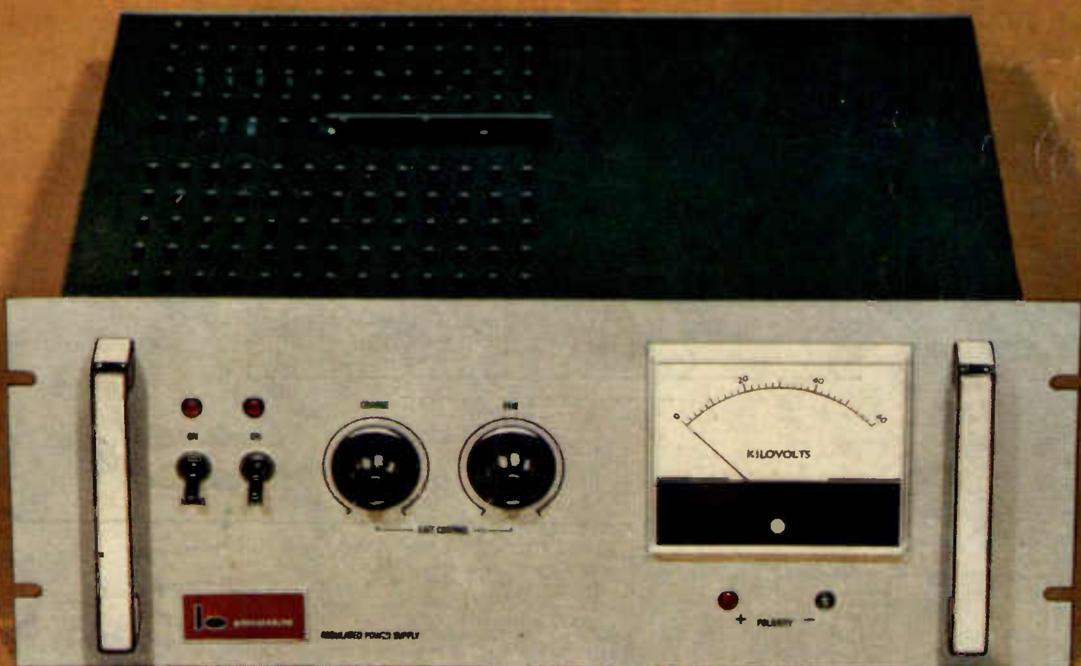
Electronic Engineering

DECEMBER 1966



THREE SHILLINGS & SIXPENCE

Vol. 38 No. 466

DECEMBER 1966

Electronic Engineering

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

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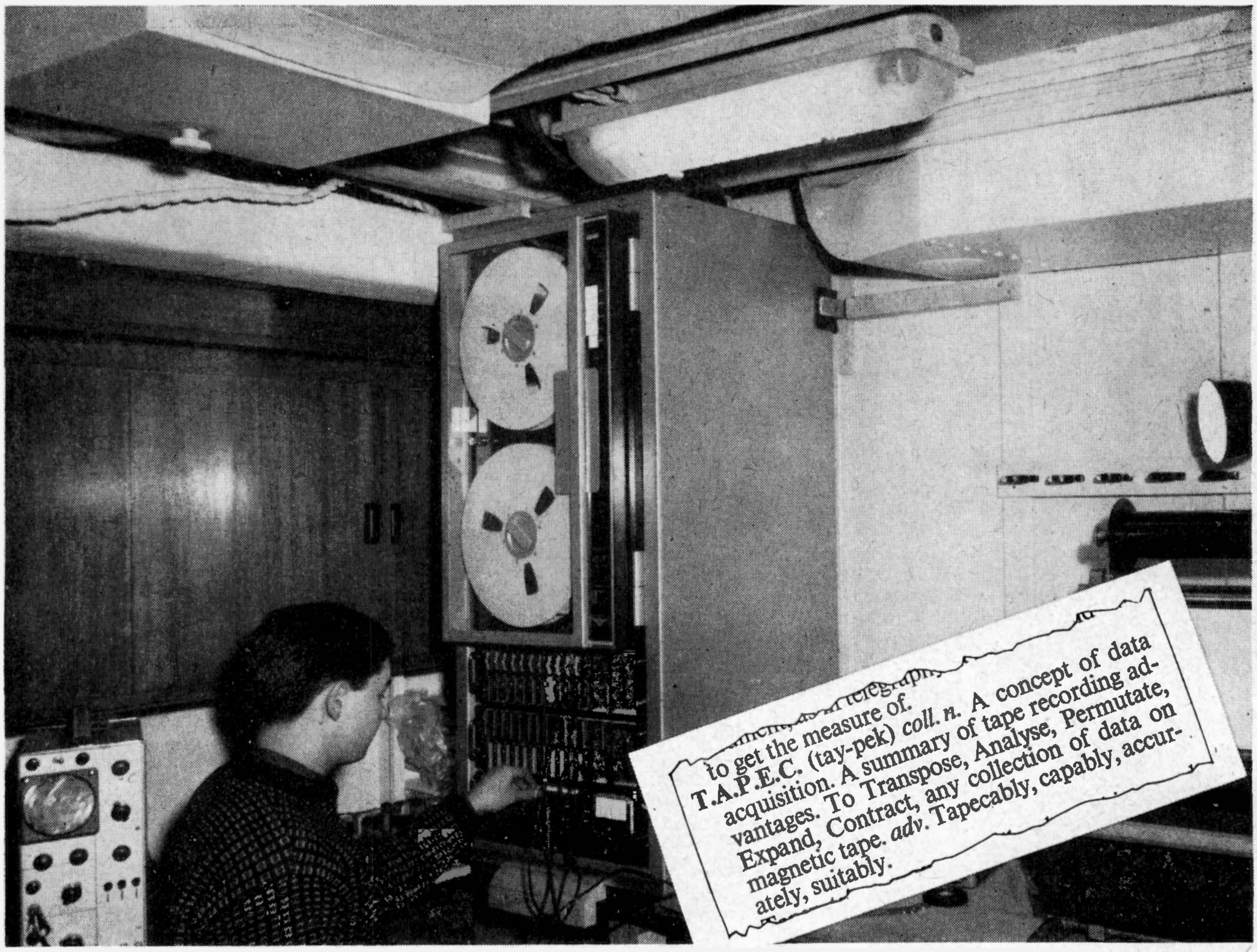

The T.A.P.E.C. technique..in Oceanography

The T.A.P.E.C. technique – data recording on magnetic tape with its unique advantages of flexibility and permanence – is seen here in use in R.R.S. Discovery for research into the nature of the ocean bed. In a typical experiment of this kind a tape recorder records the shock wave signals, caused by explosions on the ocean bed, detected by a system of transducer, geophones and hydrophones. Playback on a Thermionic Instrumentation recorder at a faster speed is used in conjunction with a chart recorder to give a continuous picture of the physical properties of the ocean bed.

Magnetic tape provides a permanent, infinitely repeatable record that can be Transposed, Analysed, Permutated, Expanded and Contracted. That's why we've coined the word T.A.P.E.C. to remind you of the advantages of Thermionic instrumentation recording equipment. Medical Research establishments, Industry, Aviation and Universities are using this equipment to put the T.A.P.E.C. Technique to work on their data acquisition problems. Why not write and find out how it can help you?

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EE 01 006 *for further details*

Precision METAL FILM Resistors

Alma

TWO RANGES

Further development by ALMA has resulted in the introduction of two ranges of Metal Film Resistors. The MA series is fully encapsulated and is expected to meet DEF 5114, whilst the MC series has a relaxed specification and is less expensive.

MA SERIES

TEMPERATURE COEFFICIENT. Better than ± 0.002% per °C.

STABILITY. Better than ± 0.1% for 2,000 hours use at full load.

ACCURACY. ±5%, ±10%, ±0.5%, ±0.2%, ±0.1%

RATINGS & DIMENSIONS.

MA1 1 watt $1\frac{1}{4}$" long × $\frac{3}{8}$" dia. 10Ω to 2MΩ.
MA1/2 $\frac{1}{2}$ watt 1" long × $\frac{3}{8}$" dia. 10Ω to 750 kΩ.
MA1/4 $\frac{1}{4}$ watt $\frac{13}{16}$" long × $\frac{3}{8}$" dia. 10Ω to 400kΩ.
MA1/8 $\frac{1}{8}$ watt $\frac{5}{8}$" long × $\frac{1}{8}$" dia. 10Ω to 150kΩ.

Fully encapsulated in epoxy resin tube.

MC SERIES

TEMPERATURE COEFFICIENT. Better than ± 0.005% per °C.

STABILITY. Better than ± 0.2% for first 1000 hours use at full load.

ACCURACY. ±5%, ±1%, ±0.5%.

RATINGS & DIMENSIONS.

MB1 1 watt $1\frac{1}{4}$" long × $\frac{5}{16}$" dia. 10Ω to 2MΩ.
MB1/2 watt $\frac{7}{8}$" long × $\frac{5}{16}$" dia. 10Ω to 750kΩ.
MB1/4 $\frac{1}{4}$ watt $\frac{5}{8}$" long × $\frac{5}{16}$" dia. 10Ω ro 400kΩ.
MB1/8 $\frac{1}{8}$ watt $\frac{7}{16}$" long × $\frac{3}{16}$" dia. 10Ω to 150kΩ.

Sealed in epoxy resin coating.

