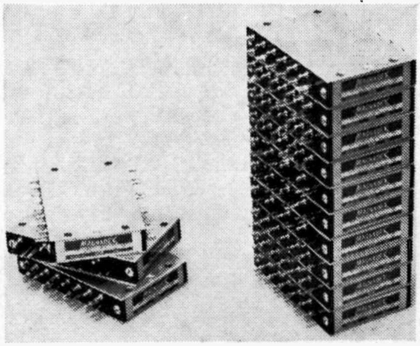
### REED RELAYS

**Distributed by: Automatic Controls & Components Ltd, Bridge Chambers, Bridge Street, Leatherhead, Surrey**

*(Illustrated above right)*

The 'Magnadek' reed relay deck, manufactured by Bon Automation Ltd, is designed to accommodate a permutation of coils and reed switches within a standard body, thus meeting virtually every switching requirement.

The Magnadek is supplied in six standard arrangements giving a maximum of six single or double wound 24V d.c. coils, suitable for switching 100At reed switches. Up to ten reed switches can be accommodated in each deck. Each coil has separate solder connexions and the reed switches can be



changed at any time without removing the deck from the complete assembly.

*For more information circle* **No. 303**

### INTEGRATED CIRCUIT ADDERS

**Motorola Semiconductor Products Inc., York House, Empire Way, Wembley, Middlesex**

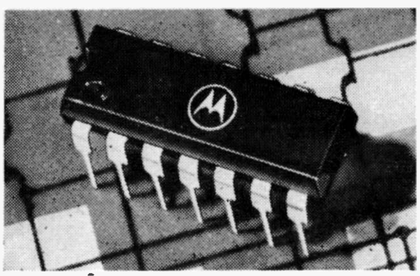
*(Illustrated below)*

Motorola Semiconductor Products Inc has introduced a new concept in logic circuits which more than doubles the operating speed of integrated circuit full adders and subtractors.

Simultaneous gating, called series current steering, is used, which eliminates the successive delays experienced with cascaded gating. Instead, the complete gating function—whether sum, carry, difference or borrow—is performed simultaneously. This represents a saving in gate delays of at least one and frequently three or more. Coupled with the inherently high speed of current mode ECL circuits, the series gating technique provides 4 or 5 to 1 reduction in propagation delay compared to other monolithic full adders.

The new full adder is the Motorola MC1019 and the subtractor is the MC1021. Both units are part of the Motorola MECL II family of high speed emitter coupled integrated circuits. This family is fully compatible in logic level with the original MECL series, but offers high speed and even lower can count.

While the slowest propagation delay in the adder is very short—a maximum of 15nsec for the addend to sum or carry—the design is such that the highest speed response is provided where it will most improve system performance. For example, in the MC1019P full adder, the carry input to carry output propagation delay is the shortest—a maximum of only 8nsec. This means that in a parallel or ripple adder, the propagation of the carry across the adder will be as short as possible. This same system orientated design is shown in the MC1021 full sub-



tractor which shows an 8nsec maximum propagation time for the borrow in to borrow out.

The inputs to the adder are single rail (function only) for the addend and augend, and double rail (function and its complement) for the carry input. Both the sum and carry outputs are double rail (SUM and $\overline{\text{SUM}}$ and $\overline{\text{CARRY}}$ and CARRY). For the full subtractor the corresponding relations exist. (The minuend and subtrahend inputs are single while the borrow in, difference, and borrow out are double rail.)

Like all MECL II circuits, the MC1019 and MC1021 are designed for operation from a −5·2V d.c. power supply. The a.c. fan out is 15. Both the MC1019P and MC1021P are housed in the 14-pin Unibloc plastic package. This is a dual in-line configuration with 100 mil spacing between adjacent leads. The devices are rated for operation over the commercial/industrial temperature range of 0 to +75°C.

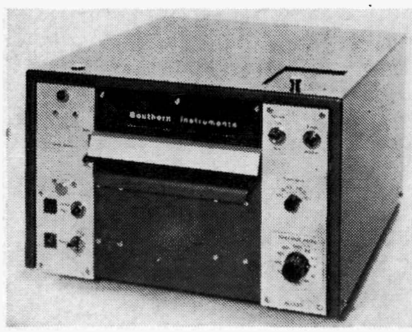
*For more information circle* **No. 304**

### ULTRA-VIOLET RECORDER

**Southern Instruments Ltd, Frimley Road, Camberley, Surrey**

*(Illustrated below)*

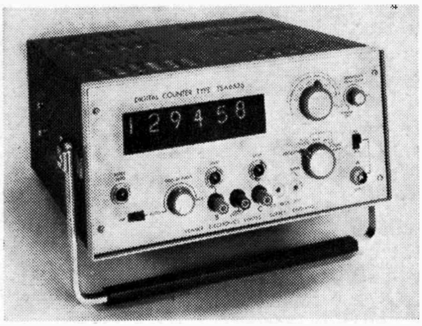
A new addition to the range of ultra-violet recorders is being introduced by




Southern Instruments Ltd. It offers a number of improvements, the main being the use of optical grid lines, full width timing lines, and the facility for adding a further six speeds.

The instrument has ten data recording channels, two datum traces and an event marker. The optical grid lines occur every 2mm and full width timing lines are at intervals of 0·01, 0·1, 1 and 10sec. Paper capacity is 106·7m (350ft) using u.v. recording paper of up to 152·4mm wide (6in). Paper speed range is from 2 500mm/sec to 4mm/sec in 12 steps. The addition of a further six speeds will extend the lower end of the range down to 0·54mm/min. By use of a high quality light and optical system writing speeds in excess of 1 016m/sec are achieved. The twin electrode mercury super pressure lamp used has an independent power supply, and a guaranteed life of 200 hours. A comprehensive range of S.M.I. galvanometers is available and other facilities offered are remote control, trace identification and paper take-up.

*For more information circle* **No. 305**



### 12·5MHz COUNTER

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

*(Illustrated above)*

A 12·5MHz all silicon frequency meter/counter/timer, type TSA 6636, has been introduced by Venner Electronics Ltd. Designed for general test and research use, the instrument gives a six-digit readout, including decimal point display, on neon number tubes.

For frequency measurement (10Hz to 12·5MHz), gating times are 1$\mu$sec to 10sec in decade steps. Input sensitivity is 75mV into 250k$\Omega$. Single and multi-period measurement covers a 10Hz to 1MHz range with periods from 1$\mu$ to $10^7$. As a timer, the TSA 6636 covers 1$\mu$sec to 10sec in decade steps, single and two-line start/stop facilities being provided. Provision is also made for gated counting.

Display time is variable from 0·5 to 5sec or infinite, and accuracy is $\pm 1$ count $\pm$ crystal stability ($\pm 1 \times 10^{-6}$).

An optional extra is a five-line code printer output.

*For more information circle* **No. 306**

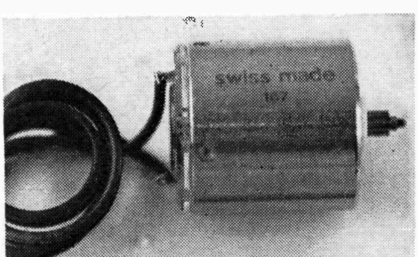
### MICROMINIATURE MOTOR

**Reno S.A., 165 rue Numa-Droy, 2300 La Chaux-de-Fonds, Switzerland**

*(Illustrated below)*

The Escap 15 motor, which is of extremely small dimensions, is a new product of the Swiss watch industry.

Its high starting-torque and the short time it takes to reach its full speed are due to its original design: the rotor, in the form of a cylinder open at one end, is formed only by the winding. The turns of wire are oblique and crossed over one another, so that they require no support. Thus the entire available air-gap space is filled with copper, which makes the motor highly efficient. Moreover, as the rotor turns round a stator consisting of a two-pole permanent magnet, it has been possible to give it an exceptionally large diameter




in relation to the overall dimensions of the motor.

The bearing is of self-lubricating powder-pressed bronze. The brushes are of gold and the collector is of silver. Lastly, a reduction gear covering the housing can be coupled to the motor.

These motors measure 15mm $\times$ 16mm and weigh 12g. The maximum voltage is 2V and the minimum starting voltage without load is 60mV. The normal operating speed is about 10 000rev/min, but reduction gearing can be fitted. The starting torque is 5·5g.cm.

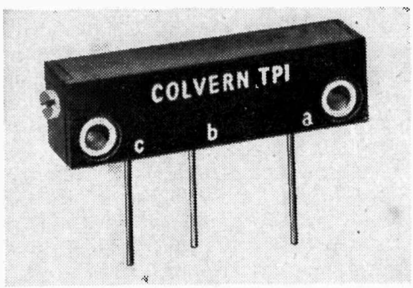
*For more information circle* **No. 307**

### TRIMMER POTENTIOMETER

**Colvern Ltd, Spring Gardens, Romford, Essex**

*(Illustrated below)*

The trimmer potentiometer type T.P.1 introduced by Colvern is intended as a low cost, high grade item for commercial applications. With the case moulded in glass loaded nylon, a stainless steel lead screw positions a beryllium copper contact incorporating a slipping clutch to avoid overturning. The pick off is lead through a separate bus bar to the terminal pin and the end connexions to the resistance element are welded to ensure good stability. Dissipation is 1W at 70°C derated to zero at 125°C. Resistance range is 10$\Omega$ to 20K$\Omega$ $\pm$ 10 per cent. Dimensions are 1·250in $\times$ 0·3in $\times$ 0·360in. Terminal pins on 0·1in module.




*For more information circle* **No. 308**

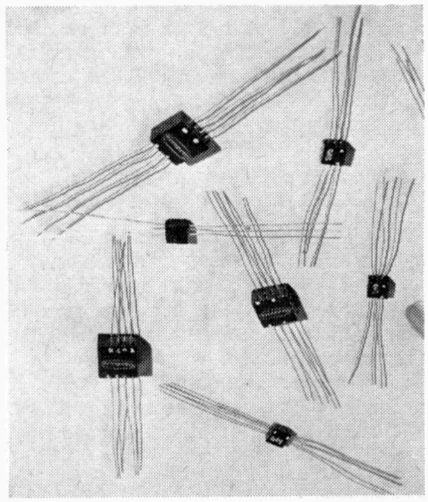
### MICRO-MINIATURE TRANSFORMERS

**Gardners Transformers Ltd, Somerford, Christchurch, Hampshire**

*(Illustrated above right)*

To be known as the Lilliput series these transformers will be available for wide band carrier frequency line coupling, carrier frequency inter-stage coupling, invertor driver transformers, pulse transformers and s.c.r. trigger transformers. They are designed to be compatible with many other miniature components used on printed circuit boards and are provided with dual-in-line output leads in a flat pack system.

Very high performance has been achieved by the use of new forms of construction offering more flexibility than true toroidal fabrication and permitting the use of ultra thin highly permeable nickel iron cores down to ·000125in thickness. By these means switching times of less than 10nsec have been achieved. Also available in this series are ferrite cored screen trans-



*For more information circle* **No. 309**

formers suitable for direct insertion into printed boards.

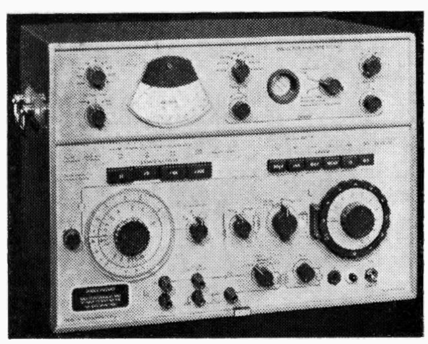
### INDUCTOR ANALYSER

**Marconi Instruments Ltd, St. Albans, Hertfordshire**

*(Illustrated below)*

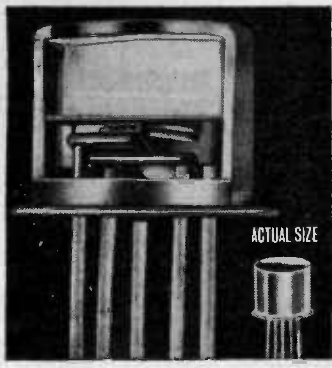
Marconi Instruments' new inductor analyser, type TF 2702 covers virtually every aspect of inductor analysis, and is very easy to balance. It is both a self-contained low- and medium-current inductance bridge (up to 0·5A) and the nucleus of a complete high-power inductor test assembly. The robust circuit elements and sensitive detector permit measurements at currents ranging from 10A to 10$\mu$A and voltages from 500V to 10mV. The inductance range is 0·3$\mu$H to 21 000H and it can be used at any frequency from 10Hz to 20kHz. Up to 3kHz the accuracy is $\pm 1$ per cent and at higher frequencies it is within 1·25 per cent.

The instrument has facilities not previously combined in one instrument. The current rating of 10A extends up to 21H and at 3A it is possible to measure 210H. Switched alternative arrangements permit inductors on test to be measured in either the series or parallel mode, depending upon the application, so that the required value is measured direct; this eliminates the need for calculations based on frequency and $Q$. Balance is achieved by a variable capacitor, and this ensures that even low Q inductors can be measured without the interaction of a sliding balance. The resistance balance has at least three scales on each

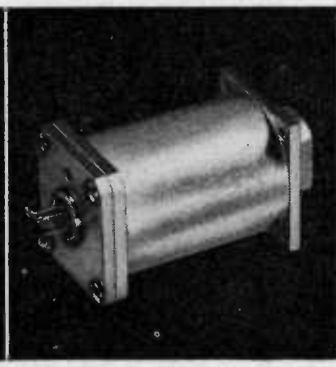


For more information circle No. 105

# ROTAX COMPONENTS & TRANSDUCERS




Teledyne Relay



401 Transducer


200 Series Encoder


HT-250 Trimmer Potentiometer

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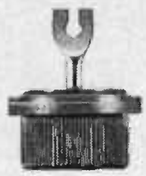
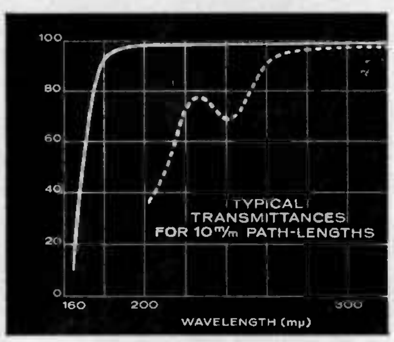
## Its price is not your main consideration



LUCAS
SEMICONDUCTORS

For more information circle No. 106



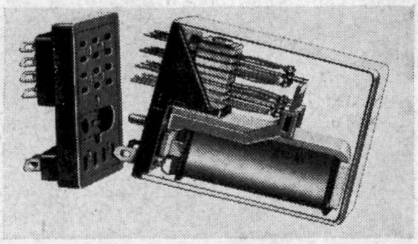

100
80
60
40
20
0
160
200
300
WAVELENGTH (mμ)
TYPICAL TRANSMITTANCES FOR 10 m/m PATH-LENGTHS

inductance range, so that the balance setting is not cramped.

A small cathode-ray tube is provided for easy searching and preliminary balance for unknown values, and for the display of overload characteristics of the iron core; this is a particularly useful feature when cores are overloaded at unexpectedly low current levels. The final balance is indicated on a meter, which also acts as a monitor for the a.c. inductor voltage.

Internal oscillators supply low power at the supply frequency and at 1 and 10kHz. These supplies can be monitored and d.c. bias applied directly into the bridge up to 0·5A or up to 10A using an optional accessory, a.c./d.c. mixer unit TM 8339.

*For more information circle* **No. 310**



## MINIATURE A.C. RELAY

**Oliver Pell Control Ltd, Cambridge Row, Woolwich, London, S.E.18**

*(Illustrated above)*

What is claimed to be the first British miniature plug-in a.c. relay to fit the internationally recognized socket has just been announced by Oliver Pell Control Ltd. This is the Varley VPAC relay, which is available in three standard contact arrangements: with two changeover contacts, four changeovers, or six makes or breaks. All of these normally handle 1A, but the two-changeover model can also be ordered with 5A or 5A (special duty) contacts.

Standard coils range from 6V to 250V operating voltage, and standard contacts are of gold-flashed fine silver, or hard gold alloy. Operate time from random switch-on is 5 to 15msec.

Dimensions are length 0·960in, width 0·730in, height 1·580in, and with the maximum number of contacts the length becomes 1·180in.

As with all Varley miniature relays the dust cover is made of transparent plastic, and all are delivered in throw-away sterile plastic packs.

*For more information circle* **No. 311**

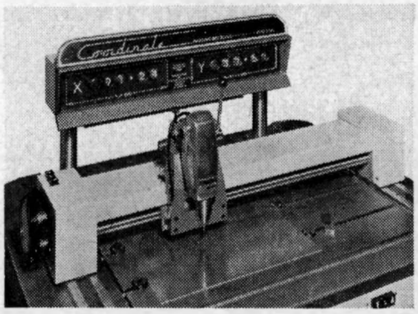
## CO-ORDINATE MARKING-OUT TABLES

**C. & N. (Electrical) Ltd, The Green, Gosport, Hampshire**

*(Illustrated above right)*

Just over a year ago C. & N. (Electrical) Ltd introduced the 7800 co-ordinate marking out table that enabled the previously laborious and expensive marking out of sheet material to be carried out by semi-skilled labour, with greater accuracy, in a much shorter time.

A new and improved version is now available, among the improvements is




the incorporation of a floating zero that enables both X and Y co-ordinates to be zeroed at any position in their travel. This dispenses with the need to work from a fixed datum. The capacity of the machine is 19in in the X plane and 12·5in in the Y plane, maximum clearance under the workhead is 1in. The digital read-out gives increments of 0·001in or to an accuracy of ±0·002in. The display also indicates 'plus or minus' dimensions when working either side of zero datum. The co-ordinate lead screws are electrically driven, final adjustment being made by handwheels.