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COMPLETE NEW RANGE OF CERAMIC TRIODES

Efficient high-frequency operation in industrial heating applications

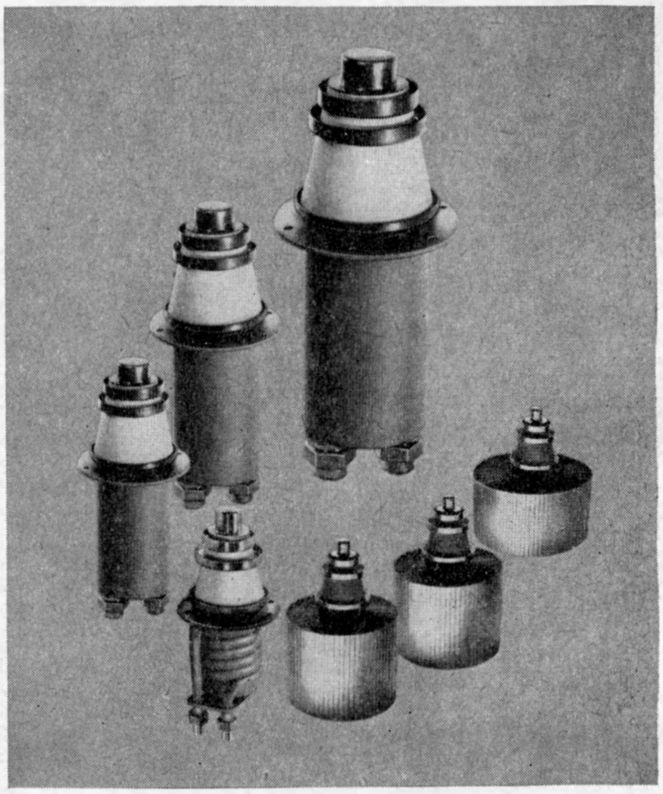
A 3·9 to 240kW range of ceramic triodes for industrial applications is the latest product of Mullard's 46 years experience in the high power transmitting and industrial valve market. Mullard ceramic triodes, specifically designed for industrial heating, are produced by an entirely new technique resulting in a valve design which is physically robust, reliable, efficient and electrically tolerant of the arduous operating conditions encountered in typical factory process work.

Small dimensions, combined with the low-loss performance of the ceramic material, ceramic/metal seals and coaxial connections used in the valve assembly, give extremely good performance in the v.h.f. band: a smaller electrode structure of optimised design also adds to the high standard of performance.

To permit complete equipment design flexibility, alternative cooling arrangements are provided for some of the valves; cooling methods are by forced air or by water flow through a jacket or integral helix. A special helix technique is used on the higher power valves in the range.

Improvements in electrical characteristics are achieved by measures which increase the grid dissipation reserve and the effective emission area of the cathode. Anode voltage surge tolerance is assured by an anode voltage maximum rating which is nearly double the operating figure quoted, whilst the filament can tolerate temporary variations within −10% to +5% of its nominal value. Added to these advantages, valves in this range are designed to operate at a low anode voltage with high efficiency and output performance, features which can considerably lower the overall cost of systems in which ceramic triodes are used.

Approx. Output at Full Ratings (kW)	Type Number	Cooling	Max. Frequency (MHz)	P_a Max. (kW)	V_a Max. (kV)	I_k Max. (A)
3·9	YD1150	air	160	2·5	7·2	1·4
	YD1151	water jacket				
	YD1152	integral helix				
8·8	YD1160	air	160	5·0	7·2	2·8
	YD1161	water jacket				
	YD1162	integral helix				
15·4	YD1170	air	120	10	7·2	4·8
	YD1171	water jacket				
	YD1172	integral helix				
30	YD1182	integral helix	80	20	8·4	8·0
60	YD1192	integral helix	30	40	9·6	14·5
120	YD1202	integral helix	30	60	14·0	20
240	YD1212	integral helix	30	120	15·6	30



Selections from the Mullard range of ceramic triodes

Valve quality is guaranteed by an exhaustive test procedure backed by an extensive research programme in the Mullard applications laboratories. Here, long-term trials of these valves are carried out in test equipment under simulated operating conditions.

Mullard's conservative rating policy is the final feature which ensures that, so long as the user adheres to the recommended operating conditions, maximum valve life and reliability are assured because the margin between these conditions and the valve's maximum limits is so large that it is very unlikely that the latter will ever be exceeded, even under the most adverse conditions likely to be encountered in industrial use.

For further details of these valves, please use the reply card of this journal (see reference opposite).

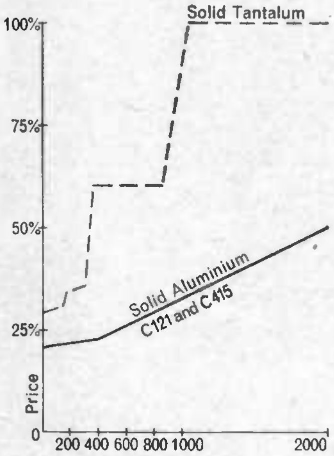
What's new . . . . . from Mullard

Electrolytic Capacitor Breakthrough

inexpensive alternatives to solid tantalum types

A breakthrough in electrolyte techniques enables Mullard to offer a comprehensive range of aluminium electrolytic capacitors in which the normal liquid electrolyte has been replaced by solid semiconductor material. This unique construction gives these components, even under severe environmental conditions, the advantages normally associated only with solid tantalum types, namely, extreme reliability, excellent long term stability and good low temperature characteristics. The shelf-life and reforming problems associated with wet and the so-called 'dry' types are also eliminated.

Reliability trials carried out over 15 000 000 component hours have provided failure rate and service life figures which are compaiable with those achieved with their more expensive solid tantalum counterparts. Lower material costs allow Mullard to offer items in the solid aluminium range at substantially lower prices than corresponding tantalum components. This factor, as shown on the adjacent chart, becomes increasingly important as the CV product is increased.



Cost comparison between solid aluminium and tantalum electrolytic capacitors.

The initial release of Mullard solid aluminium capacitors was made on a restricted basis under type number C415 with capacitance-voltage values limited to 500. Now in full production, further introductions under type number C121 have extended the CV product of this range to 2200.

Working voltages

4V	C415	16 to 100μF
	C121	180 to 390μF
6·3V	C415	12·5 to 80μF
	C121	150 to 330μF
10V	C415	8 to 50μF
	C121	100 to 220μF
16V	C415	5 to 32μF
	C121	56 to 120μF
25V	C415	3·2 to 20μF
	C121	39 to 82μF
40V	C415	2 to 12·5μF

From this brief introduction, it is obvious that, for similar performance, solid aluminium elecrolytics offer substantial price advantages over tantalum types: they cannot, of course, equal the minimum dimension characteristics of the latter type.

BTY34 Thyristors—

New Rationalised Range

The Mullard 6·4A thyristor range has been recently rationalised to provide devices with 100V, 200V 300V and 400V ratings in place of the original BTY33 to BTY39 voltage selection in which the devices were rated in multiples of 50V. Type number BTY34 with a voltage suffix is used for items in the new rationalised range. The new releases also have the added attraction of a higher junction temperature ($T_jmax = 150°C$), a feature on which it is intended to base a completely new 'higher temperature' range of thyristors.

Now in pioduction at the Mullard Stockport factory, the BTY34 range of thyristors is offered on an immediate availability basis.

TO-5 2AMP Transistors

ACY17 Family Uprated

Transistors in the Mullard ACY17 family—one of the most important groups of germanium p-n-p devices available today—have recently been uprated from a 1A to a 2A device level. This uprating has opened up a new range of applications in the medium speed switching fields; as an example, these devices can now be used to drive OC28 power transistors over their complete switching range. Junction temperature ratings have also been increased and peak pulse powers of 5W can now be handled over the full current range. All of these features add up to give this family of devices an extremely attractive performance/cost ratio.

Transistors in the ACY17 family are medium speed devices for use in switching, pulse, oscillatory and general purpose industrial circuits. The applications coverage of the family group has also been extended to cover low-noise wideband applications by the addition of two new devices, type ACY44 and its CV equivalent type CV7740, both of which have a maximum noise figure of only 4dB over the 200Hz to 10kHz range. Special selections of both linear and switching types can be made available to meet individual customer requirements.

To assist the designer, information on the family has been completely revised and its range extended to cover operation up to 2A. It now includes operation under transient and repetitive pulse conditions, and gives transient thermal resistance curves, avalanche characteristics and energy ratings in the avalanche region.

Items in the ACY17 family together with their 'CV' equivalents are:

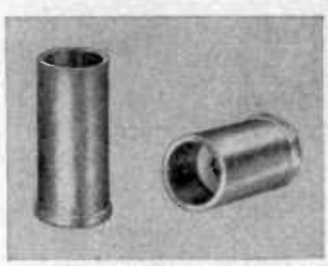
ACY17 (CV7376)
ACY18 (CV8130 & CV7436)
ACY19 (CV7437)
ACY20 (CV7438)
ACY21 (CV7439)
ACY22
ACY39
ACY40
ACY41
ACY44 (CV7740)

Low Noise Mixer Diodes AAY50 and AAY50R

Low noise mixer diodes AAY50 and AAY50R (development types 79AAY and 80AAY) have been designed for X-band equipments using the standard British type of coaxial diode. The AAY50 and AAY50R form a reverse pair and are plug-in replacements for types SIM2/5 (CV2154/5) and GEM3/4 (CV7108/9).

The measured overall noise figure at X-band is typically 6·7dB including a 2·0dB i.f. amplifier. This represents an improvement on types CV2154 and CV7108 of about 3·0dB and 1·5dB respectively.

The AAY50 and AAY50R conform to the SO-26 coaxial outline and are hermetically sealed.



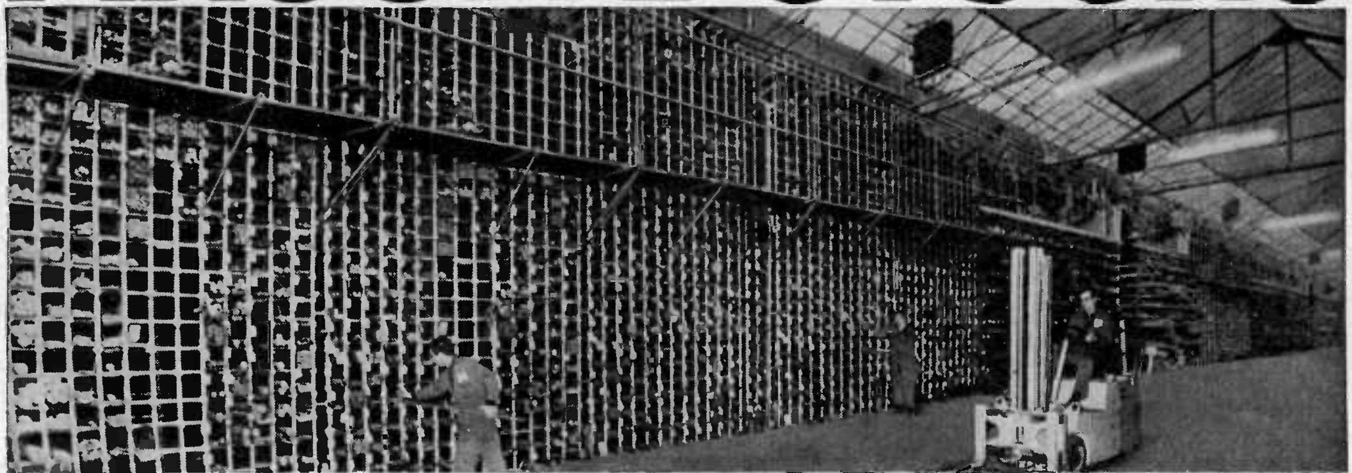
FURTHER DETAILS of the Mullard products described is this advertisement can be obtained from the address below or through the Reader Information Service of Electronic Engineering using the appropriate code number shown below.