Inch and metric versions are available together with a range of accessories including punching and scribing tools, a drill head and an optical inspection attachment.

*For more information circle* **No. 312**

## DATA ACQUISITION SYSTEM

**Smiths Industries Ltd, Kelvin House, Wembley Park Drive, Wembley, Middlesex**

*(Illustrated below)*

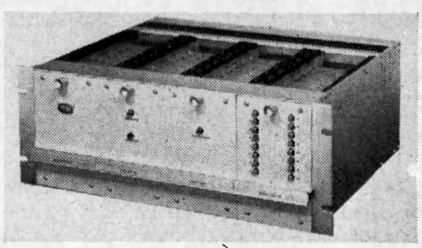
The Industrial Instrument Division of Smiths Industries Ltd has introduced a new range of data acquisition equipment known as 'Scan-Tel'.

The range is an extension to the existing Smiths telemetry systems. Designed to operate over the temperature range −5°C to +55°C to fully meet industrial conditions of operation, the system uses solid state silicon devices throughout and the printed circuit boards are interconnected using either wrapped joints or 'hypertac' plug-in connectors.

The equipment is designed to fit into standard 19in P.O. racks and is made of a number of styled plug-in sub-units. By using various combinations of these sub-units, systems may readily be built up to cover applications of data acquisition, including telemetry and telecontrol, at highly competitive prices.

In addition to telemetry and telecontrol the range embraces the transducer, data-logging and print-out aspects of data acquisition to give fully comprehensive systems.

*For more information circle* **No. 313**



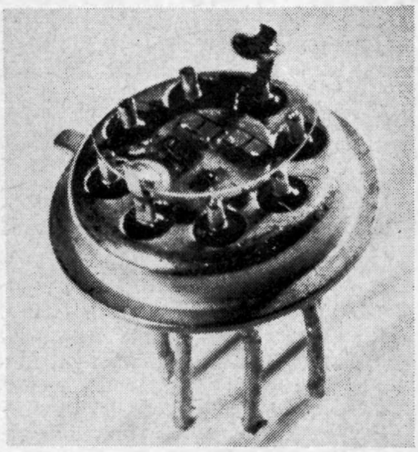
## MICRO-OSCILLATOR

**The Marconi Co. Ltd, Chelmsford, Essex**

*(Illustrated below)*

A completely new micro-oscillator, probably the smallest in the world, has been developed by The Marconi Co. Ltd. Completely sealed inside a transistor can, this unit is by far the smallest in the Marconi range of solid state, packaged oscillators, which provide design engineers with a comprehensive selection of low cost, 'ready-to-use' oscillators for a wide variety of tasks, both for production equipment and in experimental use.

This tiny oscillator, type F3185, consists simply of an integrated circuit feed-back amplifier with a crystal mounted



above it. The micro-circuits are attached to the TO-5 header in the normal position, while the crystal is supported on its two connexions to this circuit. In the prototype, a 5th overtone, AT-cut crystal is used, to give an oscillator frequency of 100MHz. In this case, an external inductor and trimming capacitor are necessary, but future models will employ crystals operating at their fundamental frequencies, which will not require any additional circuit elements. Where this is not possible, these external components will be designed into a slightly larger container taking full advantage of the small size of the oscillator itself.

The complete oscillator is sealed inside the TO-5 header can, together with its crystal, and is therefore completely isolated from the atmospheric environment. Temperature stabilization is not provided in the same unit, but the small size of the oscillator makes this a simple matter in cases where high stability is necessary. The complete unit is sufficiently small in size to make it possible to mount a number of oscillators in a single crystal oven.

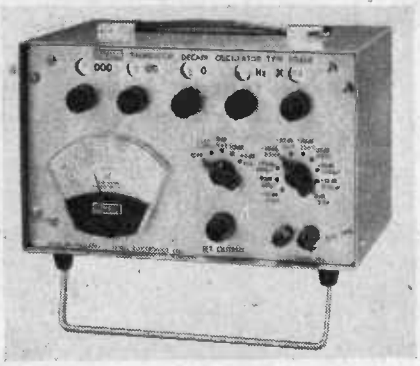
*For more information circle* **No. 314**

## TRANSISTOR DECADE OSCILLATOR

**Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire**

*(Illustrated on page 402)*

Very pure and stable sinusoidal oscillations at frequencies from 0·2Hz to



1·22MHz are produced by this lightweight portable signal source. Frequency is selected by four additive decade controls with in-line digital read-out and a five position multiplier switch. The first three controls are twelve position switches marked 0, 1, 2 . . . 11 and the fourth decade control is continuously variable with a linear calibration 0 to 10 divided into 40 divisions. Any frequency may be selected with a discrimination better than $\pm 0{\cdot}03$ per cent or 1/10th of the specified frequency accuracy.

The output is monitored by a meter with an expanded scale covering a range of 6dB. A continuously variable control and a step attenuator permits the output to be varied from $-94$dBm to $+10$dBm into 600Ω. The output impedance is 600Ω at all settings.

The power supply is four internal batteries with a life of 400 hours but a.c. mains operation is possible by replacement of the batteries with a Levell power supply unit.

*For more information circle* **No. 315**

### RATIOMETER

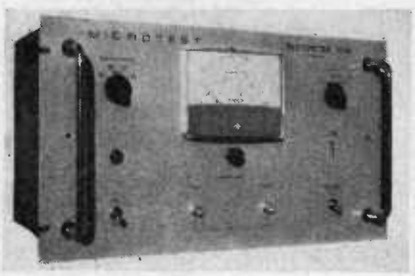
**Microtest Ltd, 9 Old Bridge Street, Kingston-upon-Thames, Surrey**

*(Illustrated below)*

The type 701A ratiometer was designed for reflectometer applications in conjunction with unlevelled signal sources or sweep generators.

The ratiometer can provide accurate results independent of incident power variations as high as 20 to 1. The front panel meter displays v.s.w.r. and/or reflection coefficient (per cent). An output socket is provided which allows oscilloscope presentation to be made in conjunction with swept frequency measurements.

Three types of r.f. detectors may be used in conjunction with the type 701A, viz. crystals or bolometers at either 4·3mA or 8·7mA bias current. As the bias current is derived from a constant current source bolometer burn-out due to the instrument is prevented.



*For more information circle* **No. 316**

### D.C. BRIDGE SUPPLY AND INDICATOR

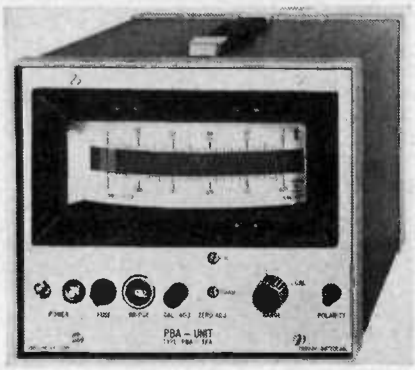
**Vibro-Meter S.A., Haletop Cevio Centre, Wythenshawe, Manchester, 22**

*(Illustrated below)*

Vibro-Meter SA, the Swiss instrument company, has introduced a new d.c. measuring system which provides an accurate indication of load, pressure or stress when used with resistive transducers or strain gauges.

The indicating instrument is particularly noteworthy; a 'Gossen Pantam Meter' is fitted which uses a light beam technique both to indicate the measured value and to operate up to four photoelectric switches for warning and control purposes.

The bridge supply and indicating unit, Vibro-Meter type PBA-1/A incorporates a chopper amplifier; the sensitivity of which can be adjusted to accept signals as low as 1mV for f.s.d. An output signal of 1V is available to drive recording



instruments. The amplifier also includes coarse and fine balance controls for the transducers and a pre-set calibration facility to reset or check the amplifier gain. The special feature of the instrument is the long term stability and accuracy.

The accuracy is within 0·25 per cent f.s.d. variations in temperature affect the accuracy by less than 0·2 per cent and the stability against mains variation is less than 0·5 per cent for 240V$\pm$10 per cent deviation.

Vibro-Meter also offer resistive pressure transducers, load cells and strain gauges compatible with this equipment.

*For more information circle* **No. 317**

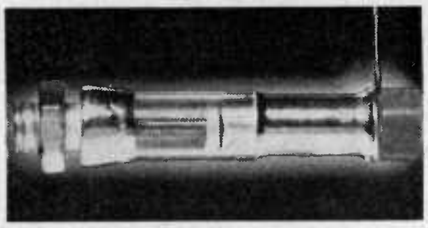
### QUARTZ TRIMMER CAPACITORS

**Oxley Developments Co. Ltd, Priory Park, Ulverston, Lancashire**

*(Illustrated above right)*

Oxley Developments Co. Ltd has recently introduced a trimmer capacitor of unusual design, utilizing quartz as a dielectric, and has been specifically developed to meet the needs of modern industrial and professional electronic equipment.

The patented design incorporates a non-rotating, self-aligning piston, anti-backlash drive mechanism and non-extending adjustment screw.



The materials and construction combine to create a capacitor which displays excellent low loss, high frequency characteristics.

These ultra high stability capacitors are available for printed circuit board and chassis mounting and are available in capacitance values from 2 to 10pF; nominal temperature coefficient of +20in $10^6$/°C; power factor less than ·0005 at 10kHz; insulation resistance greater than $2 \times 16^6$MΩ; working voltage 1kV d.c.

*For more information circle* **No. 318**

### MODULAR PLUG-IN CONTROLS

**Distributed by: Britec Ltd, 17 Charing Cross Road, London, W.C.2**

*(Illustrated below)*

'Elesta' series 14S modular plug-in controls combine easily with each other to form extremely compact systems to control machines and automatic processes.

These modular plug-in controls measure only 2 7/8in × 1 3/32in wide × 4 3/16in high, and consist of electronic timers type ZS.14S for interval, delay and sequence timing, of universal switching amplifiers type KS.14S for photoelectric control, contact thermometers or sensitive probes, and of type MR.14S relays having up to four changeover, or six normally-opened, or six normally-closed contacts.

*Electronic Timers:*

Type ZS.14S10 = unstabilized timers.
Type ZS.14S20 = stabilized timers.

These plug-in *RC* timers incorporate 'Elesta' precision cold-cathode tubes with a virtually unlimited operating life, they are particularly recommended where reliability and long service life under arduous conditions are of prime importance.

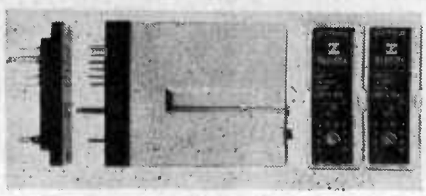
Time ranges: 0·1-1sec, 0·5-5sec, 2-20sec. These time ranges can be extended by connecting an external capacitor to terminals 3 and 6 of the socket.

Supply voltage: 220-250V a.c. 40/60Hz.

Repetition accuracy: $\pm$2 per cent with constant supply voltage.

Influence of supply: ZS.14S10 = $\pm$20 per cent for $\pm$10 per cent.
variations ZS.14S20 = $\pm$5 per cent for $\pm$10 per cent.

Contact rating: 250V a.c. 6A, non-inductive.



Remote control: Type ZB.41 potentiometer with scaled dial allows for remote control of the time scale, if desired.

Sockets: All timers are supplied with socket ZB.35 14-pin.

*Universal Switching Amplifiers: Type KS.14S*

These are versatile plug-in controls, suitable for operation with sensitive high-impedance contacts as well as with photo-resistors. Used in conjunction with the 'Elesta' range of photo-electric light sources and photo-receivers, and with miniature lamp transformer type ZS.40/20, they form compact photo-electric systems to monitor meters, to control level of liquids or powders, to guard automatic presses, open doors, control traffic, etc.

Supply voltage: 220-250V a.c. 40/60Hz.

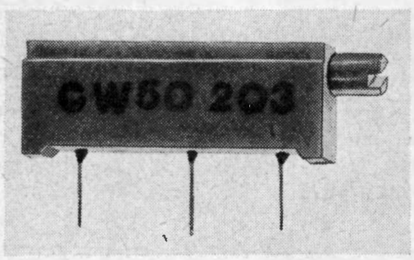
Contact rating: 250V a.c. 6A, non-inductive.

Max. counting rate: 4 to 5 per second (min. input pulse duration 100msec).

*Plug-in Relays: Type MR.14S*

The MR.14S is a plug-in relay, supplied in a transparent case having the same dimensions as the 14S modular controls. These relays are available with coil voltages from 3V to 240V a.c., or 3V to 300V d.c., and with slugged coils for use with cold-cathode tubes. The maximum number of contacts are either four change-over, or six normally-opened, or six normally-closed—rated at 250V a.c. 6A non-inductive. These relays combine with ZS timers and KS controls to provide auxiliary switching facilities.

*For more information circle* **No. 319**




## TRIMMER POTENTIOMETERS

**Reliance Controls Ltd, Drakes Way, Swindon, Wiltshire**

*(Illustrated above)*

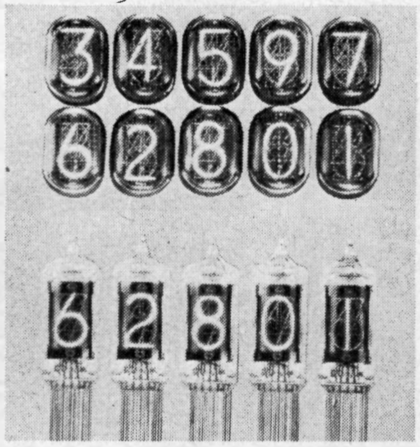
The CW.50 rectangular trimmer potentiometer is 0·74in long and is fitted with printed circuit base pins. It is available in resistances up to 20kΩ and is rated at 1W at 40°C with a temperature range of −65°C to +125°C. It has an insulation resistance of 100MΩ at 500V d.c. and is designed to meet vibration and shock to MIL-STD-202B and acceleration to MIL-R-27208A. It should have an electrical life of at least 1000 hours under full load. A slipping clutch is fitted to protect the unit at end of travel.

*For more information circle* **No. 320**

## NUMERICAL INDICATOR TUBES

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

*(Illustrated above right)*




Two new cold-cathode numerical indicator tubes announced by Mullard Ltd make possible more compact digital read-out systems in calculating machines, electronic measuring instruments and instrumentation and control systems. Details as follows:

*ZM1162*

The ZM1162 is a ten-digit, end-viewing, numerical indicator tube. Its rectangular shape within minimum dimensions ensures that compact stacking in both the vertical and horizontal plane can be achieved without any compromise in performance. The characters are approximately 15mm high.

A 14-pin plug-in type, the ZM1162 can be used as an advantageous replacement for many of the rectangular tubes at present being imported into the U.K. Overall dimensions are 27mm high × 20mm wide × 23mm seated depth. Under typical d.c. operating conditions the tube requires a minimum d.c. anode voltage of 170V. The tube may also be operated from a half-wave single-phase supply. The nominal cathode current is 2mA—a current level at which drive transistors such as the Mullard type BSX21 are suitable.

The ZM1162 may be controlled by either mechanically or electronically switching the supply voltage to its individual cathodes.

*ZM1172*

Side-viewing ZM1172 indicator tubes are the Mullard compromise between optimum stacking and minimum price. Stacking in the horizontal plane is within dimensions similar to those of the new rectangular types, although in the vertical plane more space is required—a feature which, for in-line read-outs, is usually unimportant. The characters are approximately 15mm high. Electrical performance and readability is as for the ZM1162. Overall dimensions are 9mm diameter × 47·5mm seated height.

The ZM1172 also has a special fine-mesh sputter guard which provides maximum protection for the viewing window combined with minimum obstruction of the viewing area. If required, a low-cost plastic support (type 56022) can be supplied to enable these tubes to be mounted on the edge of a typical printed circuit board.

Both of the new tubes have conservative life figures similar to those quoted for other Mullard gas-filled indicator tubes. However, when considering this minimum figure, it should be appreciated that this is the result of strict laboratory tests. The test requirements consist of illuminating a single digit under maximum current conditions and then, after a specified period, carrying out a check for discernible loss in illumination under minimum current conditions. Continuously illuminating the same digit is a particularly severe measure because it eliminates the self-cleaning effect produced when, as under normal working conditions, different digits are illuminated in turn.

The stringent test conditions applied to representative samples ensure that, under normal operating conditions, users will obtain the maximum possible life that can be achieved for tubes of this type.

*For more information circle* **No. 321**

## DIGITAL VOLTMETER

**Fairchild Instrumentation Ltd, Grove House, London Road, Isleworth, Middlesex**

*(Illustrated below)*

Use of integrated circuits has allowed Fairchild Instrumentation Ltd to announce a new low-price four-digit integrating digital voltmeter with 0·01 per cent accuracy.

The new model 7000 has complete front panel control of all functions and ranges and uses function cards instead of expensive plug-ins for its extended capabilities. The model 7000 combines maximum performance with high


accuracy and is suited for use in laboratory, production and general testing situations.

Using the dual slope integrating technique, the model 7000 combines the excellent noise rejection capability of integration with accuracy and stability of automatic comparison to an internal source.

Read-out is made with bright, easy-to-read in-line indicators. Four full decades plus a fifth digit give full scale read-out of 10 000 with display storage. Automatic polarity, decimal point and measurement units are standard features. Over-ranging of 20 per cent is provided with no degradation of accuracy.

The basic instrument provides d.c. voltage capability. Options in the form of plug-in boards extend the capability to a.c. voltage, resistance, d.c. current measurement, auto-ranging and b.c.d. output.

*For more information circle* **No. 322**

# SHORT NEWS ITEMS

**The Control and Automation Division** of the Institution of Electrical Engineers has arranged a Summer School on Control Theory to be held at the Renold Building of the University of Manchester Institute of Science and Technology on 3 to 14 July this year.

The School is designed to introduce practising engineers to recent theoretical developments in control engineering. An intensive course of lectures and tutorials will be given on the fundamental mathematical and dynamical theory required. This will provide a background for later study of the current literature, and for a special series of survey lectures on the main topics of current interest.

Enquiries concerning registration should be addressed to the Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

**A revised M.Sc. course** in quantum electronics is to start at the University of Southampton in October this year and will cover the following additional subjects:

- Lasers
- Opto-electronics
- Optical communications
- Holography
- Non-linear optics
- Masers

The course is normally open to university graduates with suitable science or engineering degrees (e.g. in electronics, physics, mathematics, electrical engineering), but in special cases candidates with other qualifications approved by the Professors of Electronics may be considered. The course is recognized by the Science Research Council for the award of postgraduate course studentships.

Applications for the course should be made as soon as possible to:

The Academic Registrar, The University, Southampton, Hampshire.

**Esse Metotest AB**, Sweden, has taken over the Industrial Ultrasonics Non-Destructive Testing business of Kelvin Electronics Company of Glasgow. The takeover includes the right to manufacture and sell throughout the world non-destructive testing equipment, using ultrasonic techniques, under the current relevant Smiths Patents.

Metotest, a subsidiary of Svenska Metallverken of Sweden, is already manufacturing and marketing a wide range of non-destructive testing equipment using eddy current and other techniques.

**A Summer Course** is to be held on 4 to 15 September this year at the University of Louvain. Belgium, by the Electronics Laboratories, Microwave Department. The topics to be covered are: guided waves, obstacles in waveguides, cavities, junctions and passive non-reciprocal devices, periodic structures and filters. Several seminars will be held, covering measurement techniques, microwave tubes, ferrites, low-noise devices and their use in spatial communications, microwave solid state and industrial applications.

The fee for the course is $200 (or equivalent) and further information can be obtained from:

Prof. A. S. Vander Vorst, Electronics Laboratories, 94 Kardinaal Mercierlaan, Heverlee—Leuven, Belgium.

**The fourth edition** of the Handlist of basic reference material for librarians and information officers in electrical and electronic engineering has recently been published by the ASLIB Electronics Group.

Copies are available at 10s for ASLIB members (15s for non-members) on application to:

R. S. Lawrie, Chief Librarian, Sperry Gyroscope Division, Sperry Rand Ltd, Downshire Way, Bracknell, Berkshire.

**The Cambridge Instrument Co. Ltd** has acquired the whole of the share capital of H. W. Sullivan Ltd for a consideration of 100 000 Cambridge 5s Ordinary shares ranking *pari passu* with the 5 166 795 Cambridge Ordinary shares in issue. The book value of Sullivan's net assets on 31 March, 1967 is estimated at £133 000.

Cambridge Instruments, founded 80 years ago have factories at Cambridge, Finchley, Muswell Hill and Richmond. There are Cambridge companies in the USA, France, Italy, India, Australia and Canada.

They produce measuring, controlling and recording instruments for a wide range of industrial application, including special gas analysers for fuel efficiency and high pressure steam generating plant, instruments for measuring physiological phenomena in medicine, and highly complex electron probe instruments, such as the Stereoscan scanning electron microscope.

H. W. Sullivan Ltd founded in 1895, has a new factory at Orpington, Kent, and foremost among its products are inductance, capacitance and resistance standards.

**The total sterling value** of sales of British radio valves, tubes and semiconductor devices for 1966 shows an increase over those for 1965, according to figures released by the British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA), as follows:

| | *1966* £(M) | *1965* £(M) |
|---|---|---|
| Valves & Tubes | 47·3 | 46·1 |
| Semi-conductor devices | 31·7 | 28·1 |
| | 79·0 | 74·2 |

The figures for the last quarter of 1966 were:

| | £(M) |
|---|---|
| Valves & Tubes | 11·5 |
| Semi-conductor devices | 7·6 |
| | 19·1 |

Of the total for 1966, according to figures previously released, export sales accounted for nearly £13¾M—higher than ever before.

**Advance Electronics Ltd** is offering an annual grant of £1 000 to finance a research studentship at Southampton University, Department of Electronics for research connected with electronic instrumentation and control.

Advance will be holding interviews very shortly and applicants should already have reached a standard of at least an upper 2nd class honours degree.

Applications should be addressed to:

Research Scholarship, Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex.

**G.E.C. Road Signals Ltd and The Marconi Co. Ltd** have agreed to collaborate in the field of computer controlled area road traffic signalling schemes in the UK and overseas.

G.E.C. Road Signals has a comprehensive range of road signalling equipment including traffic lights, controllers and detectors. Contracts recently won for the West London Area Traffic Experiment included the provision of the Queue Detectors, Closed Circuit Television and much other equipment.

The Marconi Company which has been involved in air traffic control has developed a Myriad microelectronic computer and has recently entered the field of road traffic control.

This Myriad computer has been chosen by the Road Research Laboratories of the Ministry of Transport to provide the central process control facilities for an experimental system in Glasgow, which is aimed at assessing different methods of area traffic control. It forms part of a major programme by the Ministry of Transport, to solve the increasing problem of traffic congestion in Britain's major towns and cities.

The Glasgow Computer Centre is located in Broomielaw, in central Glasgow, and some 80 traffic signals with associated traffic detecting sensors will be connected to the computer in an area covering about a square mile in the crowded central business and shopping district, an area which includes four bridges over the Clyde. Various types of traffic sensor will be used, and the computer system has to be able to accept 'real time' information from a wide variety of sources.

Comprehensive area traffic schemes will now be put forward by G.E.C. Road Signals Limited, who will provide the road signalling systems, while Marconi will provide the central data processing systems. This collaboration will also provide customers with a wide and competitive choice of associated equipment, such as closed circuit television, from the ranges of the two Companies or from other sources.

The joint activity will be centred on the offices of G.E.C. Road Signals Ltd, at East Lane, Wembley.

**A 6 000 mile undersea telephone and telegraph cable** carrying 360 telephone circuits between Cape Town, South Africa and Lisbon, Portugal, is to be manufactured and laid by Standard Telephones and Cables Ltd.

The value of the order is estimated at £22M.

The cable will be equipped with 643 repeaters at intervals of 9½ nautical miles and 51 equalizers. It will travel from Cape Town to Ascension Island (2 554n. miles) then to Cape Verde Island (1 723n. miles), Canary Islands (902n. miles) and finally to Lisbon (762n. miles).

At the Canary Islands, the cable will have connexions with the 160 circuit cable to Cadiz, Spain, supplied by S.T.C.

The cable will be of the single coaxial form and the deep-water section will consist of a central strength member clad with a welded copper tube conductor, a layer of polyethelene dielectric and outer conductor made from a single copper tape. The outer conductor, 1in in diameter, is covered in a layer of high density polyethelene to a diameter of 1¼in.

In shallow water, at depths up to about 40 fathoms, heavy armouring wires are used to prevent damage by such hazards as currents, chafing and ships' anchors.

Two-way working will be achieved by using separate 1MHz wide bands of frequencies: 312 to 1 428kHz in one direction and 1 848 to 2 964kHz in the other. Each band accommodates 360 telephone channels 3kHz wide.

At Cape Town, the Canaries and Lisbon, multiplexing and other electronic equipment will 'stack' the channels, dropping circuits off as necessary at the Canaries. The equipment includes fully duplicated carrier supply equipments, supergroup translating equipment, group translating equipments, and 3kHz channel translating equipments. For dropping off circuits at the Canaries and for routeing beyond Lisbon 'through supergroup' and 'through group' filter equipments will be used.

The process of stacking the speech channels into the transmission spectra will be performed mainly by S.T.C.'s new Mk6 multiplexing equipment.

Groups of 16 speech channels form the starting point for the translation. A separate carrier is applied to each of the 16 channels in each group so that, by choosing the appropriate sideband resulting from each modulation process, translated groups of 16 channels re-appear spanning the range 60 to 108kHz. There being 360 channels, this process results in $360/16 = 22\frac{1}{2}$ groups, each spanning 60 to 108kHz. These are taken 5 at a time (once at 2½), and to each group a separate carrier is once again applied to produce a translated set of groups each within the band 312 to 552kHz. Each of the 4 sets, and the half set, forms a 'supergroup'. Finally, each of the supergroups undergoes another, similar translation process to place the set of supergroups either in the 312 to 1 428kHz band for one direction of transmission, or 1 848 to 2 964kHz for the reverse direction.

The repeaters, using thermionic valves, have an expected life of over 20 years, and consist of only one, wideband, amplifier for the high and low frequency transmission bands moving in opposite directions through the repeater, although in fact two amplifiers are used in parallel with a common feedback path. If one amplifier path fails, the gain of the complete amplifier is not appreciably affected.

40dB of overall voltage and current feedback is applied to both amplifiers and a fault within either changes the overall gain by less than 0·1dB. Each amplifier has three stages and in conjunction with feedback and other components a gain characteristic that rises suitably with frequency is obtained.

Power for the repeaters will be fed into the cable over the centre conductor at Cape Town, Ascension and Santa Cruz in the Canaries, at voltages up to 10kV using accurately stabilized constant supply equipment.

**A Marconi-Thomson Secar** secondary radar system is to be despatched to the Vnokovo Airport near Moscow for a six months evaluation trial following negotiations with the Soviet State Committee for Science and Technology.

Vnokovo is one of the two main airports of Moscow and is the site of the air traffic control centre.

Secar was developed and is manufactured jointly by the Marconi/Thomson Consortium, formed from The Marcóni Company and Compagnie Francaise Thomson-Houston Hotchkiss-Brandt of France. The system is designed to extract information from an aircraft in flight to supplement the normal radar position finding function. This is achieved by transmitting a number of interrogation signals to the aircraft from a ground station. These signals are received in the aircraft by a transponder which automatically transmits an appropriate reply in the form of a digital code.

Secar systems have been supplied to Eurocontrol for installation at the regional air traffic control centres at Brussels and Shannon for monitoring air traffic in the upper air space.

**The Post Office** has placed a £200 000 contract with Redifon Ltd for the supply and installation of equipment for a new high-power v.l.f. transmitter (GBZ) at Criggion, in Wales. This follows the recent re-opening of the similarly-modernized v.l.f. transmitter at Rugby (GBR).

The new Criggion transmitter will be of similar design to that provided at Rugby. It will enable the output to be increased from 200kW to 500kW and permit F1 as well as A1 modulation at telegraph speeds up to 50 baud. The present aerial system would be too small to sustain the large voltages produced and a new aerial system covering twice the present area and requiring three further 700ft masts is being provided. When completed in 1968, the effective power of the station will be increased by four times.

**A new type of radio terminal equipment,** known as Lincompex (*Lin*ked *Com*pressor and *Ex*pander), which will greatly improve the quality and efficiency of radio telephone circuits, is now going into service on a number of high-frequency circuits operated by the General Post Office of the United

Kingdom and the American Telephone and Telegraph Company.

The new system largely eliminates variations in speech volume and timbre caused by fading of the radio circuit and is very little affected by radio noise, which is effectively suppressed between syllables of speech. It also dispenses with the anti-singing device which seriously interferes with the smooth flow of conversation in the conventional terminal system and which causes premature shut-down of the circuit when radio noise is present.

The equipment is already in operation on circuits provided between New York and Uruguay and also on circuits to Argentine, Brazil, Chile and Peru. The GPO will shortly introduce the equipment on circuits between London and India, Ceylon, Kenya, Nigeria, South Africa and other overseas terminals.

Lincompex equipment was developed and designed by the GPO and is manufactured in the United Kingdom to a specification jointly agreed between engineers of the GPO and the American Telephone and Telegraph Company. It has something in common with the compander systems used on long-distance cable circuits in that the speech signal is heavily compressed. This achieves a high level of modulation of the transmitter irrespective of speech amplitude and thus helps to minimize the effects of radio noise. The information required for restoring the original variations in speech amplitude is passed at syllabic rate to complementary expanders at the receiving end of the circuit by means of a narrow-band frequency-modulated control signal contained within the normal speech channel. The overall bandwidth requirements are, therefore, unchanged and modern independent sideband radio equipment will accept one or more Lincompex channels.

GEC (Electronics) Ltd has received an order valued at over £100 000 for Lincompex terminal equipment from the USA.

**An International Symposium,** organized by the International Federation of Automatic Control (IFAC), is to be held at Haifa, Israel on 11 to 14 September.

The Conference will deal with the application and utilization of control of natural resources and public utilities, with subjects ranging from gas supplies to traffic lights and refineries.

**A new ultrasonic research facility** has been installed at the Feltham research laboratories of EMI Electronics Ltd to be used primarily for investigating the radar cross-section of aircraft and other radar targets.

The Feltham facility includes a small tank for experimental use and a large concrete pool measuring 32 × 22 × 14ft, containing 60 000 gallons of water. This pool is believed to be the largest of its kind in Britain. It houses the ultrasonic source which is capable of movement in three axes and the target which has freedom of movement in four axes. A high degree of positional accuracy is achieved by closed loop servos coupled to an EMICON B100 numerical control system. Control can be either manual or by punched tape input from a central console. Closed circuit television enables the operator to view clearly the source and the target.

In addition to the programme of work on radar cross-section EMI Electronics indicate that time will be available on the facility for investigations involving more conventional use of ultrasonics such as sonar and oceanography.

**G.E.C. (Telecommunications) Ltd,** of England, has won a contract worth over £500 000 for a 450-mile microwave radio link between Tripolis in southern Greece and Zakros on the island of Crete.

The new link will go via the islands of Kythira and Antikythira. Semiconductored equipment will be used throughout: 6 000MHz for the main section and 7 000MHz for 'spur' routes on Crete.

The former is suitable for transmitting up to 960 telephone circuits or television, and the latter can carry a maximum of 300 circuits. In both cases there will be a standby channel which will take over automatically should the working channel fail.

The company's microwave systems already link Patras with the island of Corfu, Athens with Mt. Parnis and the island of Syros, and Larrissa with Mt. Pillion.

**Datafair 67,** organized by the British Computer Society, is to take place at Southampton University on 26 to 29 September.

There will be a symposium of about 80 papers in about three parallel sessions—hardware, software and applications—taking place in the afternoons of each day.

During the mornings four parallel events are to be set up, presentations by computer and equipment manufacturers, consultants, service bureaux, etc.; British Computer Society discussion groups, university presentations; visits. In the evenings there will be similar parallel events of presentations, visits and discussion groups.

Three special one-day seminars have been organized during the week. A discussion on computer education in schools on Tuesday, 26 September, will be followed on Wednesday, 27, by a seminar on computer studies on computer education. The final seminar on Thursday, 28 September, is to deal with the computer as it affects business management.

Accommodation for 950 delegates has been arranged at the Southampton University halls of residence and allocation of rooms will be made by the British Computer Society. Further details of Datafair may be obtained from The British Computer Society, 23 Dorset Gardens, London, N.W.1.

**A microwave print drier** has recently been installed by Coventry Newspapers. The drier was developed and manufactured by a joint company Eden Fisher & Hirst (Microwaves) and at Coventry it is applied to a Halley-Aller web offset machine with four-colour newsprint. The web (the continuous sheet of newsprint) is passed between the two halves of a waveguide made up of a labyrinth with a total length of 60ft. The speed is 900ft/min but future development is aimed at increasing this to 2000ft/min.

The magnetron power source operates at 2·4GHz with a c.w. output of 5kW. The waveguide labyrinth is terminated in a water load so that changes in working load do not cause reflection back to the magnetron; the water also provides cooling for the valve. The drying process depends on the differential dielectric loss between the paper and the coloured inks, so that the paper is minimally heated, most of the energy being absorbed by the inks. In this way the hitherto intractable problem of the paper becoming too dry, and therefore reduced in elasticity, is avoided. Stabilizers set opposite each other at intervals across the whole length of the paper in the drier maintain the web in position. Two systems of stabilization may be employed: either pneumatic pressure on the stabilizer face or by means of two guide path strips which have a low coefficient of friction and a low dielectric constant.

The waveguide labyrinth is so constructed that the top half hinges upwards for inspection or starting a run. If the drier is opened in this way when microwave power is present a microswitch reduces the e.h.t. to the magnetron so that no danger occurs from mismatching and there is, moreover, no microwave field hazardous to the operator. The total leakage of power under normal operating condition is about 1mW, approximately 1/10 of the maximum laid down in safety regulations.

**The Wayne Kerr Co. Ltd,** which specializes in electronic measurement, is to endow a chair in measurement science at the new University of Surrey. The chair will be in the Department of Chemical Physics, headed by Professor V. S. Griffiths. The holder of the chair, who is yet to be appointed, will devote the bulk of his time to research and will lead the research team which is to carry out, under the Ministry of Technology contract announced in April, an investigation into the establishment of measurement standards and techniques in the radio frequency field.