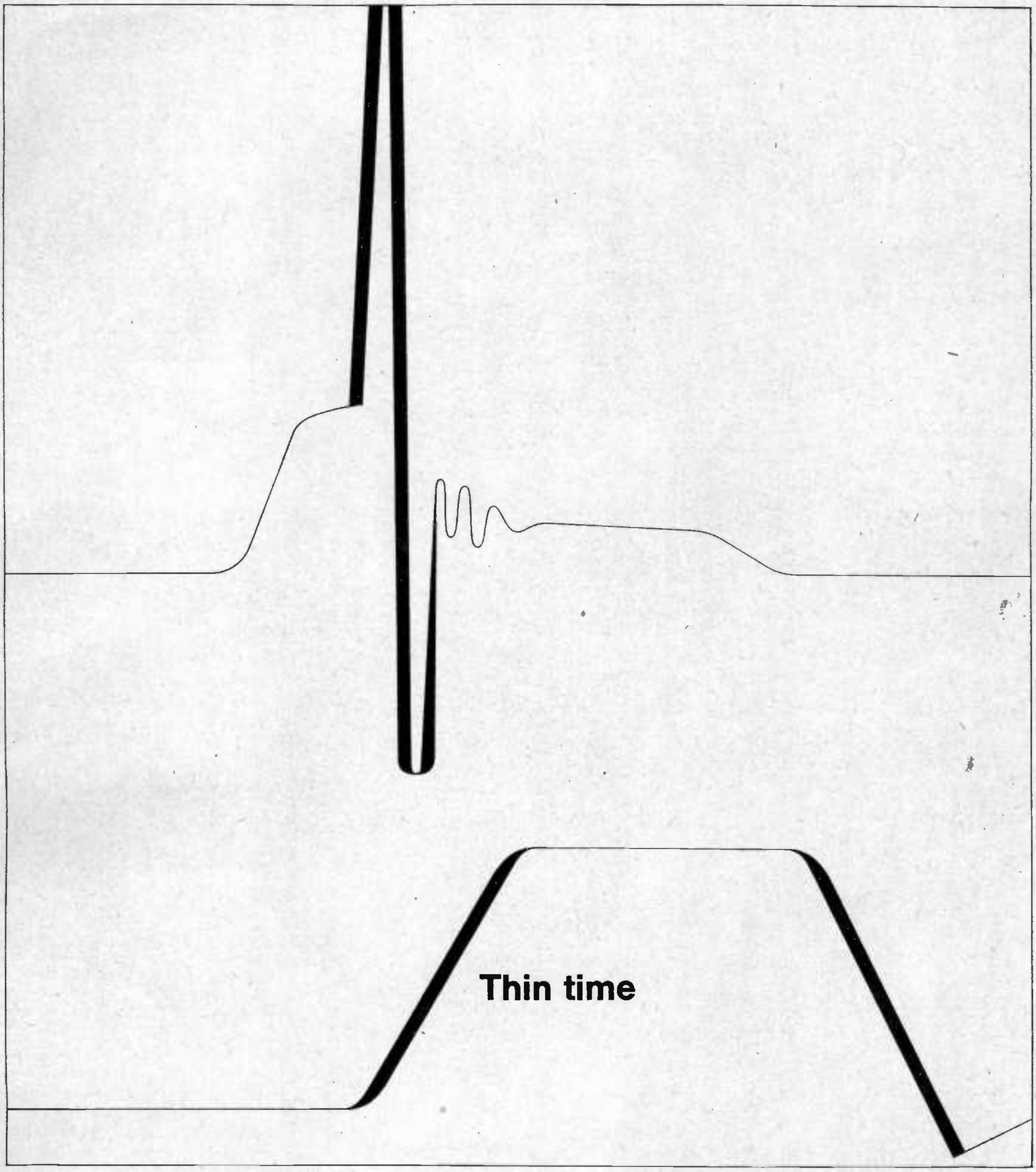
Ceramic Triodes . . .	EE 01 350
C121, C415 Aluminium Electrolytics	EE 01 351
BTY34 Thyristors . .	EE 01 352
AAY50, AAY50R Mixer Diodes	EE 01 353
TO-5 2A Transistors	EE 01 354

Mullard

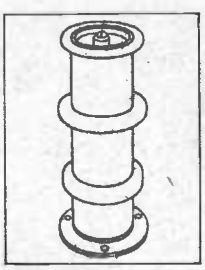
Mullard Limited, Mullard House, Torrington Place, London, WC1. Telephone: LANgham 6633

CAM43





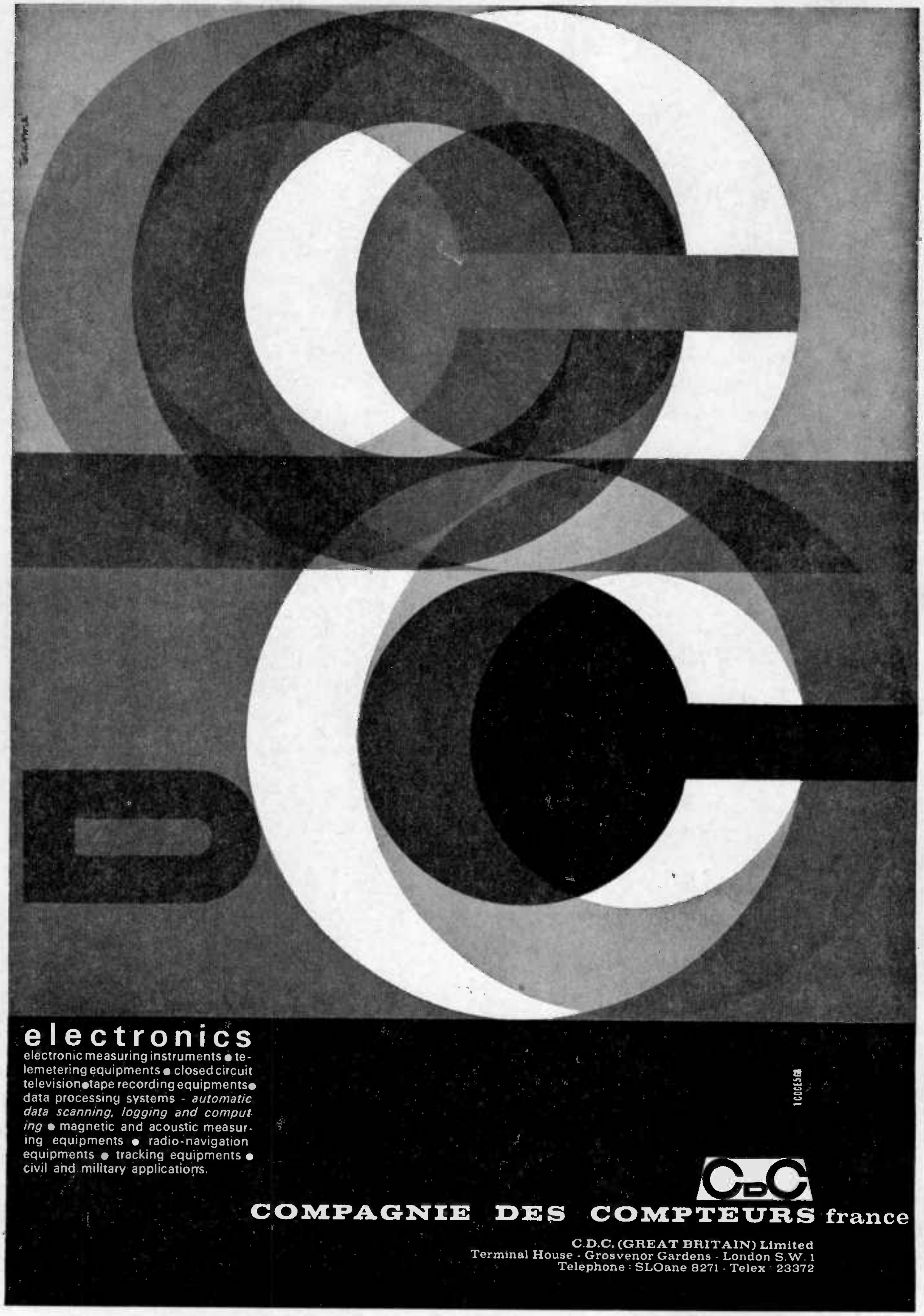
English Electric Valve's range of ceramic hydrogen thyratrons is unique. Because each is a tetrode with inherently low dynamic inductance the firing time may be made accurate to less than one nanosecond, and pulse rise times of less than 50 nanoseconds. The anode delay time drift is shorter and the trigger powers required are considerably less than with other types of thyratron.



For high speed switching applications ceramic thyratrons are better than the corresponding glass tubes wherever high-peak, high-mean characteristics have to be met. Higher hold-off voltages (40kV per gap) are possible by using specially designed thyratrons with deuterium filling. EEV will be glad to consider special development and manufacture for customers' own requirements.

ENGLISH ELECTRIC VALVE COMPANY LIMITED

CHELMSFORD, ESSEX, ENGLAND. TELEPHONE: CHELMSFORD 53491 TELEX: 99103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- * AMPLITUDE STABILITY
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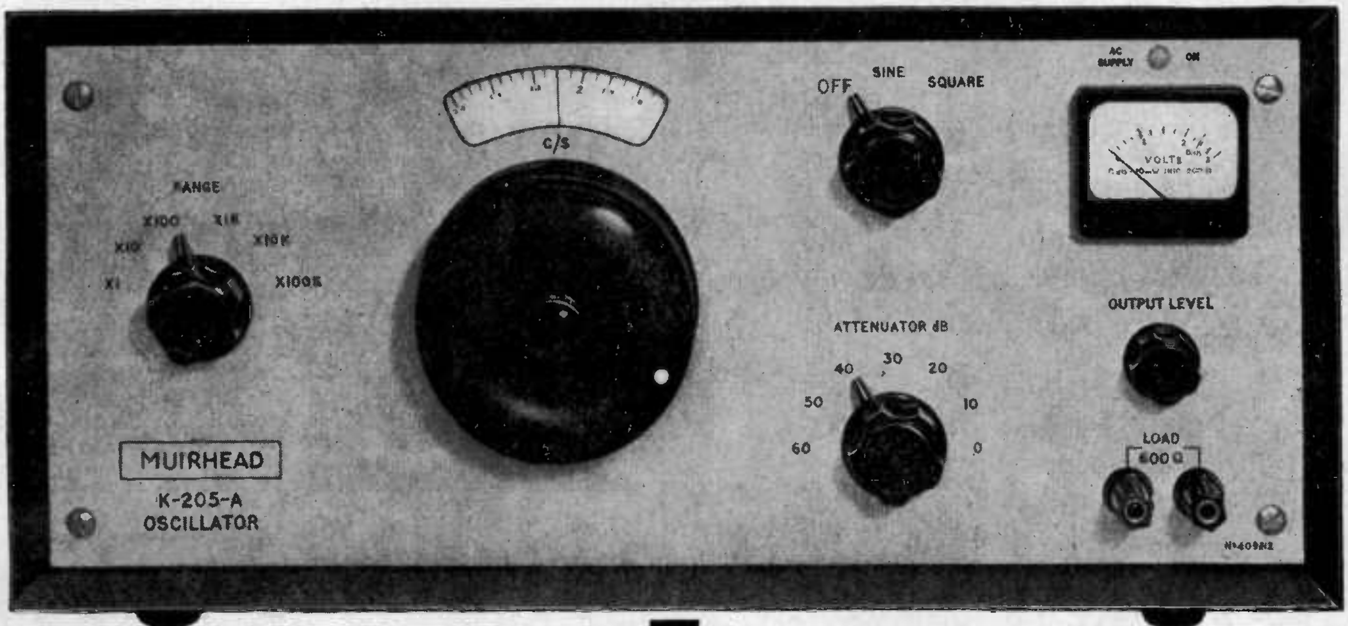

. . . AND THE PRICE?

Muirhead ask £175 for their 1 c/s—1 Mc/s continuously variable oscillator. No other equipment offers such features at the price—frequency stability and accuracy, a high amplitude stability, low distortion and sine or square wave outputs. Special circuit techniques assure the high amplitude and frequency stability. Any decade between X1 and X100K can be selected on the six-position range switch. Accuracy of setting over the entire frequency range is ensured by a 50 :1 precision slow-motion drive. Frequency is ± 1% above 10 c/s. Spot frequencies to an accuracy of 0.005% can be obtained by coupling to a Muirhead frequency comparator.
The Muirhead K-205-A is ideal for laboratory or field work and operates from AC mains supply or low voltage external D.C.

* £175

Please send me full details on the Continuously Variable Oscillator

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MUIRHEAD & CO. LIMITED

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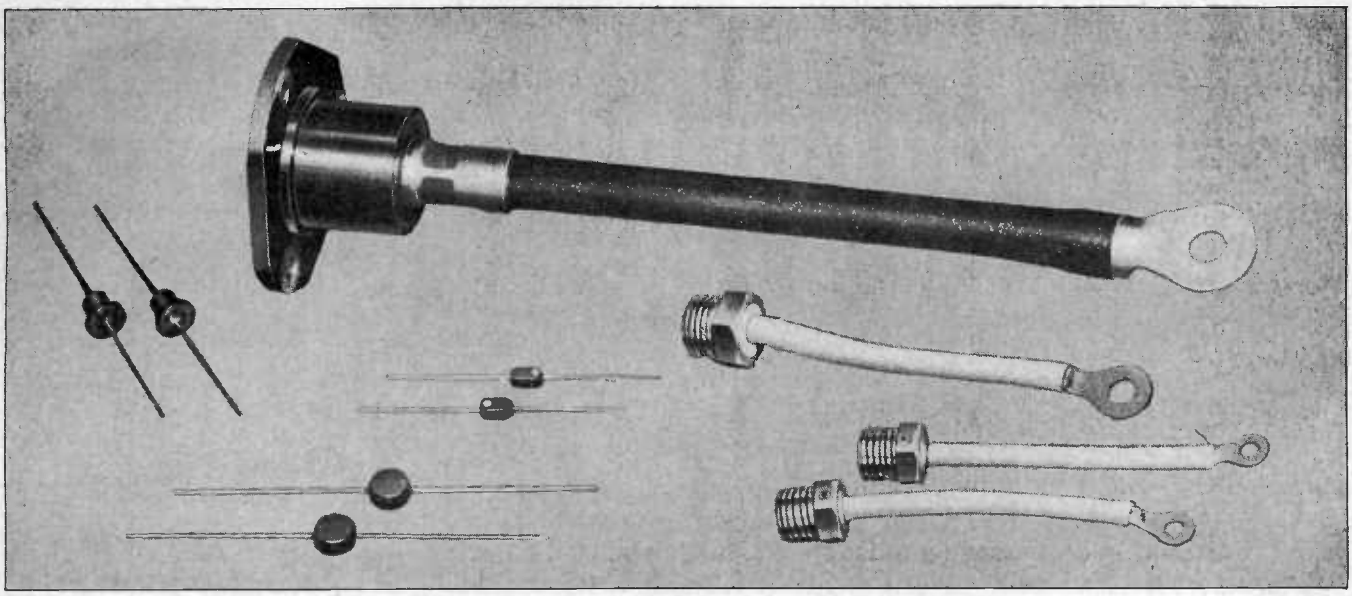
EE 01 015 *for further details*



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Shindengen is far in front as Japan's top producer in the semiconductor field. Shindengen research not only has a long history —but today the products coming from this research satisfy most of Japan's domestic demand and are exported to every industrialized nation of the world. Strict quality control over mass production techniques results in efficiency as well as economy. Especially, the following three silicon elements have many applications in entertainment and industrial electronic equipment:

TYPE	OUT PUT CURRENT(D.C.A)	PEAK REVERSE VOLTAGE(V)
S1 A	0.1	600, 800, 1000, 1500
S2 E	0.5	600, 1000
S2 A	1.5	200, 400, 600, 800, 1000

SHINDENGEN ELECTRIC MFG. CO., LTD.

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MAIN PRODUCTS:
- Silicon Rectifier
- Silicon Bridge Rectifier
- Power Transistor
- Thyristor
- Silicon Symmetrical Diode
- Selenium Rectifier

For further information contact the following agencies:

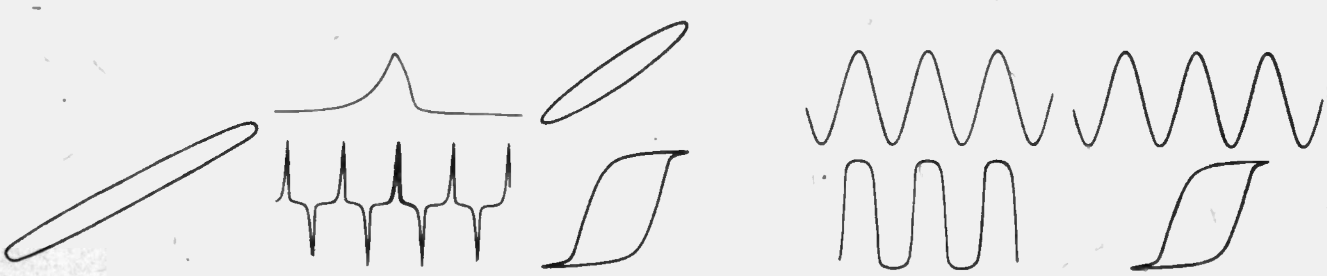
* Scandinavia: **Gadelius & Co., A.B.**, 1A Stockholm Ö Sweden (Phone: 08/23 28 00)
* E.E.C. Districts: **Gloire Trading Co., Ltd.** Central P.O. Box No. 430 Osaka, Japan

EE 01 016 *for further details*

EE 01 019 *for further details*

How many ?

How many instruments do you need to get these displays?



X-Y plots with 100 μV/cm sensitivity

The entire signal plus a magnified portion

Two signals against a common third

Two signals versus time simultaneously

X-Y and time plots simultaneously

How many instruments do you need to get these displays?

One!

The new hp model 132A dual-beam scope.

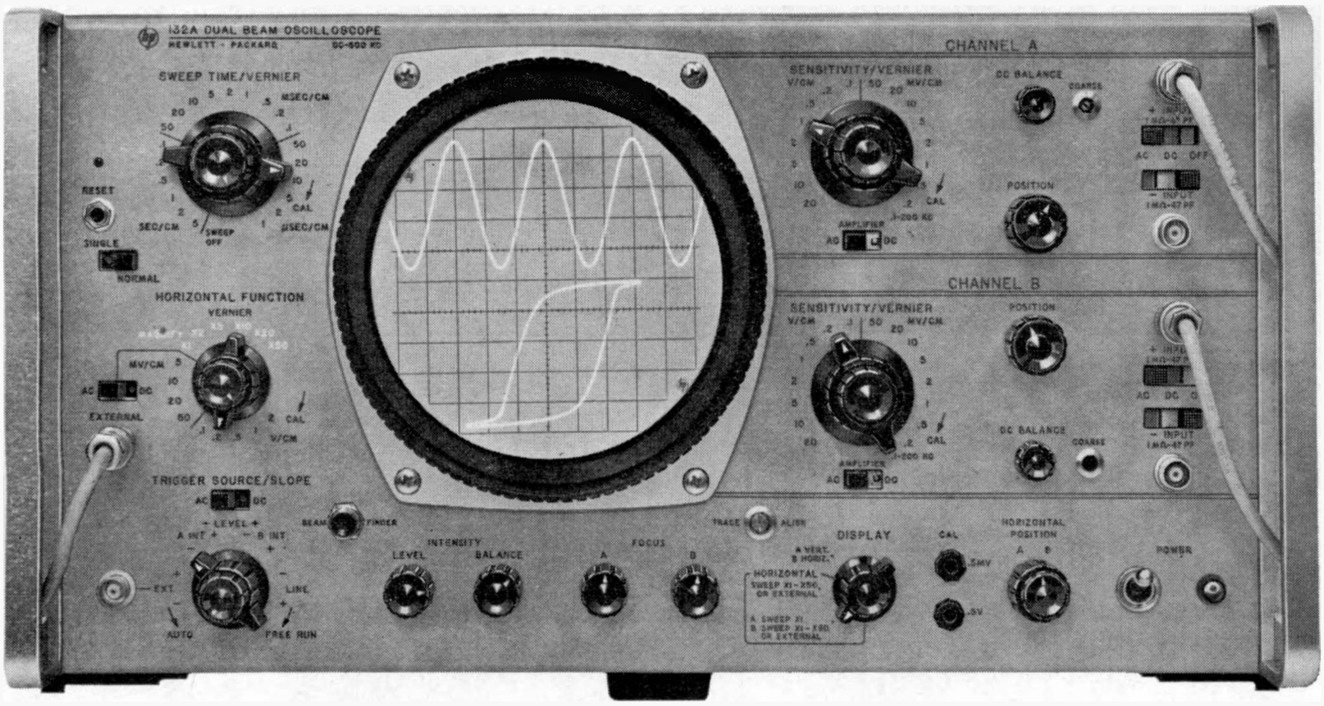
All X-Y or Y-T measuring problems having one, two or even three independant low-level signals can be solved with the hp 132A, due to its unique combination of performance and features:

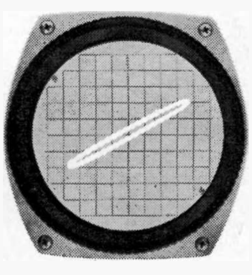
Two completely independant beams to display signals at different sweeps or make simultaneous X-Y and Y-T plots

Two 100 μV/cm vertical amplifiers with a 4000:1 common mode rejection up to 200 KHz, 1 mV/cm to 20 V/cm up to 500 KHz

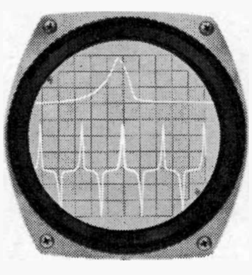
Recorder outputs from each vertical amplifier

Price: £ 495

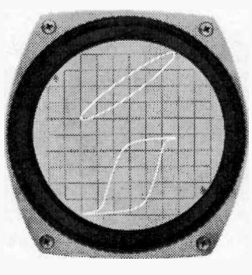




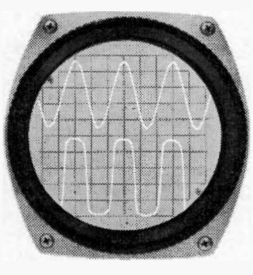
X-Y plots with 100 μV/cm sensitivity
For X-Y plots of low-level signals, just plot A against B. No need for external pre-amplifiers, sensitivity is already 100 μV/cm. Phase shift is less than 2° up to 50 KHz.



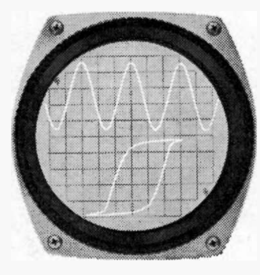
The entire signal plus a magnified portion
Channel B sweep can be magnified up to 50X, while leaving channel A unmagnified. Using horizontal shift any part of the unmagnified wave form can be observed and identified.



Two signals against a common third
Plot both amplifiers against a common 5 mV/cm 300 KHz X-amplifier. Measure phase shift between two circuit points. Relative phase shift between two channels is 2° up to 10 KHz.



Two signals versus time simultaneously
Measure gain, delay time and pulse response. For comparing two signals at the same time the 132A is a regular dual-beam scope. No need for low-level signal preamplifiers.



X-Y and time plots simultaneously
Use channel B and the horizontal amplifier for the X-Y plot while displaying the signal versus time on channel A. Ideal for servo and audio work previously requiring two scopes.

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HEWLETT hp PACKARD
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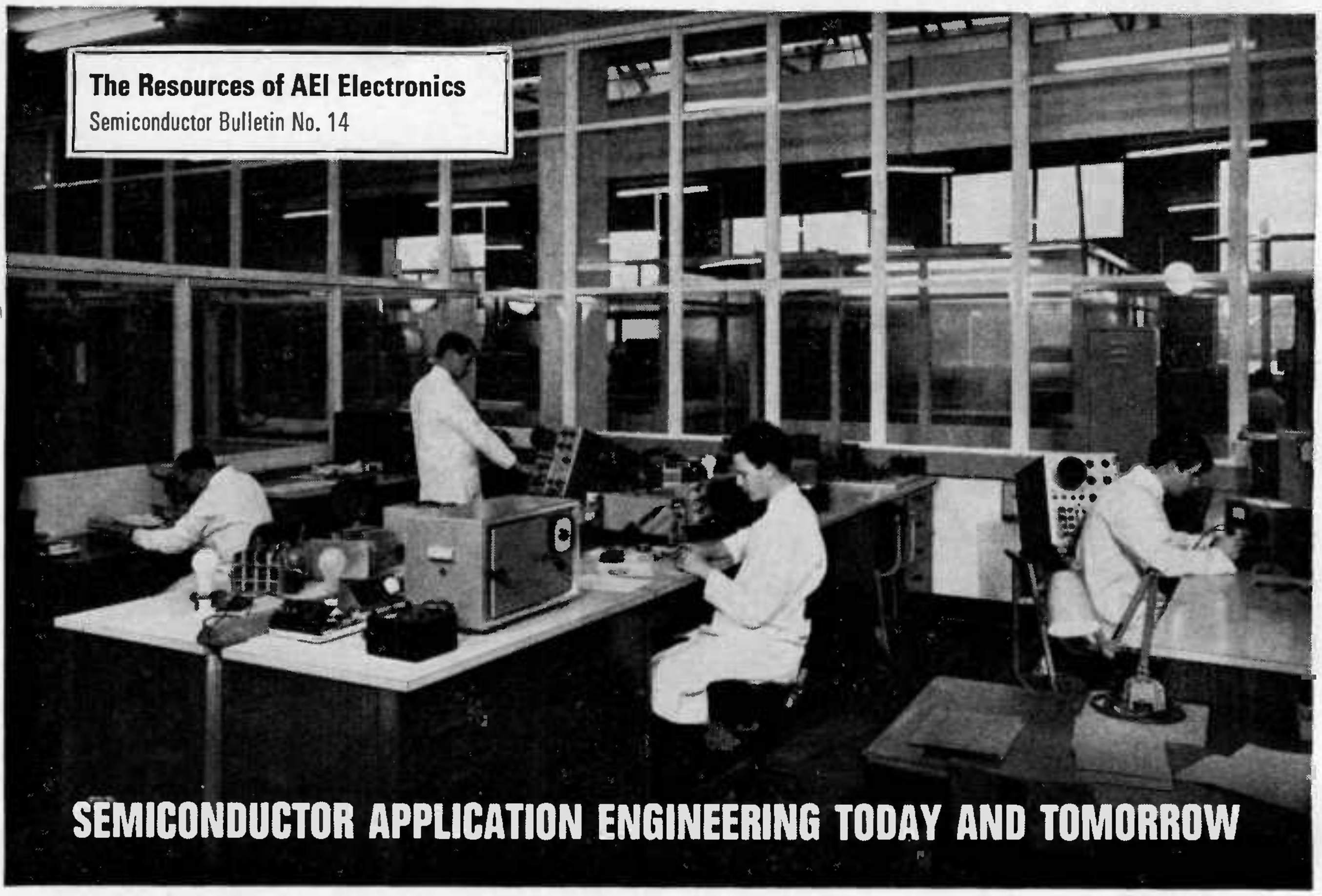

Fig. 1 Part of the Customer Engineering Department at AEI's Lincoln semiconductor plant. Research undertaken by the department in recent months includes development of a control circuit for a domestic food mixer, a new type of drive control for printing machines, and advanced temperature control equipment.

An earlier bulletin in this series—namely, HOW TO ASSESS THE RELIABILITY OF RELIABILITY REPORTS: AEI Semiconductor Bulletin No. 13*—explained why it is that a report allegedly defining the reliability of semiconductors may itself be unreliable. It was shown that this curious state of affairs results partly from faulty statistics and partly from inadequate sampling and testing. It was then demonstrated that by testing an adequate sample and employing sound statistical principles, it is perfectly possible to achieve really reliable reliability reports. This bulletin considers a rather different topic which, although closely connected to reliability, is, if anything, of even greater consequence to the non-specialist user of semiconductor devices; namely, semiconductor application engineering.

The nature of control and its limitations

There is a sense in which semiconductor manufacture is more closely akin to arable farming than to conventional engineering. For just as the farmer cannot possibly guarantee that all the grain in a given crop will be to a given standard, so the semiconductor manufacturer cannot possibly guarantee that all the devices in a given production run will meet a particular specification for either electrical properties or physical characteristics. On the contrary, a certain percentage will inevitably fall outside the requisite specification, and another percentage will, equally inevitably, fall well within it. In between there is, of course, a percentage which just meet the prescribed tolerances. In other words, if you plot the standards attained by the devices in a particular run, you will, regardless of whether or not the production line is in control, obtain a unimodal graph of the kind shown in fig. 2. It may be argued that this is true of all forms of batch production. And it is – but not to the same extent.

The exact form of this curve can, within certain limits, be adjusted by the manufacturer. But there is no point in his doing so unilaterally in an arbitrary manner. He must consider the applications in which the devices will be used. And this is where the customer comes in – and the situation sometimes becomes complicated!

The semiconductor manufacturer can, in theory, produce a specially designed device – a planar diode, say, or a planar transistor – for a specific application. This may be an economic proposition if the quantity involved is very large, or if the quantity is reasonably large and the specification so modest that the scrap rate is negligible. But customers frequently require semiconductor devices in large quantities which they also require to have very high standards of both performance and reliability. In general, therefore, custom-built devices are too costly to produce to be acceptable, to any but the largest (and wealthiest) customers.

Importance of practical experience in circuit design

Accordingly, the vast majority of users prefer to select standard devices from the manufacturers' established range, and rely upon circuit design to achieve the particular effects they want. This puts a tremendous onus on the manufacturer – and for two reasons.

In the first place because it means that it is very largely up to the semiconductor manufacturer to decide what types of devices to produce in what quantities. In the second place, because solid-state circuit design is still a highly specialised field in which a limited number of people in the 'electrics' industry have practical experience – hence the popularity of the solid-state assemblies and sub-assemblies manufactured by AEI.

It is, in fact, no exaggeration to say that the typical semiconductor user is, unless he happens to be in the electronic field, always in much the same position as the man who has enough bricks to build a house, but does not know very much about architecture. In other words, he has what he needs, but lacks the knowledge to use it to its best advantage.