For more information circle No. 111

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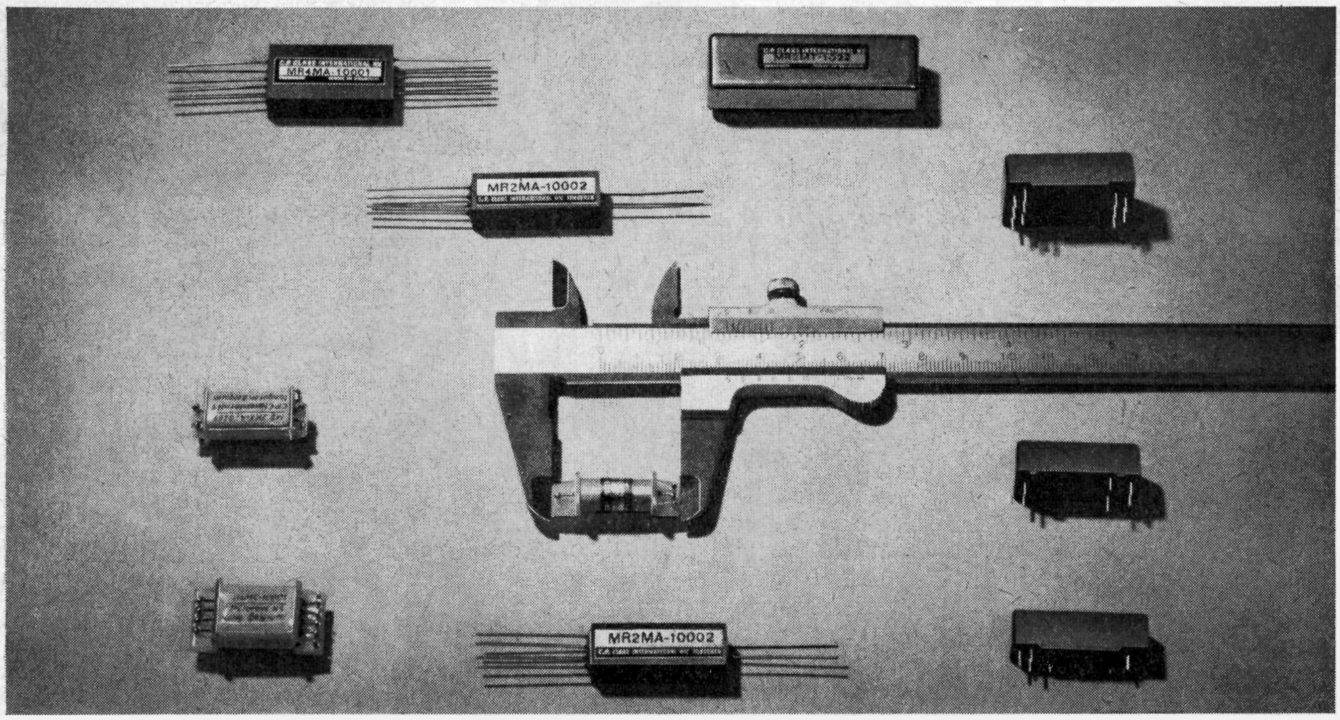


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# Résumés des Principaux Articles

**Calcul des courbes V-1 de thermistors NTC** par M. R. McCann

Résumé de l'article aux pages 346 à 351

*Lorsqu'on augmente légèrement le courant passant à travers un thermistor N.T.C., la tension correspondante augmente jusqu'à un maximum de $E_{max}$ puis commence à baisser. Cette caractéristique est due à l'auto-échauffement du thermistor.*

*Les formules de base comprenant la résistance, la température et la puissance sont utilisées pour obtenir un diagramme indiquant des courbes de tension et de courant normalisées pour des thermistors à coefficient de température négatif. A partir de ces courbes, de la connaissance de la résistance à froid et de $E_{max}$, on peut déterminer la caractéristique de tension-courant du thermistor dans n'importe quelle température ambiante. Un exemple de ce calcul est indiqué pour un thermistor monté dans des conditions d'environnement différentes. L'analyse comprend une méthode simple pour calculer $E_{max}$ (sur la caractéristique V-I) et sa variation en fonction de la température ambiante. On obtient une ligne de charge résistive universelle pouvant être utilisée en liaison avec des caractéristiques de tension-courant reportées à l'aide d'échelles logarithmiques.*

*Plusieurs exemples de caractéristiques de tension-courant de différents types de thermistors (caractéristiques mesurées dans des conditions d'utilisation effective) sont indiqués et comparés à des caractéristiques calculées à partir de courbes normalisées.*

*L'analyse est étendue à des thermistors montés sous vide dont le comportement subit une dérivation considérable par rapport à celui de thermistors montés dans d'autres milieux.*

**Un amplificateur à transistor et à large bande de 50 à 500 MHz** par A. E. Hilling

Résumé de l'article aux pages 352 à 355

*L'amplificateur à large bande dont il est question dans cet article a un gain de 21,5 ±1,5dB de 50 à 500MHz. Un indice de bruit maximum de 8dBs' obtient à 500MHz et les rapports maxima d'amplitude de tension d'entrée et de sortie sont de 2,0:1 pour 50Ω. Le facteur de distorsion d'intermodulation de troisième degré est maintenu à un faible niveau par l'emploi du transistor Mullard BFY90.*

**Circuit de blocage de comparateur à paire de diodes Esaki couplées et se faisant face** par Y. Murata

Résumé de l'article aux pages 356 à 361

*Par l'emploi du circuit à paire de diodes Esaki couplées et se faisant face on obtient des circuits à mémoire et à fonction de comparaison d'une application étendue.*

*Le fonctionnement du circuit à paires de diodes Esaki couplées et se faisant face a été étudié au moyen de plusieurs expériences ainsi que par l'analyse numérique à l'aide d'une calculatrice numérique. Les limites de fonctionnement du circuit en ce qui concerne la charge, l'entraînement, la vitesse de réponse, etc. ont été déterminées dans le cas des diodes Esaki SONY, type 1T1101. Il a été constaté que le circuit pouvait servir d'élément à mémoire de qualité ainsi que d'élément de commutation à action rapide.*

**Un circuit électronique à transistor de jonction** par T. K. Cowell

Résumé de l'article aux pages 362 à 366

*La résistance d'entrée maxima pouvant être obtenue avec des circuits à transistors de jonction conventionnels est limitée par la résistance de fuite du collecteur et de la base ainsi que par le gain de courant effectif aux courants de fonctionnement nécessairement faibles.*

*Le montage du circuit de base décrit dans cet article, soit un circuit complémentaire à charge cathodique par transistor et à anneaux de garde au collecteur, obvie à ces limites et peut donner, s'il est muni d'étages en cascade, des résistances d'entrée dépassant 1 million de mégohms.*

**Un circuit monostable à haute stabilité et à cycles de régime élevés** par R. C. French

Résumé de l'article aux pages 372 à 373

*L'auteur décrit un circuit monostable capable d'assurer des cycles de régime pouvant atteindre 99% et produisant une impulsion dont la longueur demeure stable à 1% près par rapport à une variation de tension d'alimentation de 10%.*

*Un changement de longueur d'impulsion ne dépassant pas 1,5% se produit lorsque le cycle de régime varie entre 0 et 99%.*

**Un inverseur stable de 50Hz pour matériel à fonctionnement secteur** par W. T. Maloney

Résumé de l'article aux pages 374 à 375

*L'auteur décrit un inverseur d'une bonne stabilité de fréquence aux fréquences secteur. L'inverseur fonctionne sur alimentation de 12V et fournit une sortie d'environ 40W à 240V. Il est démontré que moyennant quelques légères modifications un fonctionnement sûr peut être obtenu à partir d'une source de 6V.*

**La représentation sur tube cathodique d'indications de vitesse proportionnelle** par M. R. Green et K. Lord

*Résumé de l'article aux pages 376 à 380*

*Cet article décrit des circuits permettant de présenter des informations au moyen de deux faisceaux électroniques sur un tube à rayons cathodiques. Ce dernier a donc les mêmes fonctions qu'un instrument à aiguilles ordinaire mais ne comporte, évidemment, aucune pièce mobile mécanique. Le modèle cité est un appareil de bord fournissant des indications d'altitude et de vitesse verticale sur un même écran.*

**Un convertisseur analogique numérique simple à lecture réciproque** par L. Davison et R. Wilson

*Résumé de l'article aux pages 381 à 383*

*Ce convertisseur constitue un moyen de lecture pratique et direct de la vitesse des compteurs numériques dont le contenu représente des temps transitoires. Non seulement il évite les pertes de temps lorsqu'il s'agit d'indications de résultats comportant plusieurs chiffres mais il élimine également toute possibilité d'erreur résultant d'une fausse lecture du compteur, d'une panne de la lampe de lecture ou d'une erreur de calcul.*

**Un filtre à T jumelé et à fréquence médiane réglable** par K. G. Beauchamp

*Résumé de l'article aux pages 384 à 387*

*Une difficulté majeure dans la réalisation de filtres à T jumelé couvrant une gamme étendue de fréquences est d'assurer un moyen simple de réglage de fréquence tout en ne réduisant pas, en même temps, le facteur Q de façon drastique.*

*Dans le modèle décrit dans cet article, une variation d'un seul paramètre permet de contrôler la fréquence avec un effet très réduit sur le facteur Q.*

**Ungénérateur précis d'ondes triangulaires à balayage de fréquence étendu** par G. Klein et H. Hagenbeuk

*Résumé de l'article aux pages 388 à 390*

*Cet article décrit un générateur d'ondes triangulaires qui allie un balayage étendu de fréquence ($\simeq 10^4$) à une forme d'onde symétrique très précise. La réponse d'amplitude et la symétrie donnent des variations inférieures à 0,1%, la stabilité de fréquence étant supérieure à 1:10⁴. Le circuit peut servir de base à un oscillateur à ondes sinusoïdales à balayage unique ou de convertisseur de fréquence de tension (courant) et de modulateur de fréquence.*

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# Zusammenfassung der wichtigsten Beiträge

**Die Berechnung von U-I-Kennlinien für NTC-Thermistoren** von M. R. McCann

*Zusammenfassung des Beitrages auf Seite 356-351*

*Wenn der durch einen Thermistor fliessende Strom langsam erhöht wird, steigt die entsprechende Spannung, bis ein Höchstwert $U_{max}$ erreicht ist, und beginnt dann zu fallen. Die Eigenschaft beruht auf der Selbsterwärmung des Thermistors.*

*Mit Hilfe der Widerstand, Temperatur und Leistung einbegreifenden Grundformeln kann man eine grafische Darstellung erstellen, deren Kennlinien normierte Spannungen und Ströme für Thermistoren mit negativem Temperaturkoeffizienten geben. Diese Kennlinien und die Kenntnis des kalten Widerstandes sowie $U_{max}$ gestatten Bestimmung der Spannung-Strom-Eigenschaften des Thermistors für jede beliebige Umgebungstemperatur. Solche Berechnungen werden beispielsweise für einen Thermistor in verschiedenen Umgebungsbedingungen vorgenommen. In der Analyse wird u.a. ein einfaches Verfahren zur Berechnung von $U_{max}$ (an der U-I-Kennlinie) und deren Änderung mit der Umgebungstemperatur gegeben. Für die mit logarithmischen Massstäben aufgetragenen Spannung-Strom-Kennlinien wird eine allgemeine ohmsche Belastungslinie abgeleitet.*

*Die unter praktischen Bedingungen gemessenen Spannung-Strom-Kennlinien verschiedener Thermistorentypen werden als Beispiele gebracht und mit den auf Grund der normierten Kurven berechneten Eigenschaften verglichen.*

*Die Analyse wird auf im Vakuum arbeitende Thermistoren ausgedehnt, deren Verhalten wesentlich von dem in anderen Medien betriebener Thermistoren abweicht.*

**Ein Breitband-Transistorverstärker für 50 . . . 500 MHz** von A. E. Hilling

*Zusammenfassung des Beitrages auf Seite 352-355*

*Ein beschriebener Breitbandverstärker hat von 50 . . . 500 MHz eine Verstärkung von 21,5 ±1,5 dB. Ein Höchstrauschmass von 8 dB wird bei 500 MHz erreicht, und das maximale Eingangs- und Ausgangsstehwellenverhältnis ist—auf 50 Ω bezogen—2,0:1. Durch Anwendung des Transistors BFY90 werden niedrige ZM-Faktoren dritter Ordnung erzielt.*

**Vergleichshalteschaltung mit gegengekoppelten Esaki-Diodenpaaren** von Y. Murata

Zusammenfassung des Beitrages auf Seite 356-361

*Durch Anwendung gegengekoppelter Esaki-Diodenpaarschaltungen lassen sich breit anwendbare Schaltungen mit Speicher- und Vergleichsfunktionen erstellen.*

*Die Arbeitsweise von Schaltungen mit gegengekoppelten Esaki-Diodenpaaren wurde in mehreren Experimenten untersucht und mit Hilfe eines Digitalrechners numerisch analysiert. Die Betriebsgrenzen der Schaltung wurden in bezug auf Belastung, Steuerung, Ansprechgeschwindigkeit usw. für die SONY-Esaki-Dioden 1T1101 geklärt. Es wird gezeigt, dass die Schaltung als gutes Speicherelement und sehr schnelles Schaltelement Anwendung finden kann.*

**Eine Flächentransistor-Elektrometerschaltung** von T. K. Cowell

Zusammenfassung des Beitrages auf Seite 362-366

*Der höchste mit einer herkömmlichen Flächentransistor-Schaltung erreichbare Eingangswiderstand ist durch den Isolationswiderstand zwischen Kollektor und Basis und die bei den unvermeidlich niedrigen Betriebsströmen reduzierte wirksame Stromverstärkung begrenzt.*

*Diese Begrenzungen werden durch die beschriebene Grundschaltungsanordnung—ein komplementäres Emitterfolgerpaar mit abgeschirmtem Kollektor—überkommen; mit hintereinandergeschalteten Stufen können Eingangswiderstände von über ein Teraohm erreicht werden.*

**Ein hochkonstanter monostabiler Oszillator mit hoher Impulskennziffer** von R. C. French

Zusammenfassung des Beitrages auf Seite 372-373

*Eine beschriebene monostabile Schaltung kann mit Impulskennziffern bis zu 99 betrieben werden und erzeugt Impulse, deren Dauer bei Speisespannungsschwankungen von 10 Prozent innerhalb 1 Prozent konstant bleibt. Bei Änderung der Impulskennziffer von 0 . . . 99 schwankt die Konstanz der Impulsdauer nur 1,5 Prozent.*

**Ein 50-Hz-Konstantwechselrichter für netzbetriebene Geräte** von W. T. Maloney

Zusammenfassung des Beitrages auf Seite 374-375

*Der beschriebene Wechselrichter hat bei Netzfrequenzen gute Frequenzkonstanz. Er arbeitet mit einer Speisespannung von 12 V und gibt bei 240 V etwa 40 W ab. Es wird gezeigt, dass mit nur geringfügigen Abwandlungen auch bei Speisung mit 6 V zuverlässiger Betrieb erreicht werden kann.*

**Darstellung von Proportional- und Geschwindigkeitsinformation auf einem Oszillografenschirm** von M. R. Green und K. Lord

Zusammenfassung des Beitrages auf Seite 376-380

*In diesem Beitrag werden Schaltungen beschrieben, die Darstellung von Information auf einem Oszillografenschirm in Form von zwei Zeigern gestatten. Das Gerät übt daher die Funktion eines Zeigerinstrumentes aus, in dem es jedoch keine mechanisch beweglichen Teile gibt. In der als Beispiel beschriebenen Ausführung als Flugzeug-Bordgerät ist das Gerät für die gleichzeitige Anzeige vuo Höhe und senkrechter Geschwindigkeit bestimmt.*

**Ein einfacher Digital-Analogumsetzer mit Reziprokwertanzeige** von L. Davison und R. Wilson

Zusammenfassung des Beitrages auf Seite 381-383

*Der beschriebene Umsetzer ist ein zweckdienliches Mittel zur Direktanzeige der Geschwindigkeit von Binärzählern, deren Inhalt Fortpflanzungszeiten darstellt. Durch den Umsetzer wird nicht nur bei der grossen Anzahl der anfallenden Ergebnisse Zeit eingespart, sondern es werden auch die durch Fehlablesung von Zählern, Ausfall von Anzeigeröhren und Berechnungsirrtümer auftretenden Fehlermöglichkeiten beseitigt.*

**Ein Doppel-T-Filter mit verstellbarer Mittelfrequenz** von K. G. Beauchamp

Zusammenfassung des Beitrages auf Seite 384-387

*Eine der Hauptschwierigkeiten beim Entwurf eines Doppel-T-Filters für einen Frequenzenbereich liegt in der Konstruktion einfacher Mittel zur Frequenzverstimmung, die nicht gleichzeitig den Gütefaktor scharf reduzieren.*

*In dem beschriebenen Entwurf erfolgt die Frequenzverstellung durch Änderung eines einzigen Parameters und mit nur geringem Einfluss auf den vorgegebenen Gütefaktor.*

**Ein Präzisionsdreieckwellen-Generator mit grossem Frequenzhub** von G. Klein und H. Hagenbeuk

Zusammenfassung des Beitrages auf Seite 388-390

*In diesem Dreieckwellengenerator wird breiter Frequenzhub ($\simeq 10^4$) mit sehr genauer symmetrischer Wellenform kombiniert. Die Abweichungen des Amplitudengangs und der Symmetrie liegen innerhalb 0,1 Prozent, und die Frequenzkonstanz ist besser als $1 \times 10^{-4}$. Die Schaltung kann als Grundlage für einen Sinuswellenoszillator mit einmaligem Durchlauf oder einen FM-Modulator und Spannungs-(Strom-) Frequenzumsetzer dienen.*

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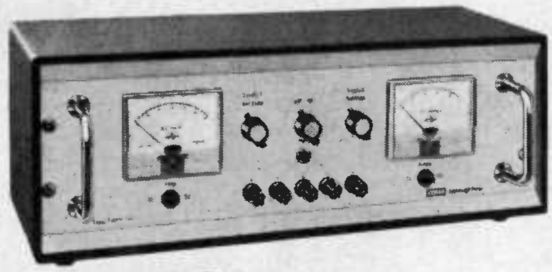

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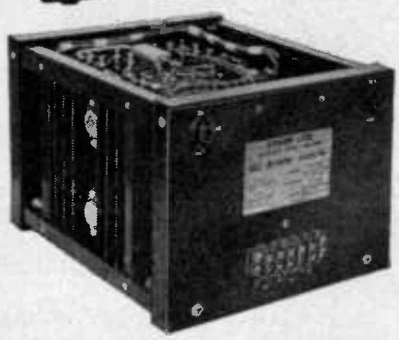

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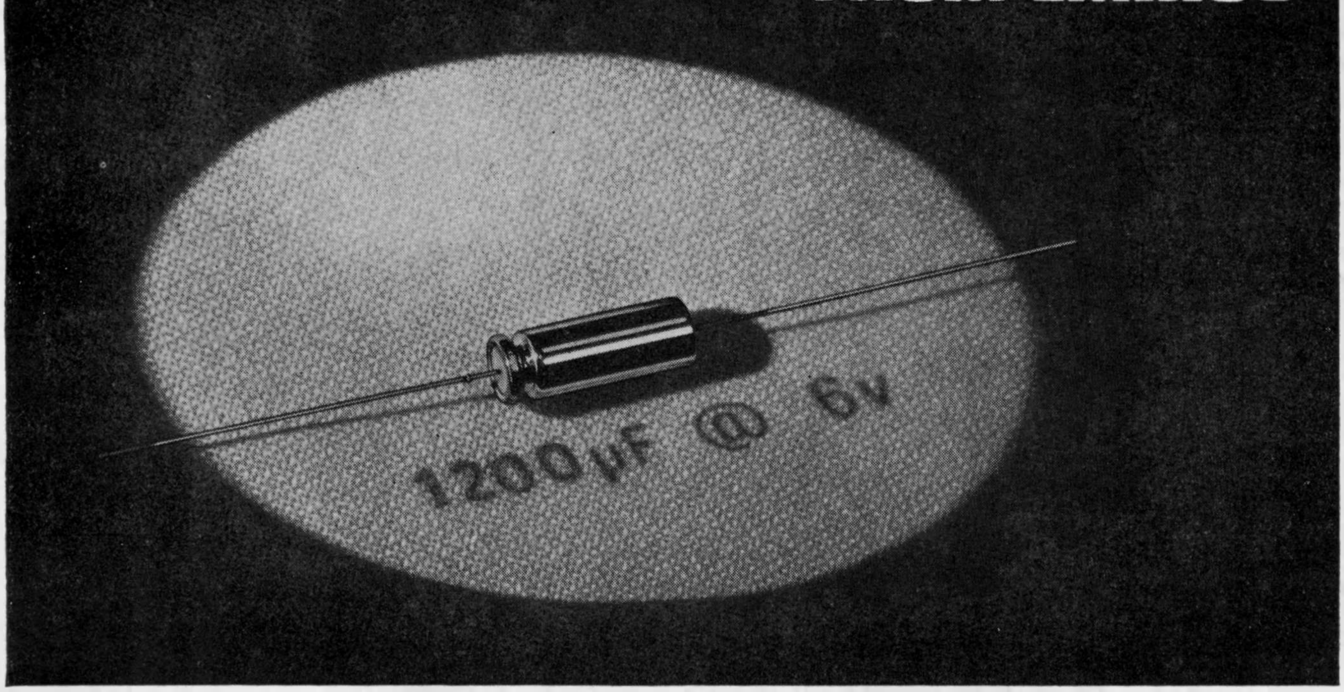


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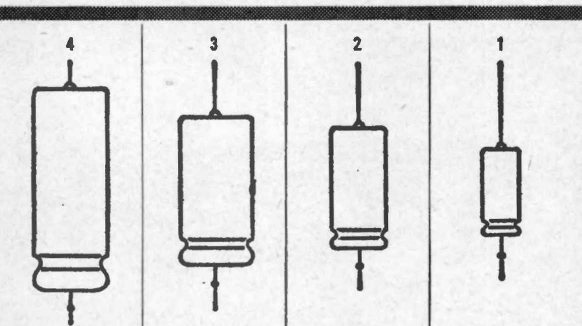

| 4 | 3 | 2 | 1 |
|---|---|---|---|
| From 80μF @ 125V<br>To 1200μF @ 6V | 25μF @ 125V<br>560μF @ 6V | 9μF @ 125V<br>270μF @ 6V | 1.7μF @ 125V<br>88 μF @ 6V |
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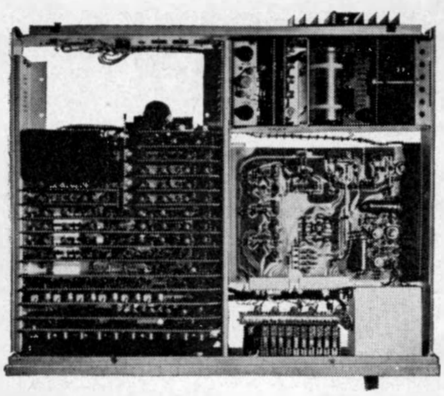
Including: A 100 MHz *DIRECT* Frequency Meter, three 500 MHz Automatic Heterodyne Converters, a 1 MHz preset Counter/Timer, a VHF/UHF calibrator using state-of-art sampling techniques. Others now in final development stage will follow quickly.



800 series information and a demonstration date from Brian Cuthew,

RACAL

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Telephone: Crowthorne 2273
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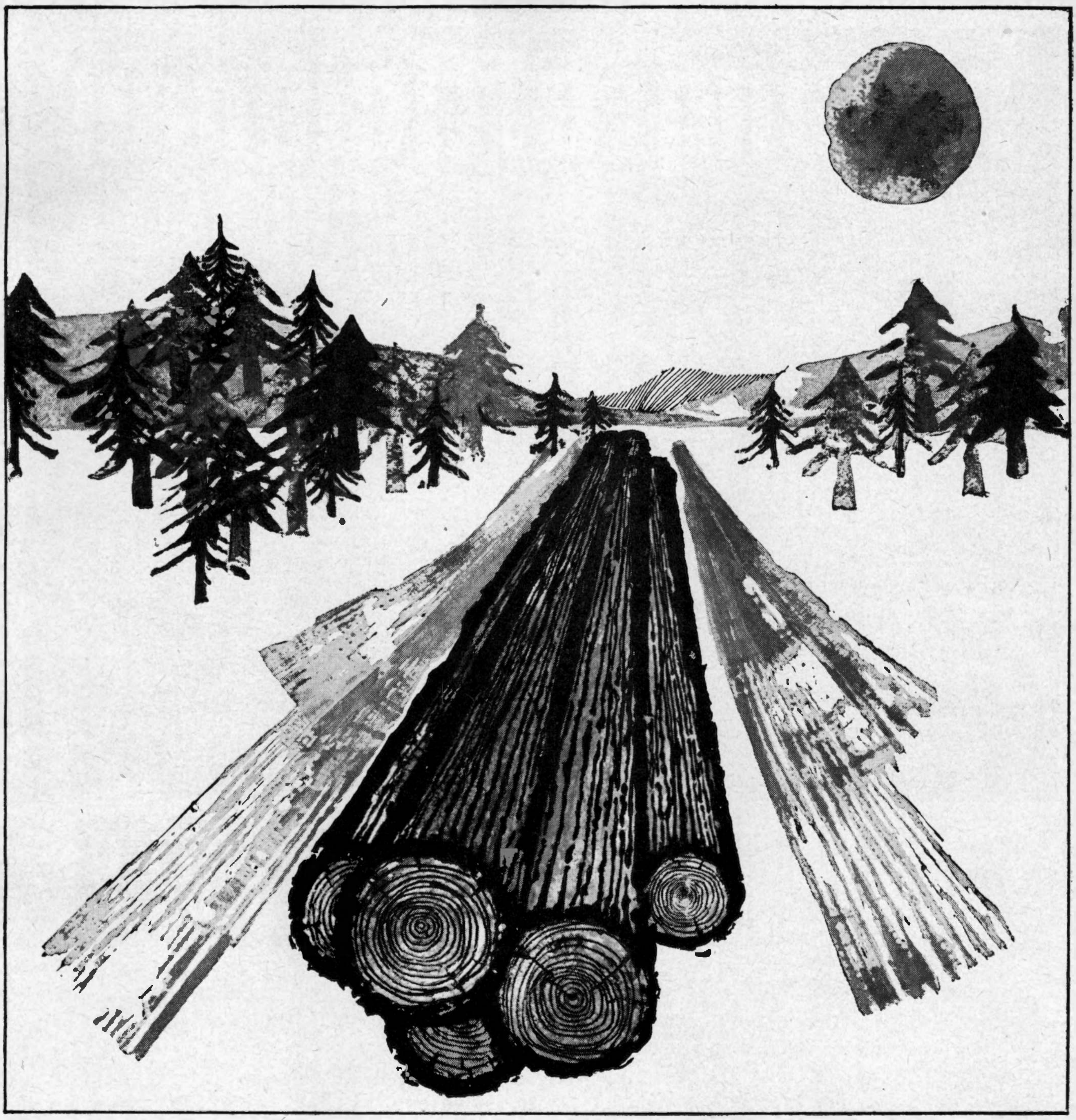

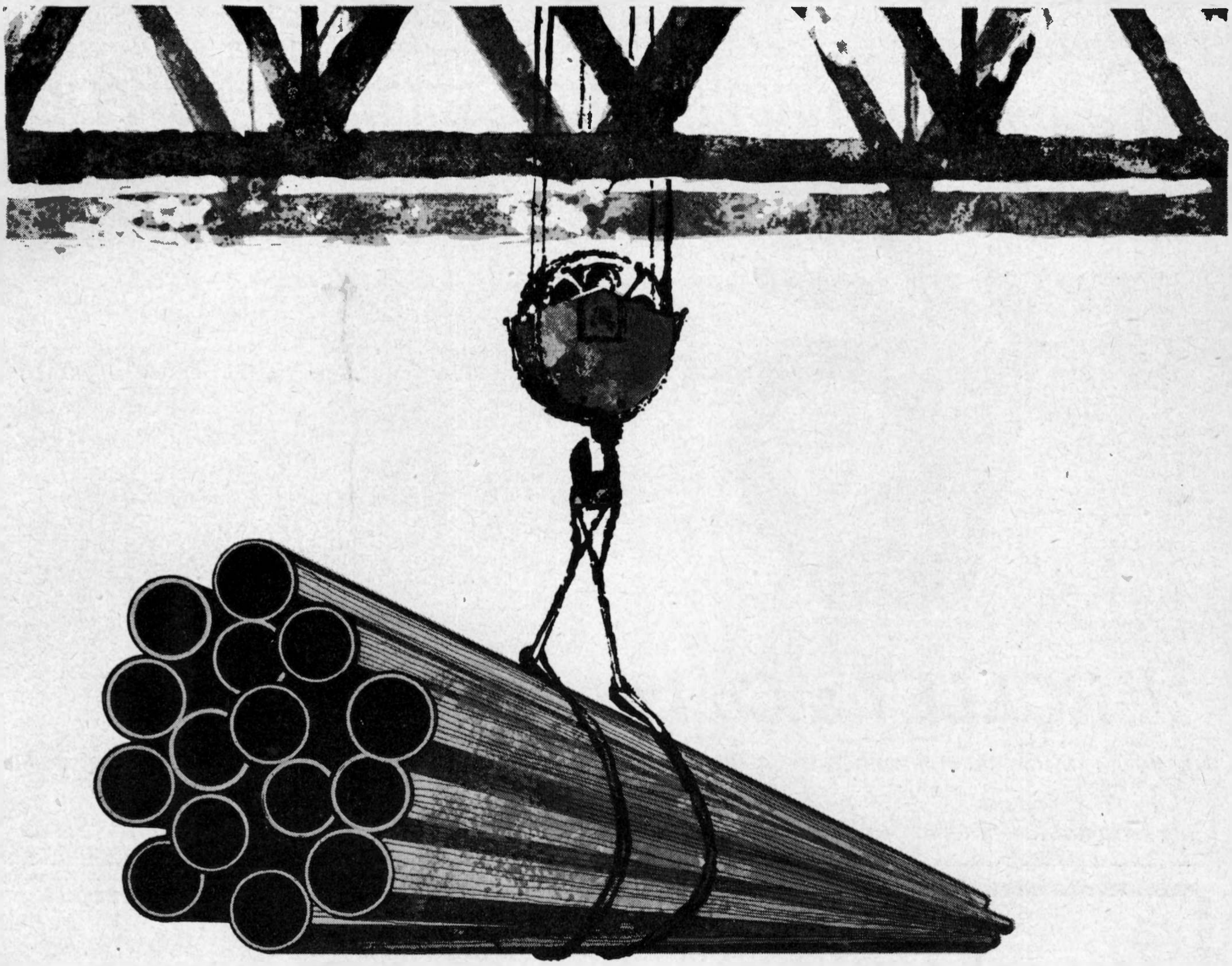

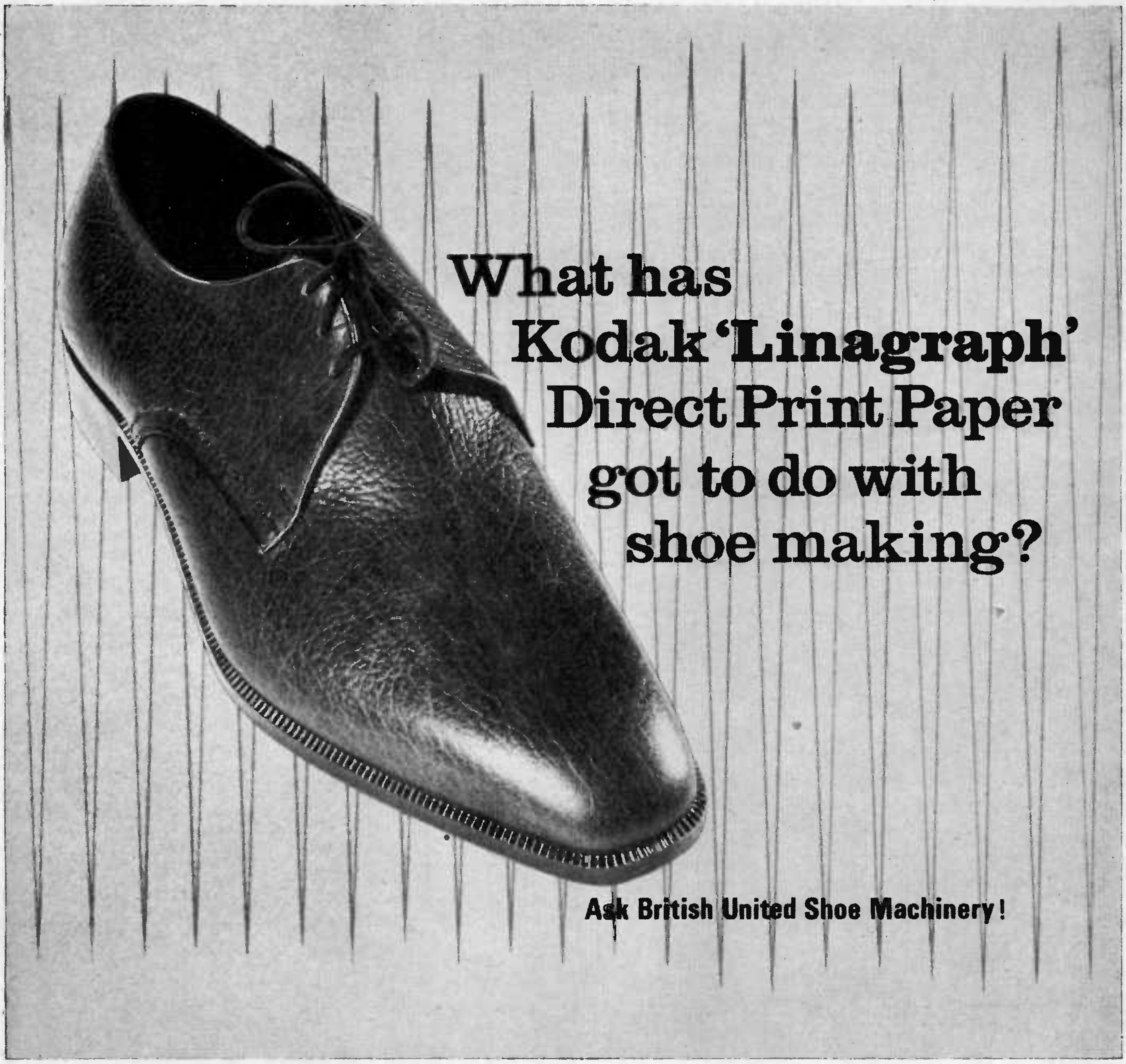
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Ask British United Shoe Machinery!

*For more information circle* **No. 130**

# Grab the new Solartron portable 'scope. Pack your bag. Travel anywhere.

# Your shaver may buck at the mains supply. But never your oscilloscope.

Meet CD 1642.

The new Solartron solid-state go-anywhere oscilloscope with the clearest cleanest trace you ever saw on a portable.

It works anywhere there's power. And anywhere there isn't.

Plug in to 100-130 V a.c. or 200-260 V a.c. at any frequency from 44-440 Hz.

Or in to 12-30 V d.c. A car battery maybe. Or use the built-in rechargeable battery.

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*Details are given in the booklet "Calibration Equipment". May we send you a copy?*

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*For more information circle* **No. 135**


## The DVM's:

### Fairchild 7200:

A full 5-digit meter made for precision measurements in the laboratory or on the production line. It operates to a high standard of accuracy, with 10$\mu$V resolution, and excellent short and long term stability. The principle of operation is based on a new concept where high accuracy and long term stability are achieved by a digital time base memory (Pat. Pend.). The basic unit provides DC voltage measurements, DC ratio measurements, and counting functions to 1MHz. Optional plug-in cards or modules provide AC measurement, frequency measurement, resistance measurement, and other capabilities. Basic unit price is $3500.00.

### Fairchild 7100A:

A full 4-digit meter with extensive capabilities for laboratory and production line measurements. The 7100A measures voltage, resistance and ratio, with AC capabilities optionally available in a plug-in unit. It features guarded construction, 10$\mu$V resolution, 0.01% performance, and excellent stability. Price is $2075.00.

### Fairchild 7000:

A small, half-rack size, portable, medium price 4-digit meter, the 7000 features 0.01% accuracy. The basic unit provides DC voltage measurements, with provisions for adding AC voltage, resistance and current measurements as well as autoranging and BCD output. The front panel of the instrument controls all measurement functions, so that capabilities may be added simply by plugging in a circuit board. Basic unit price is $1150.00.