Of course this is not to suggest that engineers in other fields cannot design circuits involving semiconductors. Most of them understand the principles involved, and many of them can design a circuit that will

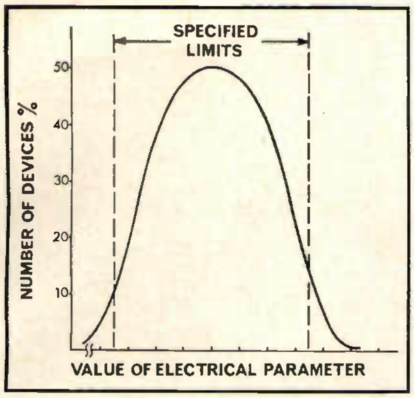

Fig. 2 Graph showing how only a certain percentage of semiconductor devices fall within the specified tolerances for a given parameter.

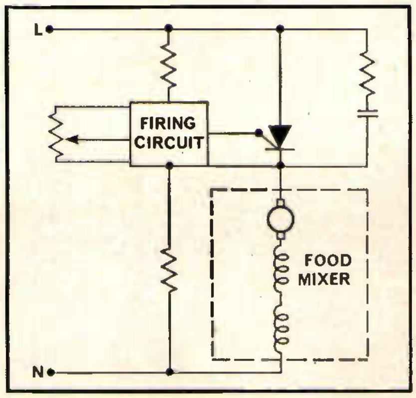

Fig. 3 Control diagram for a food mixer.

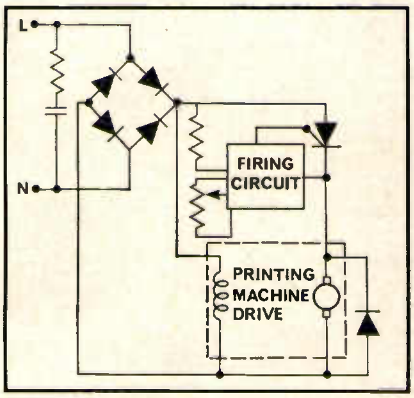

Fig. 4 Solid-state printing machine drive control diagram.

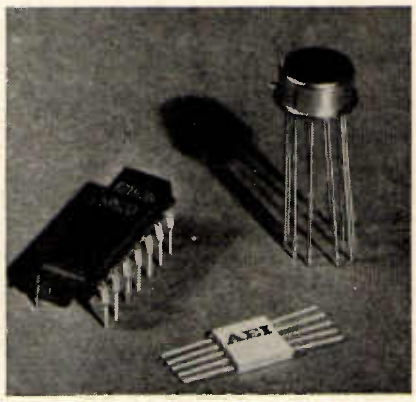

Fig. 5 A selection of the microcircuits now available from AEI Semiconductors.

work – perhaps one that will work efficiently. However, whether their design is either the *most* efficient or the *most* economical is quite another matter.

Food mixer control circuit

Consider for instance the circuit for a food mixer shown schematically in fig. 3. This design is the result of considerable research and development on the part of the Customer Engineering Department of AEI Electronics' Semiconductor factory at Lincoln. A designer inadequately briefed on solid-state techniques would probably have tried to oversimplify the circuitry, in which case the system would certainly have been inefficient, and very probably unreliable as well. A designer with a firm grounding in solid-state theory but too little practical experience would probably have avoided over simplification but have fallen into the equally serious error of oversophistication. Which would certainly have increased the cost without necessarily achieving the same measure of reliability and efficiency as that of the straightforward, but carefully thought-out circuit developed by the AEI Semiconductor Department at Lincoln.

All this is even more true when the circuit in question is a more complex one such as the printing machine drive control shown in fig. 4. For the development of such an assembly calls for knowledge and experience that no manufacturer can gain if he is not constantly engaged in semiconductor technology.

Large customer engineering departments

This explains why more and more semiconductor users in the 'electrics' industry are coming to take it for granted that there are benefits to be gained by discussing developmental work with the component manufacturer. It also explains why it is that the larger manufacturers such as AEI Semiconductors now maintain well-established customer engineering departments.

Microcircuits

These departments are already of primary importance, and in the near future they will become absolutely essential. For it is clear that the next major stage in the evolution of semiconductor technology will undoubtedly be the widespread commercial use of the monolithic or microcircuit. AEI Semiconductors are already supplying a wide range – of which a selection is shown in fig. 5. – to users in this country, and their service-proven advantage of dependability and economy make their acceptance on a broad front a certainty.

When this happens much of the responsibility for semiconductor circuit design must inevitably pass from the customer to the semiconductor manufacturer. No doubt there are in industry today numerous 'rugged individualists' who will find this fact irritating. But the only alternative open to them is to go into the semiconductor business themselves. And even if they were to do this right now, they would have to spend a great deal of time, and more money, just catching up with developments, let alone keeping pace with them in the future.

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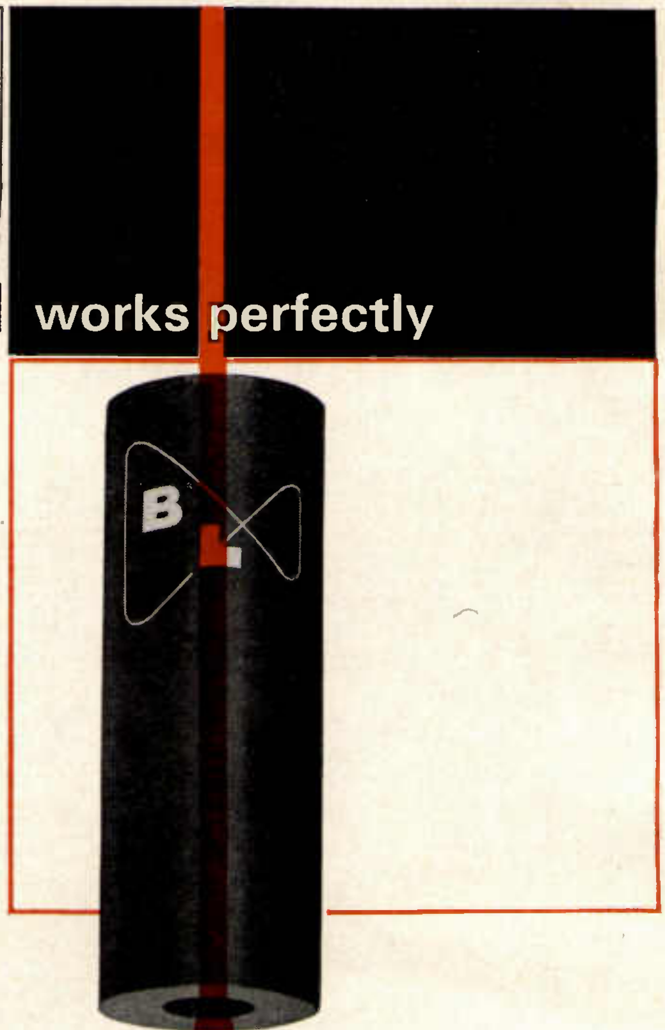

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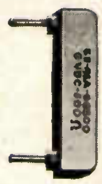

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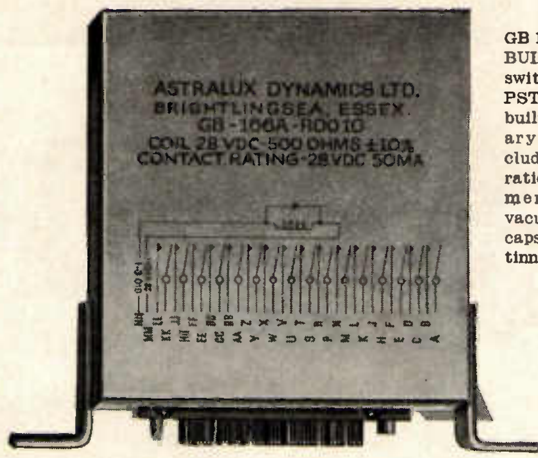

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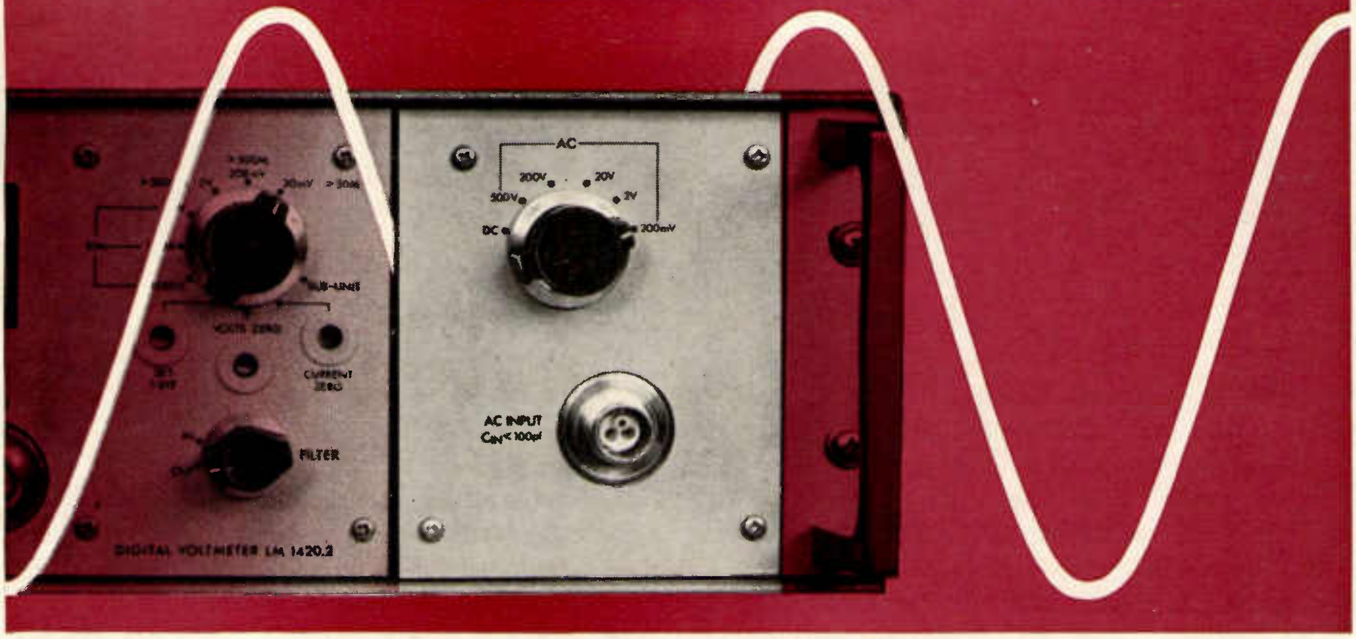

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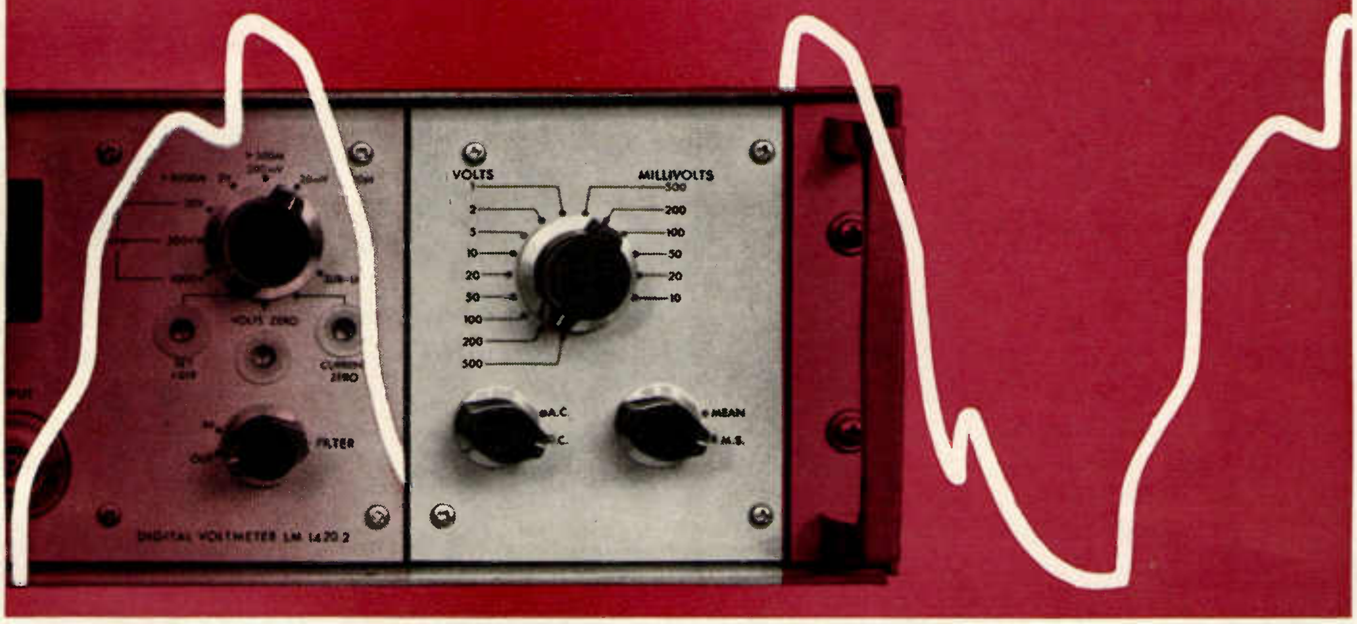

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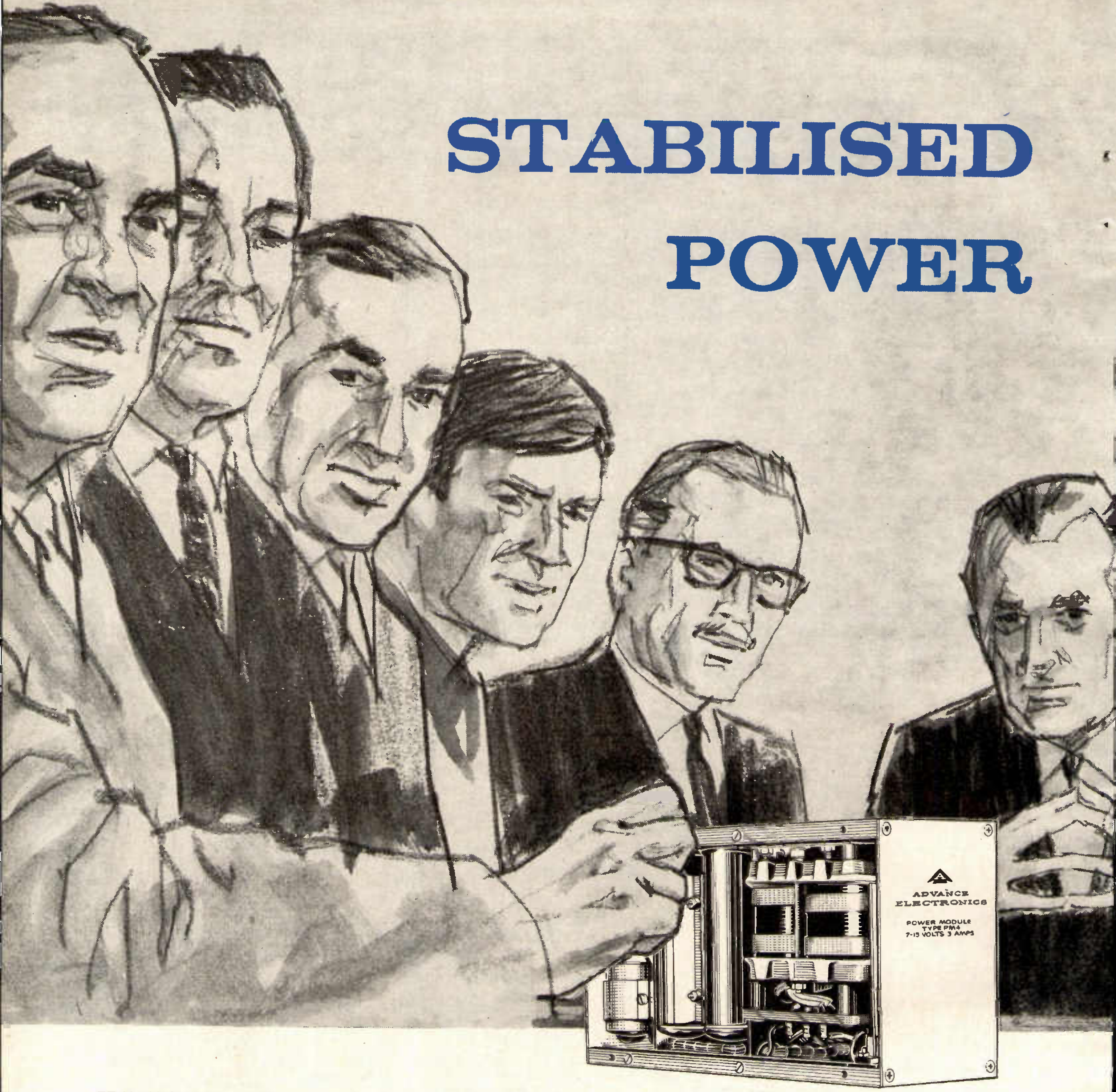