### Fairchild 7050:

This low-cost, accurate, 3-digit instrument is intended primarily as a replacement for analog-type meters and panel indicators in such applications as production testing, general testing, quality assurance, servicing and the like. Basic features include DC volts and resistance, full scale readout of 1500, input impedance greater than 1000 megohms, floating input, and readout storage (non-blinking display). Price is $299.00

FAIRCHILD
INSTRUMENTATION

Representative: Aveley Electric Ltd. South Ockendon, Essex, Tel: South Ockendon 3444, TWX: 24120 AVEL OCKENDON

European Headquarters: FAIRCHILD INSTRUMENTATION LTD., Grove House, 551 London Road, Isleworth, Middlesex, England, Tel: 560-0838, TELEX: 24693

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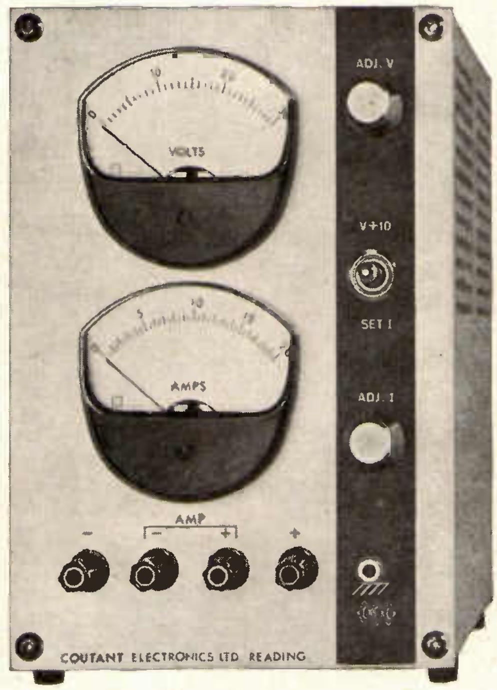


Because it is not a 'run-o'-the-mill' unit. It is Coutant designed, Coutant built and has in-bred Coutant reliability. Its star operational features include: constant current overload protection, precise voltage setting by multi-turn potentiometer, 0.005% voltage regulation, and provision for remote sensing. Each component has been included for its record of high-performance reliability. Here are all the advantages of an advanced all-silicon design in a moderately priced power supply.

| TYPE | VOLTAGE | CURRENT | HEIGHT | WIDTH | LENGTH | WEIGHT | PRICE |
|---|---|---|---|---|---|---|---|
| LA100 | 0-50 | 1 amp | 8½" 216 mm | 5⅝" 143 mm | 9½" 250 mm | 12 lbs 5.5 kg | £76 |
| LA200 | 0-30 | 2 amp | | | | | £79 |
| LA400 | 0-15 | 4 amp | | | | | £81 |
| LB200 | 0-50 | 2 amp | 8½" 216 mm | 5⅝" 143 mm | 14½" 368 mm | 18 lbs 8.2 kg | £105 |
| LB500 | 0-30 | 5 amp | | | | | £110 |
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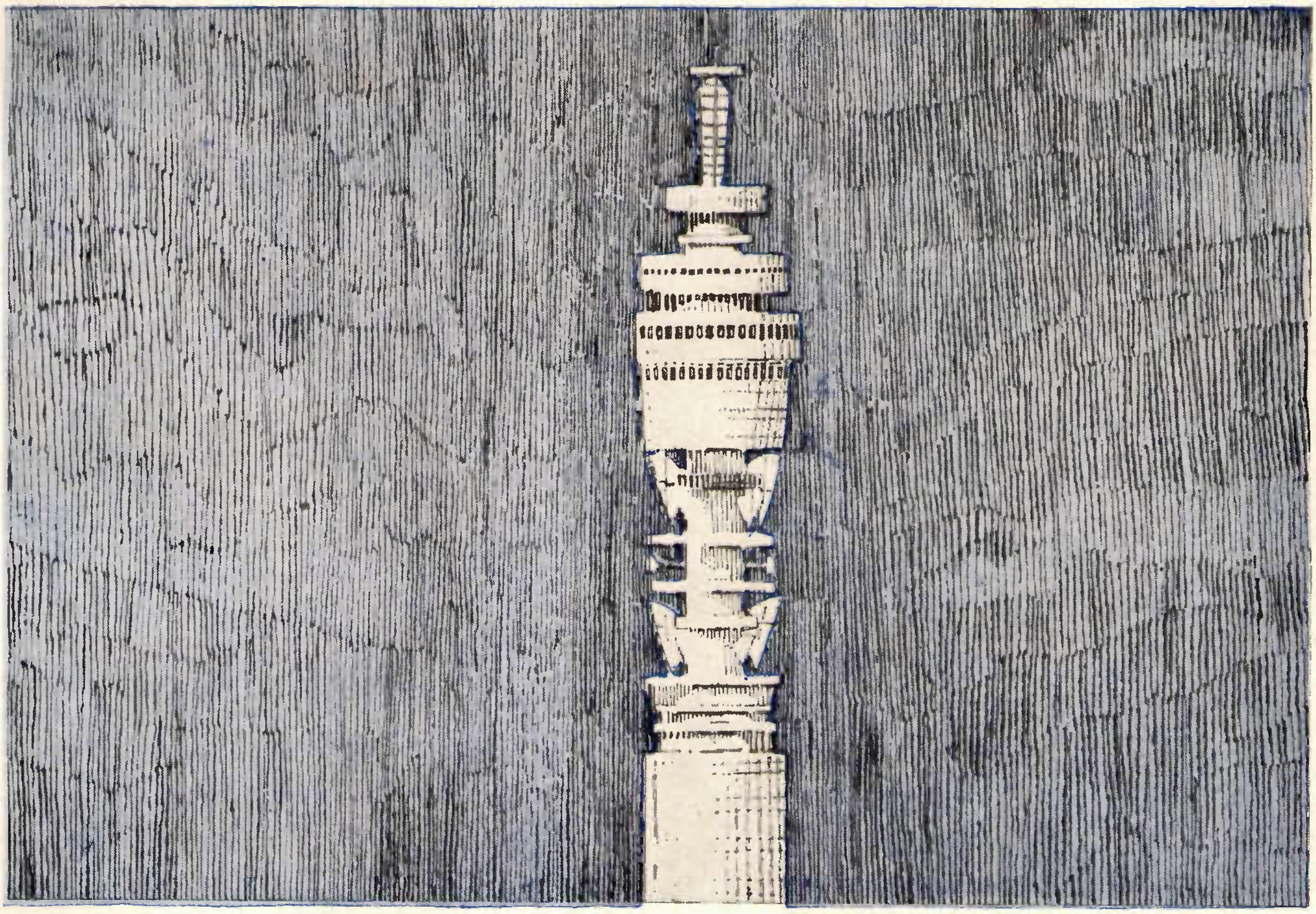
PAM CT2

**coutant electronics ltd**

Electrotech Instrument Division, Trafford Road, Richfield Estate, Reading, Berkshire. Tel: Reading (0734) 55391

*For more information circle* **No. 141**

# Erie take a pride in performance



# in communications

Erie manufacture two Ceramicons of specific interest to the communications field: the Style 390 sub-miniature axial-lead Ceramicon and the Style 395 Transcap Ceramicon, in the identical configuration. Both are identical in quality, yet they have subtle differences in performance and in application.

Designed as a conformal package, these Ceramicons are ideal for high density circuitry.

The single miniature ceramic plate of Erie high permittivity or barrier layer material gives a discrete capacitance range of 5.6pf to 1,200pf and 5,000pf. The unique manner in which the axial leads are attached to these dielectrics produces a robust construction and good high frequency performance. The capacitance ceiling is very high for insulated capacitors of this size and is complemented by optimum voltage ratings over the range —55°C to +125°C.

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In communications, as in instrumentation, aerospace and defence circuitry systems and the whole range of modern electronics . . . Erie take a pride in performance.

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Telephone: Great Yarmouth 4911.
Cables: Resistor Great Yarmouth.
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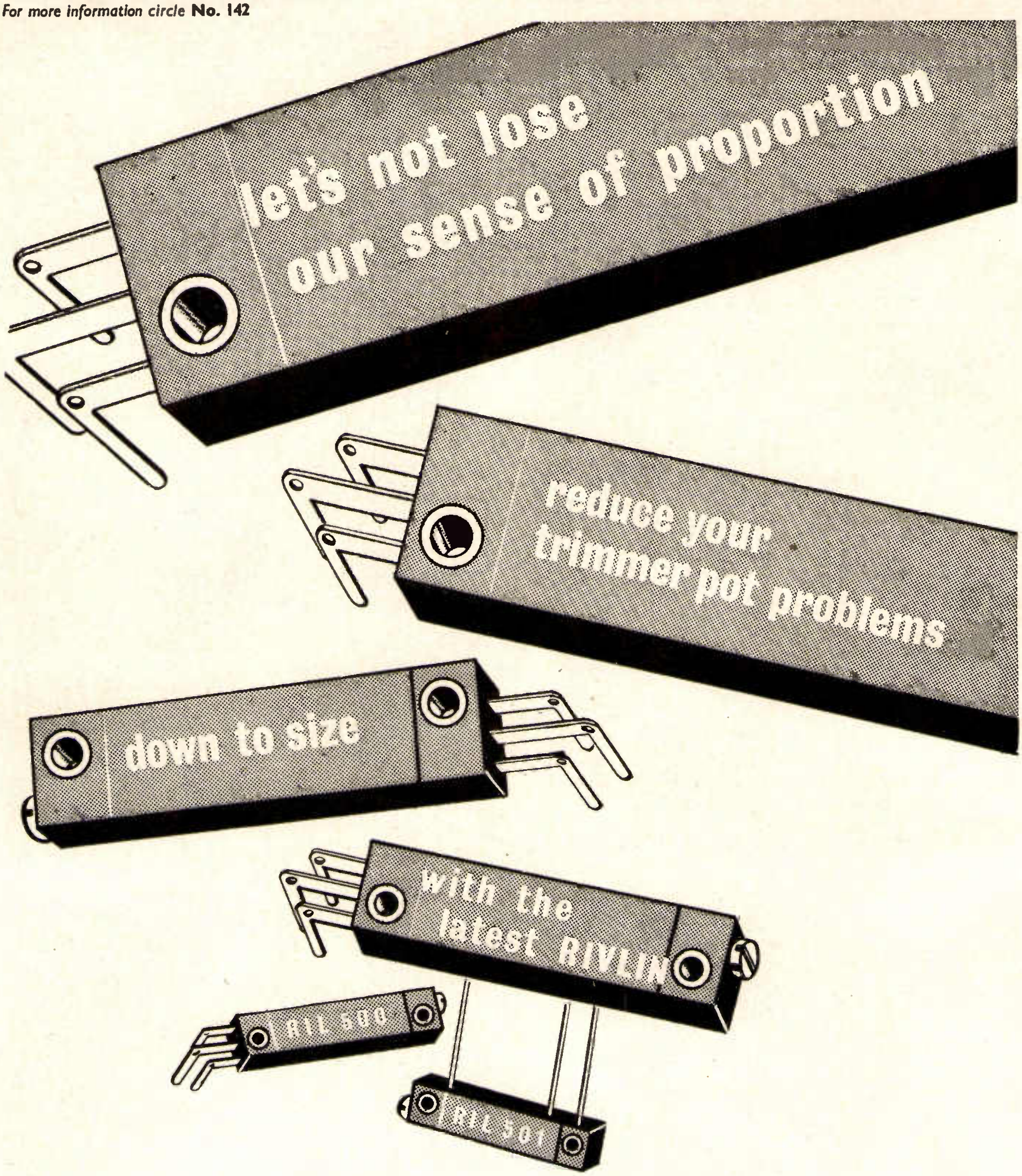
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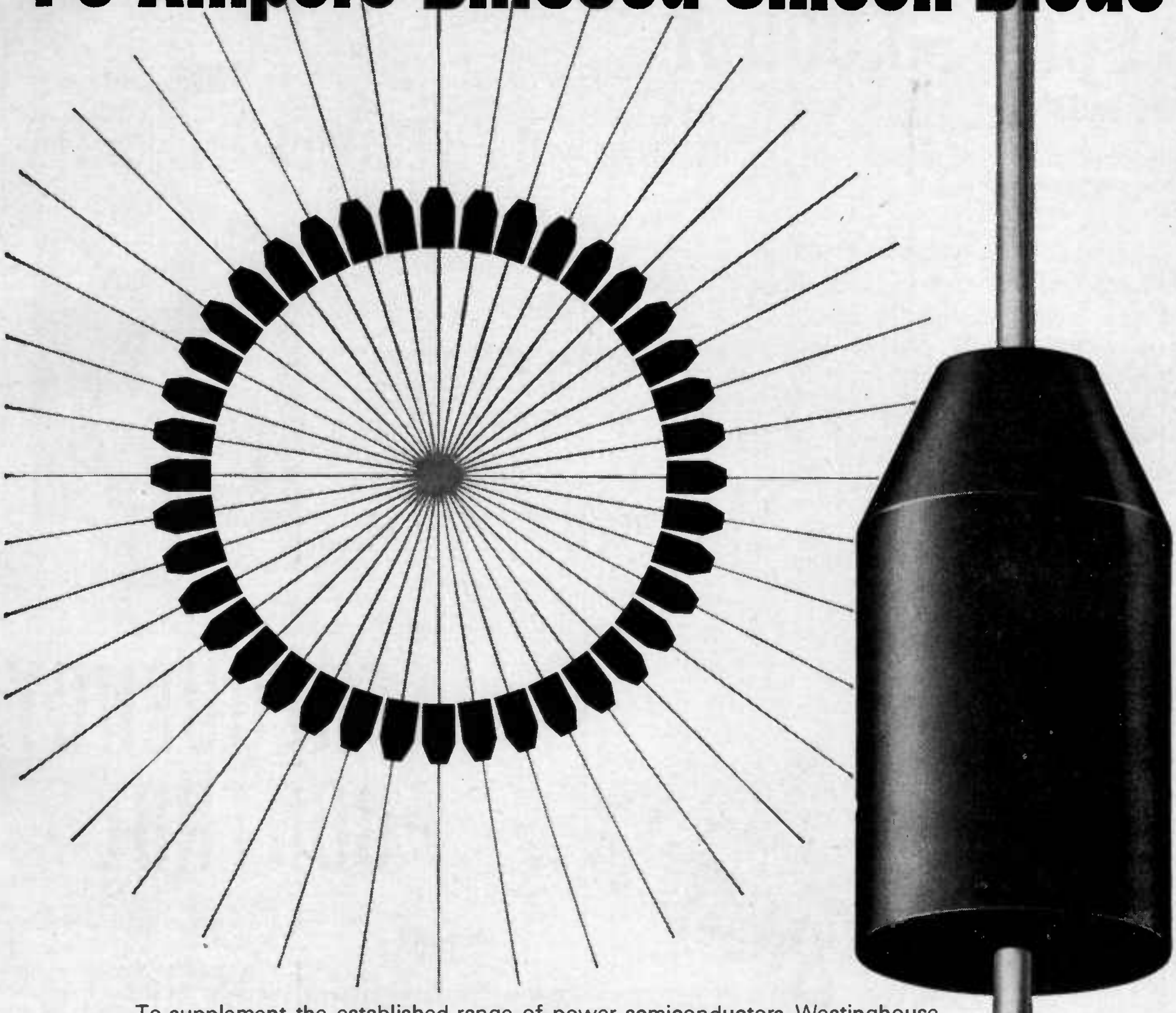

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The M.1. diodes are encapsulated in a rugged epoxy resin case, providing high insulation resistance and full protection from moisture and humidity. The combination of high operating and surge currents with minimum size results in a diode that provides greater flexibility for designers. The M.1. diodes conform to VASCA SO-78 and JEDEC DO-27 outlines.

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SEMICONDUCTOR DIVISION
**WESTINGHOUSE BRAKE AND SIGNAL COMPANY LIMITED**
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LONDON N.1.,
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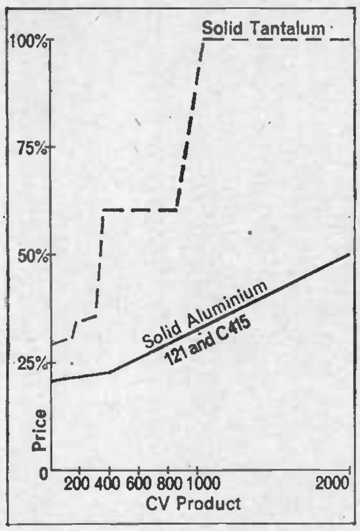
For more information circle No. 156

# Before you specify tantalum for temperature and reliability, take a look at Mullard solid aluminium electrolytics

## *–they're bigger but a fraction of the price*




15 million component-hour life tests on Mullard solid aluminium electrolytics have established a failure rate comparable with solid tantalum types.
Long term stability is as good.
So is low temperature performance.
There are no shelf life problems—reforming is eliminated.
And they are available ex-stock.

BUT THE BIG THING IS THE PRICE
Just look at this graph. Compare costs on a Capacitance x Voltage basis.

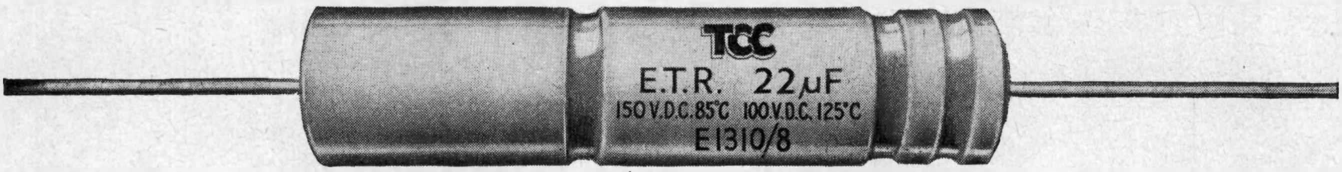
| Working Voltage | | 4V | 6.3V | 10V | 16V | 25V | 40V |
|---|---|---|---|---|---|---|---|
| Value (μF) | C415 | 16 to 100 | 12.5 to 80 | 8 to 50 | 5 to 32 | 3.2 to 20 | 2 to 12.5 |
| | 121 | 180 to 390 | 150 to 330 | 100 to 220 | 56 to 120 | 39 to 82 | 18 to 39 |

By the way, Mullard also makes solid tantalums.
Get the whole story (and a quotation) from Mullard Limited, Industrial Markets Division, Mullard House, Torrington Place, London WC1. Telephone: 01-580 6633

Mullard

IND94

# +125°C




# -55°C

# This is the only aluminium electrolytic available in the UK which operates over this temperature range

**It's smaller than an equivalent tantalum foil type, two thirds the weight and two thirds the price.** The TCC Extended Temperature Range capacitor is the big breakthrough in aluminium foil electrolytic design. Improved foil etching processing plus new formula electrolytes enable an extended temperature range to be achieved with a considerable reduction in size and weight.
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I am particularly interested in capacitors for

TC 14

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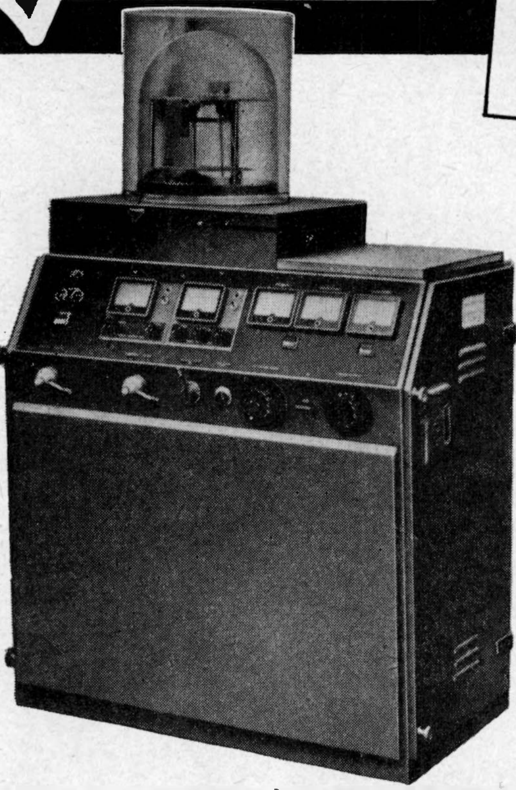
# GeneVac VACUUM COATING UNITS



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- ★ Fully fitted workchamber
- ★ $10^{-4}$ torr in 4 minutes with 4" system
- ★ 12v. 40A. LT Evaporation supply
- ★ 3000v. 100mV. Ion clean up supply

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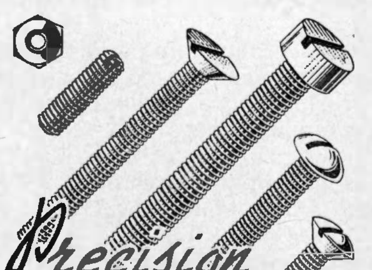

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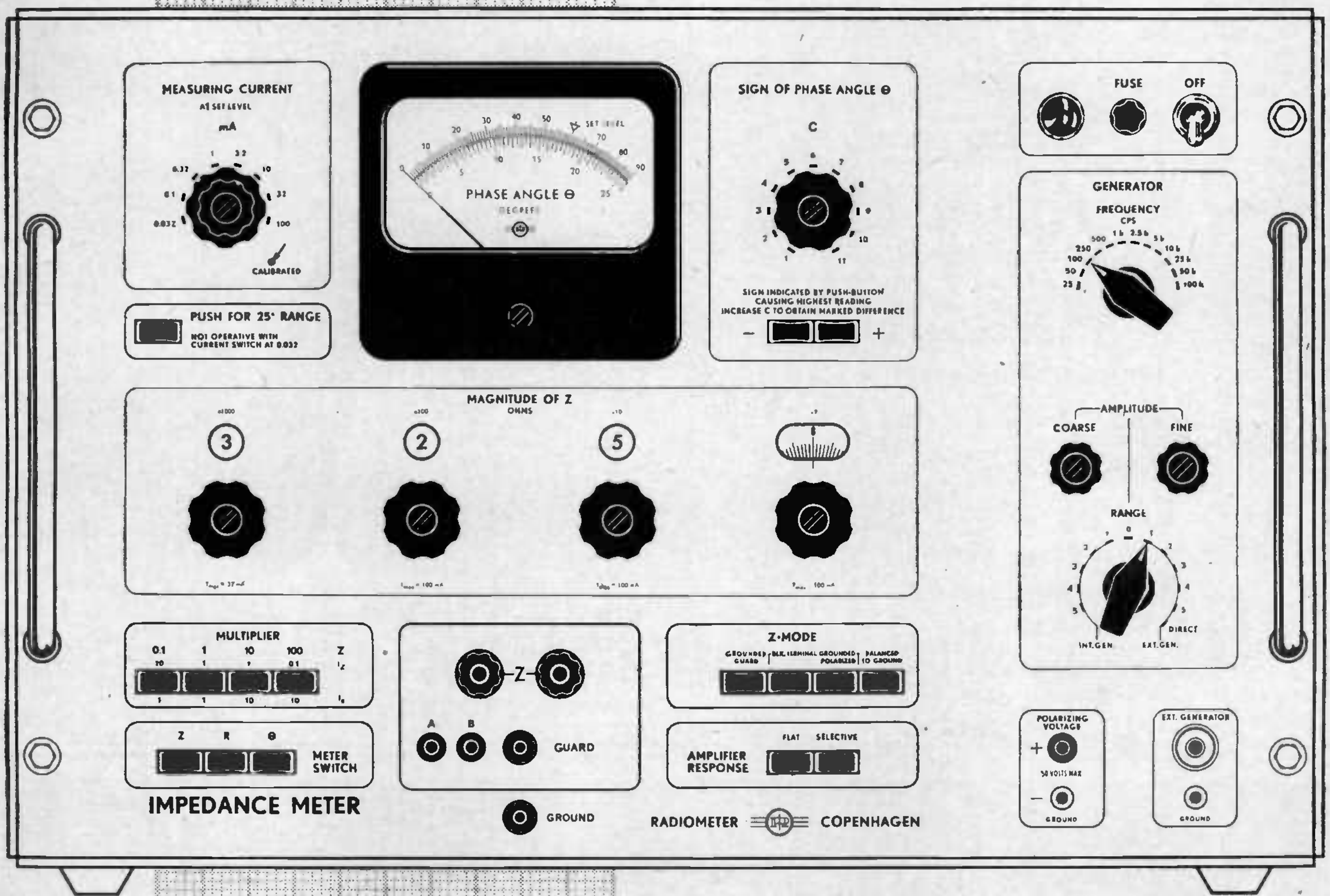
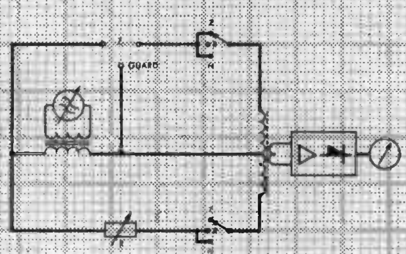
# NEW APPROACH TO IMPEDANCE MEASUREMENT

## TYPE GB11




The type GB11 Impedance Meter is a fully transistorized impedance measuring bridge with the following features:

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▶ Measurements on floating, grounded, or balanced-to-ground impedances.

▶ Measurements of dc polarized impedances. ▶ Wide frequency range.

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Accuracy: Magnitude 1 % ±0.1 Ω, phase angle ±1° (below 10° ±0.5°).

Test Current: Adjustable from 3.2 μA to 1 A.

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TK0967E

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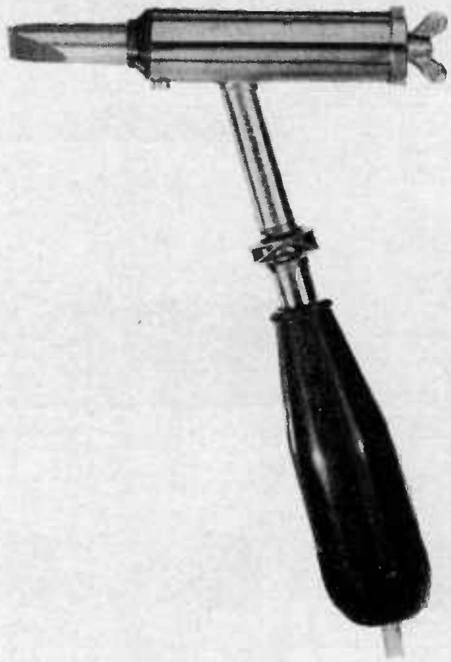





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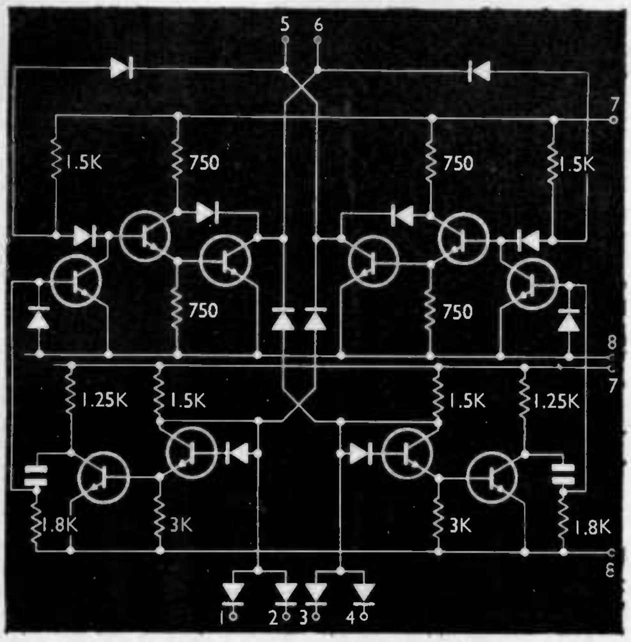
# MARCONI MICRONOR II

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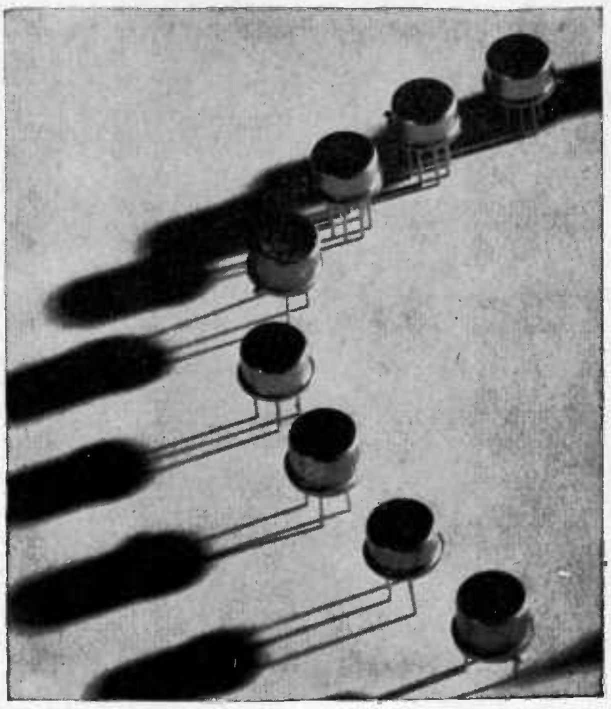

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W 8632

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### THE ROYAL FREE HOSPITAL

**ELECTRONIC ENGINEER**

Electronic Engineer required by a London Teaching Hospital. The successful candidate will be part of a team of electronic engineers but will have special responsibility for the maintenance and modification of electronic diagnostic apparatus including electro-physiological and data processing equipment in the Department of Psychological Medicine at the Royal Free Hospital, Lawn Road, Hampstead N.W.3. Experience with either biological or pulse techniques would be an advantage but is not essential. A good salary will be paid. Detailed applications stating age, qualifications and experience to the Administrator, The Royal Free Hospital, Gray's Inn Road, London W.C.1.
W 8625

### UNIVERSITY OF BIRMINGHAM

**DEPARTMENT OF CIVIL ENGINEERING**

**POSTGRADUATE RESEARCH IN WATER RESOURCES TECHNOLOGY**

Applications are invited for the posts of 1.) Research Assistant in Electronics, 2.) Research Assistant in Flood Movement. The first position is concerned with the development and application of electrical analogues to the solution of unsteady state problems. The second position will involve the analysis of flood movements in river systems and the effectiveness of control by regulation. Applicants for both appointments should possess, or anticipate possessing this year, an honours degree in Engineering or Applied Science or its equivalent. It will be possible, in appropriate circumstances, for the persons appointed to submit work for a higher degree of this University. Salary in the range £840–£900 p.a. for a period of two years to commence as soon as possible.
Applications (3 copies) naming three referees, should be sent to the Assistant Registrar (S), P.O. Box 363, University of Birmingham, Birmingham, 15.
W 8623

### UNIVERSITY OF EDINBURGH

**LABORATORY SUPERINTENDENT**

Applications are invited for the post of Superintendent of the Laboratories of the Department of Natural Philosophy (Physics). Candidates for the post should preferably be graduates in physics or in some branch of engineering, or holders of a Higher National Certificate, or equivalent. Experience in mechanical workshop practice is desirable; a talent for leadership is essential.
Duties will include assistance to members of the teaching staff in the design of apparatus for research, general supervision of the technical staff, oversight of the mechanical workshop and stores and of the maintenance of apparatus in the teaching laboratories, and the organisation and supervision of a departmental technician training programme.
Salary will be on a scale identical to that for University Lecturers (£1,470–£2,630 per annum), with initial placing according to qualifications and experience.
Applications, by letter (three copies), giving the names of two referees, should be lodged with The Secretary to the University, University of Edinburgh, Old College, South Bridge, Edinburgh, not later than 26th April, 1967.
W 8594

### UNITED NEWCASTLE UPON TYNE HOSPITALS

**MEDICAL PHYSICS UNIT**

A Technician with knowledge of electronics is required in the Department of Medical Physics at the Royal Victoria Infirmary, Newcastle. The successful candidate will be required to assist in the design, construction and maintenance of electronic apparatus, research and clinical.
Candidates should have at least O.N.C. in electrical engineering or applied physics, or an equivalent qualification.
Salary according to qualifications and experience.
(a) £711 per annum rising to £1,000 by annual increments, or
(b) £980 per annum rising to £1,300 by annual increments.
As a result of the Government's Prices and Incomes Policy these scales will not be implemented immediately but will be paid retrospectively to the date of appointment. In the interim the salary will be £696 per annum in respect of (a), or £868 per annum in respect of (b).
Applications stating age, qualifications, experience and the names and addresses of two referees should be sent to the House Governor and Secretary, Royal Victoria Infirmary, Newcastle upon Tyne 1, within 14 days of the appearance of this notice.
W 8622

### NORTHERN POLYTECHNIC

**HOLLOWAY, LONDON, N.7**

**CHIEF TECHNICIAN**

Applications are invited for the permanent and pensionable post of Chief Technician with experience in the building and servicing of electronic equipment, to take charge, in conjunction with the Organising Technician, of electronic instrumentation used in the teaching and research work of the Department of Physics. Salary scale: £1,260 × £35—£1,435.
Write, stating age, qualifications, experience, etc. to the Secretary.
W 8628

### IMPERIAL COLLEGE

**TECHNICIAN/SENIOR TECHNICIAN**

Technician/Senior Technician required for operation and maintenance of Cary 81 helium-neon laser Raman Spectrometer shortly to be installed as part of the University of London Intercollegiate Research Service. Interest in Raman Spectroscopy desirable but experience not essential if basic knowledge of electronics is adequate. Salary in University Scale (Technician) £713–£1,028 or (Senior Technician) £962–£1,240; superannuation; good holidays. Apply in writing with full details to Departmental Superintendent, Chemistry Department, Imperial College, London, S.W.7.
W 8624

## CLASSIFIEDS

*continued on pages 126 & 127*

## OFFICIAL APPOINTMENTS

### THE UNIVERSITY OF SUSSEX

**SCHOOL OF MATHEMATICAL AND PHYSICAL SCIENCES**

**TECHNICIAN AND SENIOR TECHNICIAN IN ELECTRONICS**

Applications are invited for posts of TECHNICIAN and SENIOR TECHNICIAN to join a group which carries out the design and construction of a wide range of electronic equipment for use in experimental physics research.

Applicants for the senior post should possess or expect to obtain shortly H.N.C. or equivalent. Opportunities are available for day release and salaries will be offered according to age and experience within the approved scales:—
Technician: £683–£968 per annum.
Senior Technician: £942–£1,180 per annum.
Further details and application forms may be obtained from the Laboratory Superintendent, School of Mathematical and Physical Sciences, University of Sussex, Falmer, Brighton, Sussex. (Please quote Ref. 425/1). W 8619

## SITUATIONS VACANT

**TEST AND SERVICE ENGINEER REQUIRED** by firm producing machine tool electronic control systems. Experienced persons with initiative able to work without supervision, who prefer a go-ahead, small, but independent organisation, should apply, stating salary required, to Box No. W 8620, c/o ELECTRONIC ENGINEERING.

**YOUNG AMBITIOUS ELECTRONICS** sales engineer, with good qualifications, experience and sound knowledge of semi-conductor applications, required to fill post with rapidly expanding semi-conductor distributors. Extremely good prospects, and rapidly increasing responsibility and salary for right man.
Applications with details in writing to: Box No. W 8618, c/o ELECTRONIC ENGINEERING.

**OVERSEAS**

We have a steady flow of enquiries for qualified men.
If you are interested, write to:
Major ENgagements Ltd., Sheepy Parva, Atherstone, Warwickshire.
W 8606

ELECTRONIC CONTROL DEVICES

Electronic Control Devices

We wish to recruit a young *Electronic Engineer* with experience of, and interest in, the application of *Electronic Control Devices*—such as Thyristors—to heavy industrial systems, equipment and processes. We currently have projects requiring this expertise under way in the steel, glass and chemical industries.

Salary on a Band within the overall range of £980—£3,025.

**Please write for further information to Mr. D. C. Page, Head of Personnel Services, quoting reference number E.E./66**

**ELECTRICITY COUNCIL RESEARCH CENTRE,**
**Capenhurst, near Chester**

## EDUCATIONAL

**UNIVERSITY OF SOUTHAMPTON**

**QUANTUM ELECTRONICS**

Applications are invited for the one-year post-graduate course in the Department of Electronics commencing October 1967. The course provides a thorough training in the principles, design and application of a wide range of quantum electronic devices, in particular masers and lasers. Lectures are given by members of the University's large quantum electronics research group and by scientists from industrial and governmental laboratories. An essential part of the course is a research project on an aspect of the work of the research group. A Diploma is awarded on the results of an examination in June and graduates (and with special approval those with other qualifications) may continue with the project for a further three months and submit a dissertation for an M.Sc. The course is recognised by the Science Research Council for the award of Advanced Course Studentships.
Full time research posts are also available in the Quantum Electronics Research Group.
Further details of the course and the research positions may be obtained from
The Academic Registrar, The University, Southampton. W 8601

## BUSINESS OPPORTUNITIES

**ALARM EVENT RECORDING EQUIPMENT**

The Proprietors of British Patent No. 971024 entitled 'Improvements in or relating to Recording Systems' wish to enter into negotiation for licensing manufacture or other arrangements. This system should be of particular interest to manufacturers of alarm annunciator systems. Other data handling equipment is also available. Box No. W 8633, c/o ELECTRONIC ENGINEERING.

## FOR SALE

**FOR SALE**—1 Ediswan 8 Channel Electro-Encephalograph Mk. II Machine (purchased 1951) with spare Valves, Motors, and Pens plus paper. Offers should be submitted within 14 days to the Supplies Officer, Taunton H.M.C., Taunton & Somerset Hospital, Musgrove Park Branch, Taunton, Somerset. W 8621

## CAPACITY AVAILABLE

**AIRTRONICS LTD.**, for coil winding, assembly and wiring of electronic equipment, transistorised sub-units. Sheet Metal Work. 3A, Walerand Road, London, S.E.13. Telephone: Lee Green 1706. W 370

BOX NUMBER REPLIES SHOULD BE ADDRESSED AS FOLLOWS:-

**BOX NO......,**

**ELECTRONIC ENGINEERING**

**28 ESSEX STREET, STRAND, LONDON, W.C.2**

# DEVELOPMENT ENGINEERS

G.E.C. Hammersmith invite applications from men over 22 years of age for general design, development, and pre-production work on Thermionic Vacuum Tubes. Their work would be under the guidance of a Senior Engineer and would include detailed control of assembly, processing, testing, assessment and rectification of faults and the use of microwave measurements and techniques to improve design.

The successful candidates will have U.K. Degrees in Electrical Engineering or Physics or will have had sound practical experience in valve manufacture with an H.N.C. as a minimum qualification.

Attractive starting salaries and excellent fringe benefits will be offered.

*Please write, giving brief details of experience, to: Mr. M. B. Stevenson, Personnel Officer, M–O Valve Co. Ltd., Brook Green Works, Hammersmith, W.6.*

G.E.C. **HAMMERSMITH**

W 8629

LEB **LONDON ELECTRICITY BOARD**

# ENGINEER (DEVELOPMENT)

Applications are invited for the above position in the Development Branch of the Chief Engineer's Department at Spa Road, Neckinger, S.E.1, for a young Engineer interested in the general fields of measurement and instrumentation.

Duties include the development of analogue and digital systems and general laboratory measurement work. It would be an advantage if applicants were experienced in the fields of data handling or telecontrol.

Commencing salary in scale £1,310 to £1,635 per annum inclusive.

**Application form obtainable from the Personnel Officer at 46 New Broad Street, E.C.2, to be returned completed within 14 days of the publication date of this notice. Please quote reference number PER/V/6326/AR**

W 8631

# senior telemetry systems engineer-natural gas

**The Gas Industry** is entering an exciting period of expansion. Within recent years its scientists and technologists have discovered new ways of making gas from oil and have carried through the project which brings liquefied natural gas to Britain from the Sahara Desert. Beneath the North Sea there are rich reserves of natural gas which will not only further transform the Gas Industry but will be of great importance to the nation's economy.

**An Engineering Research Station** has been set up by the Gas Council to solve the problems which arise and to establish plans for the future. The research team recruited for this new station will have special responsibility for distribution engineering and materials. Modern laboratories are under construction and will be ready for occupation later this year when advanced experimental and computing facilities will become available.

**A SENIOR TELEMETRY SYSTEMS ENGINEER,** ***with at least five years experience, familiar with current practice on microwave, radio and G.P.O. communication lines is required to lead research and development projects.***

***Immediate attention is to be given to the preparation of equipment and system standards together with design codes of practice for use in the establishment of a National Telemetry Control Scheme.***

***An interest in the assessment of general purpose electronic equipment standards is also desirable.***

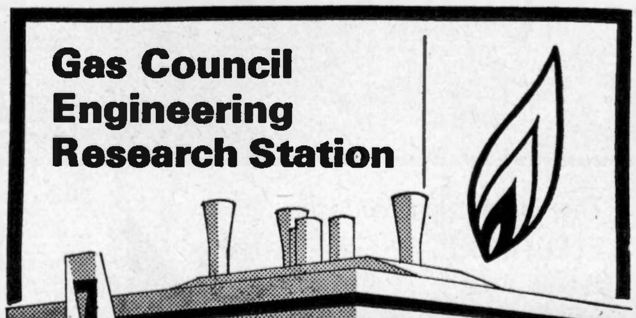
**QUALIFICATIONS: Must include a good degree.**

**STARTING SALARY:** Up to £2,370 per annum. There are excellent prospects of promotion.

**LOCATION:** Newcastle upon Tyne.

**APPLICATION FORMS FROM:** Gas Council Engineering Research Station, Benton Park Road, Newcastle upon Tyne, 7.

Please quote reference STE/DT.



# INDEX TO ADVERTISERS

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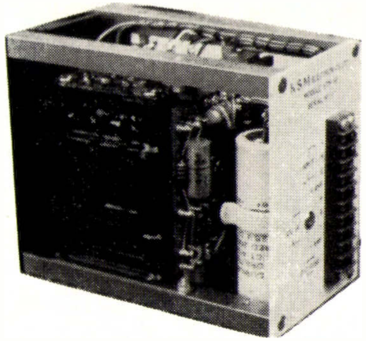
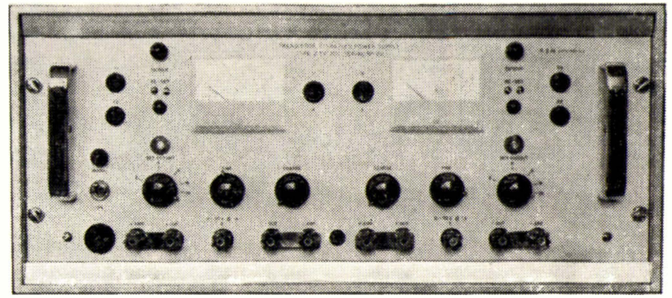
# KSM DC REGULATED POWER SUPPLIES



## 0-10KV 0-100AMPS

Short Circuit & overload protection.
Constant voltage or constant current output.
Remote sensing terminals.
All models supplied as modules, bench units, or 19" rack mounting.
Mains supply, 90-120V, 200-240V, or 3 p.h.
Complete protection against short circuits & overloads.



### KTV RANGE (SILICON) 0-100V 0-100A

Regulation .01% Line .1% Load Ripple 1 MV

| MODEL | D.C. OUTPUT (FLOATING) VOLTS | AMPS |
|---|---|---|
| KTV35 | 0–30 | 0·5A |
| KTV301 | ,, | 1A |
| KTV302 | ,, | 2A |
| KTV303 | ,, | 3A |
| KTV305 | ,, | 5A |
| KTV310 | ,, | 10A |
| KTV320 | ,, | 20A |
| KTV330 | ,, | 30A |
| KTV340 | ,, | 40A |
| KTV350 | ,, | 50A |
| KTV3100 | ,, | 100A |
| KTV55 | 0–50 | ·5A |
| KTV501 | ,, | 1A |
| KTV502 | ,, | 2A |
| KTV503 | ,, | 3A |
| KTV505 | ,, | 5A |
| KTV510 | ,, | 10A |
| KTV520 | ,, | 20A |
| KTV530 | ,, | 30A |
| KTV1001 | 0–100 | 1A |
| KTV1002 | ,, | 2A |
| KTV1003 | ,, | 3A |
| KTV1005 | ,, | 5A |
| KTV1010 | ,, | 10A |
| 2TV35 | 2 × 0–30 | 2 × 0·5A |
| 2TV301 | ,, | 2 × 1A |
| 2TV302 | ,, | 2 × 2A |
| 2TV303 | ,, | 2 × 3A |
| 2TV305 | ,, | 2 × 5A |
| 2TV310 | ,, | 2 × 10A |
| 2TV55 | 2 × 0–50 | 2 × 0·5A |
| 2TV501 | ,, | 2 × 1A |
| 2TV502 | ,, | 2 × 2A |
| 2TV503 | ,, | 2 × 3A |
| 2TV505 | ,, | 2 × 5A |

### V and VV RANGE 0-500V 0-2A

Regulation .01% Line .1% Load Ripple 1 MV

| MODEL | | D.C.OUTPUT (FLOATING) VOLTAGE | CURRENT |
|---|---|---|---|
| V | 205 | 100–200 | 50mA |
| V | 210 | ,, | 100mA |
| V | 225 | ,, | 250mA |
| V | 250 | ,, | 500mA |
| V | 2100 | ,, | 1A |
| V | 2200 | ,, | 2A |
| V | 305 | 200–300 | 50mA |
| V | 310 | ,, | 100mA |
| V | 325 | ,, | 250mA |
| V | 350 | ,, | 500mA |
| V | 3100 | ,, | 1A |
| V | 3200 | ,, | 2A |
| V | 405 | 300–400 | 50mA |
| V | 410 | ,, | 100mA |
| V | 425 | ,, | 250mA |
| V | 450 | ,, | 500mA |
| V | 4100 | ,, | 1A |
| V | 4200 | ,, | 2A |
| V | 505 | 400–500 | 50mA |
| V | 510 | ,, | 100mA |
| V | 525 | ,, | 250mA |
| V | 550 | ,, | 500mA |
| VV | 525 | 0–500 | 250mA |
| VV | 550 | 0–500 | 500mA |
| VV | 5100 | 0–500 | 1A |
| VV | 5200 | 0–500 | 2A |

### SC RANGE (SILICON S.C.R.) 0-500V 0-50A

Regulation .2% Line .2% Load Ripple 1%

| MODEL | | D.C. OUTPUT (FLOATING) VOLTAGE | CURRENT |
|---|---|---|---|
| SC | 110 | 0–100 | 10A |
| SC | 120 | ,, | 20A |
| SC | 130 | ,, | 30A |
| SC | 140 | ,, | 40A |
| SC | 150 | ,, | 50A |
| SC | 210 | 0–200 | 10A |
| SC | 220 | ,, | 20A |
| SC | 230 | ,, | 30A |
| SC | 240 | ,, | 40A |
| SC | 250 | ,, | 50A |
| SC | 310 | 0–300 | 10A |
| SC | 320 | ,, | 20A |
| SC | 330 | ,, | 30A |
| SC | 340 | ,, | 40A |
| SC | 350 | ,, | 50A |
| SC | 410 | 0–400 | 10A |
| SC | 420 | ,, | 20A |
| SC | 430 | ,, | 30A |
| SC | 440 | ,, | 40A |
| SC | 450 | ,, | 50A |
| SC | 510 | 0–500 | 10A |
| SC | 520 | ,, | 20A |
| SC | 530 | ,, | 30A |
| SC | 540 | ,, | 40A |
| SC | 550 | ,, | 50A |

NEW RANGE

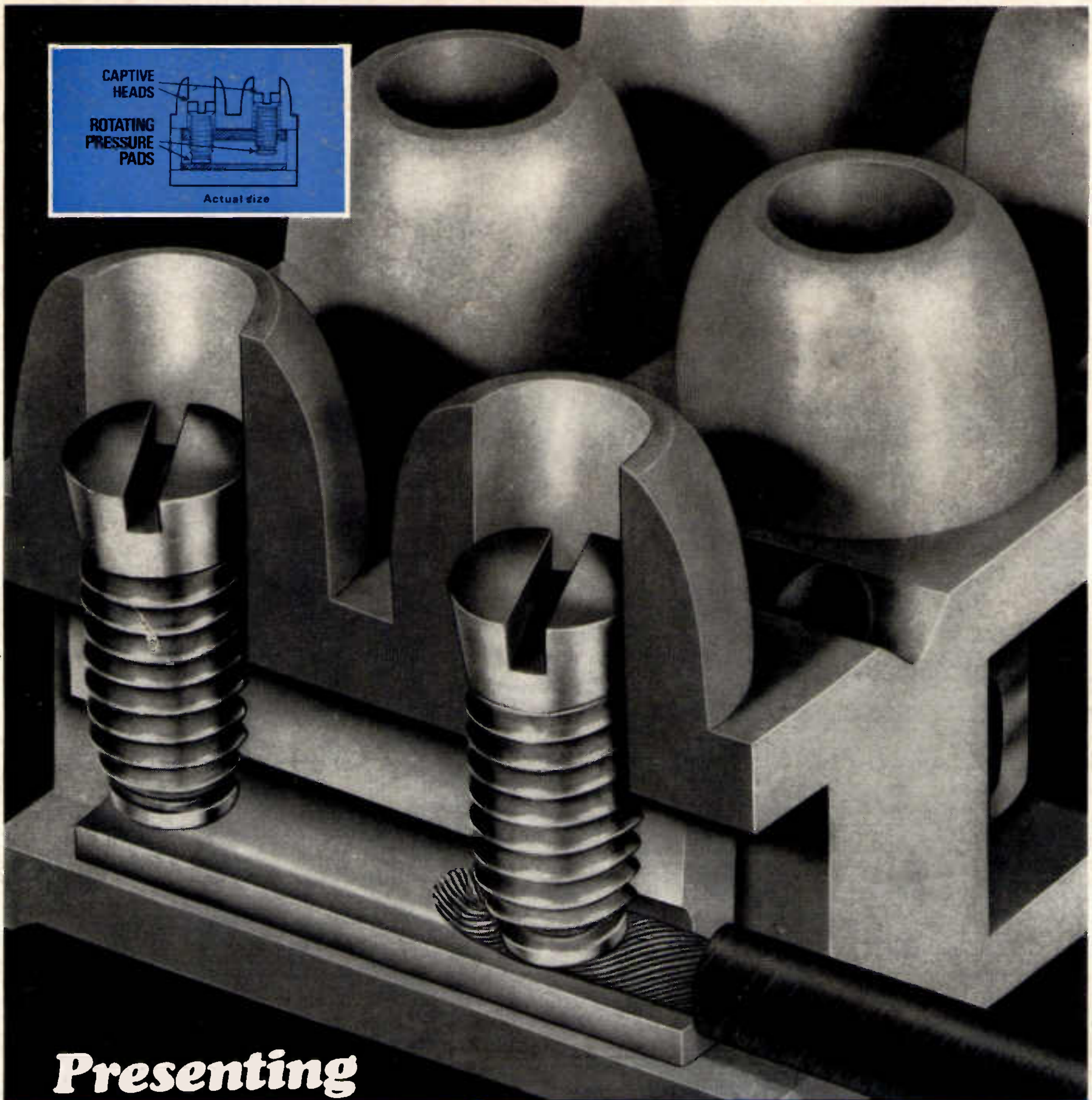
### HV and HVV RANGE 0-10KV 0-500MA

Regulation .05% Line .1% Load Ripple 2 MV

| MODEL | | D.C. OUTPUT (FLOATING) VOLTAGE | CURRENT |
|---|---|---|---|
| HV | 1005 | 1,000 | 5mA |
| HV | 1505 | 1,500 | 5mA |
| HV | 2005 | 2,000 | 5mA |
| HV | 1020 | 1,000 | 20mA |
| HV | 1520 | 1,500 | 20mA |
| HV | 2020 | 2,000 | 20mA |
| HVV | 1100 | 0–1,000 | 100mA |
| HVV | 1200 | 0–1,000 | 200mA |
| HVV | 1500 | 0–1,000 | 500mA |
| HVV | 2100 | 0–2,000 | 100mA |
| HVV | 2200 | 0–2,000 | 200mA |
| HVV | 2500 | 0–2,000 | 500mA |
| HVV | 3100 | 300–3,000 | 100mA |
| HVV | 3200 | 300–3,000 | 200mA |
| HVV | 3500 | 300–3,000 | 500mA |
| HVV | 5002 | 500–5,000 | 2mA |
| HVV | 10001 | 3,000–10,000 | 1mA |
| HVV | 10020 | 3,000–10,000 | 20mA |

# KSM ELECTRONICS LIMITED

139-149 Fonthill Road, London, N.4. Tel: ARChway 6160




***Presenting***
***the only screw of***
***it's type***
***in captivity***

**L1639/B** incorporates a unique pressure pad which is free to rotate and is capable of clamping the finest wires without damage. Moulded in flexible P.V.C. the block can be mounted in any position. Terminal screws are captive and vibration proof.

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