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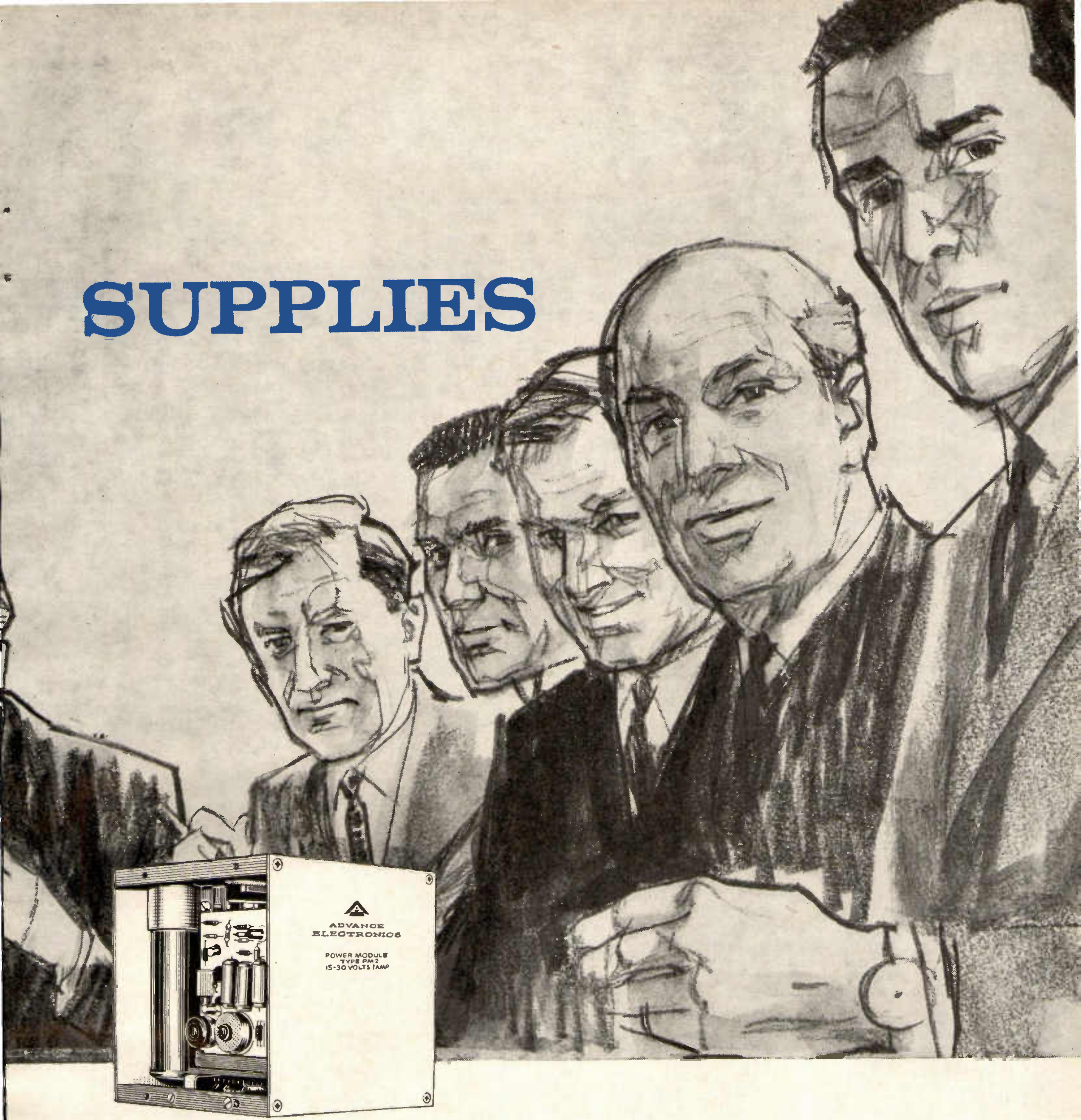
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BC107	45	100	260	250	110	2
BC108	20	100	260	250	110	2
BC109	20	100	260	300	210	2
BF173	25	25	260	550	38	7
BF167	30	25	130	350	27	4
BF180	20	20	130	675	14	2
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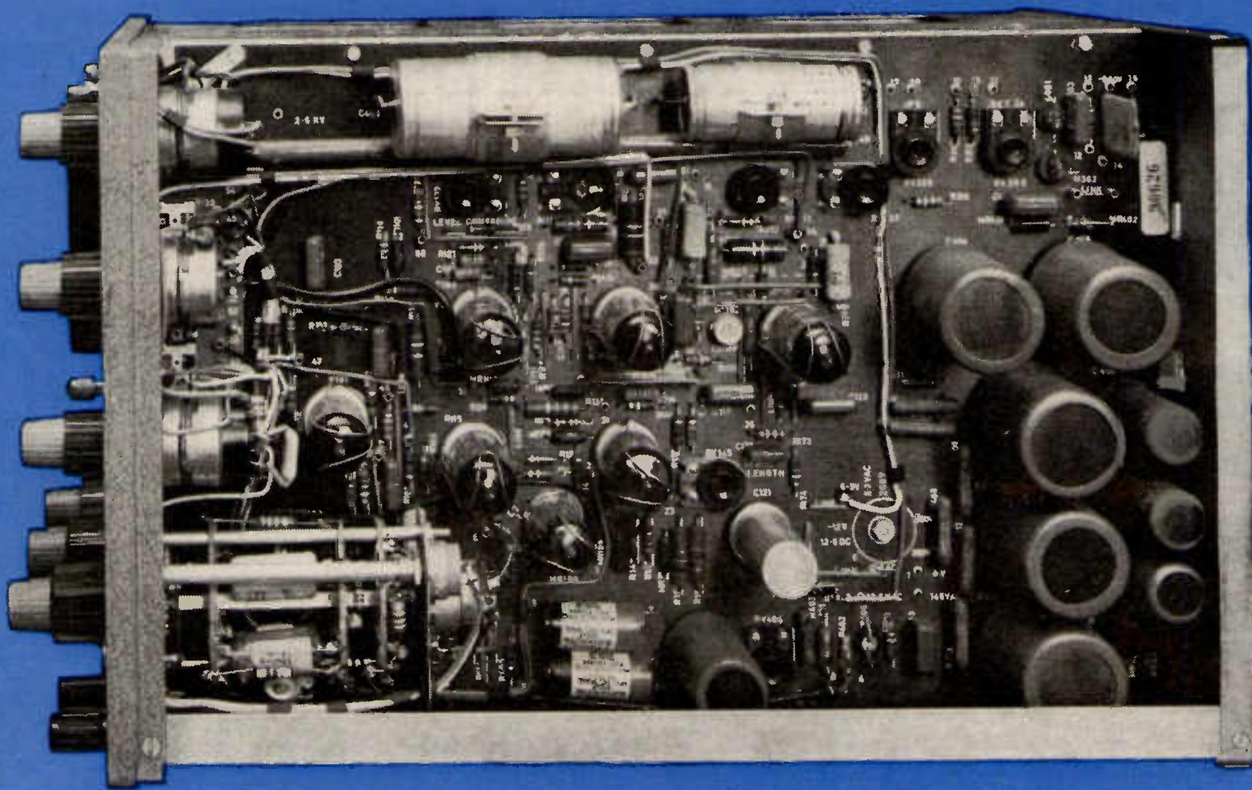

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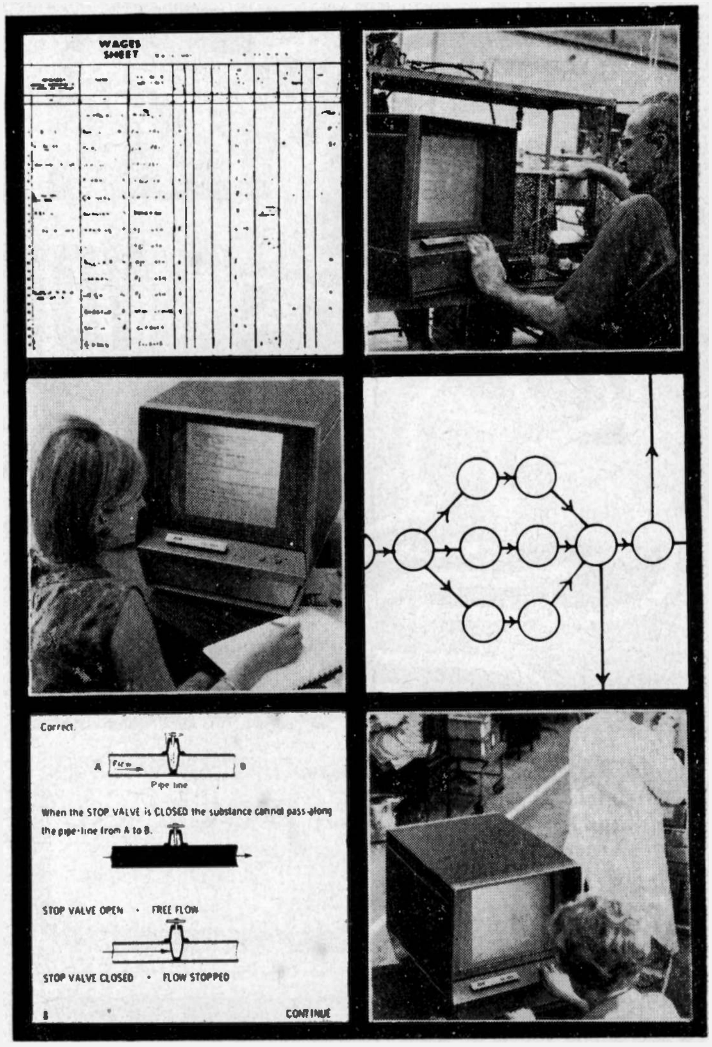

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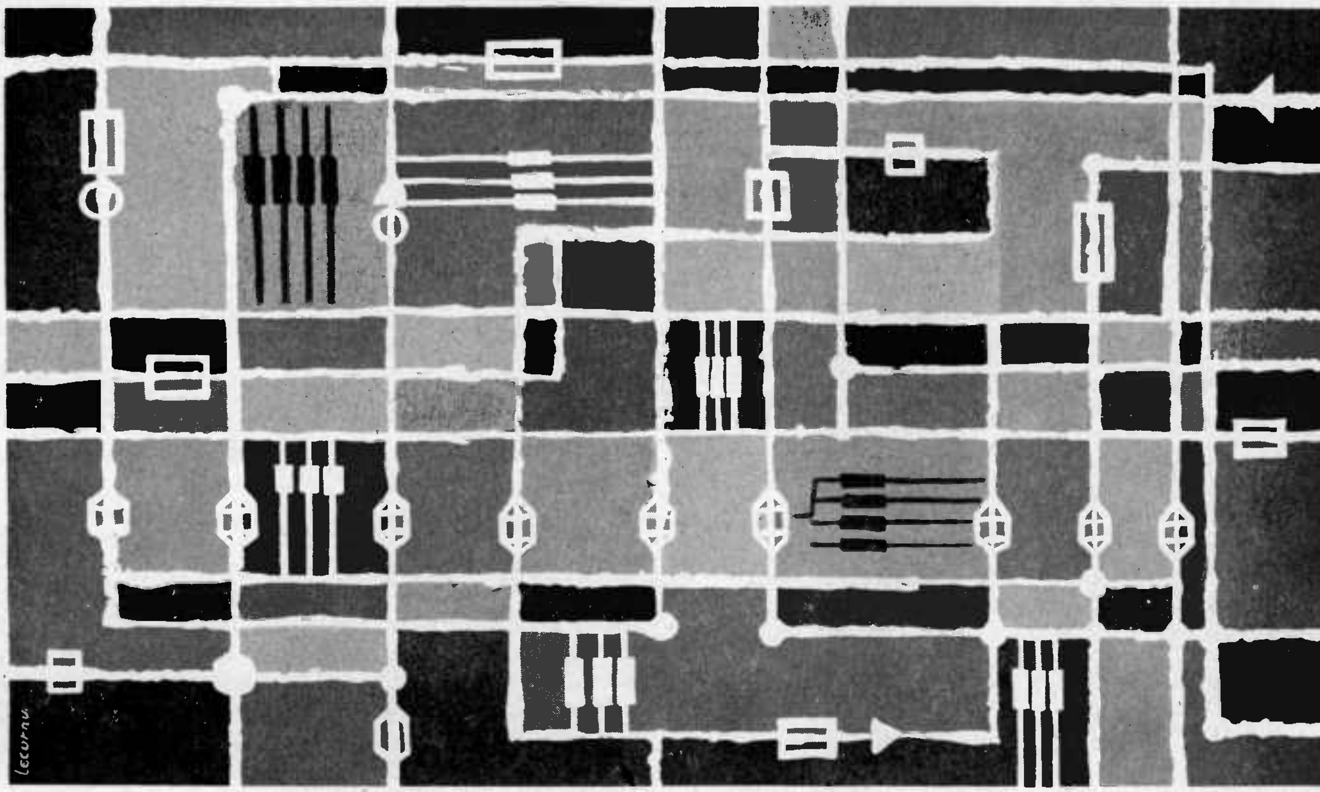

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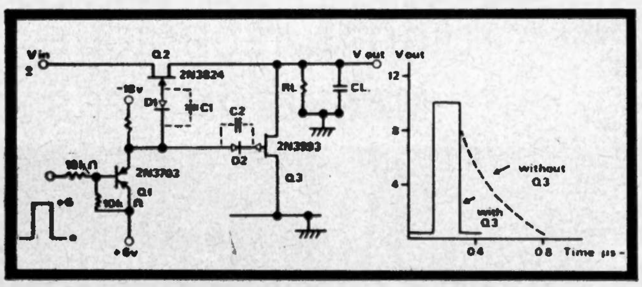
FETs improve Performance



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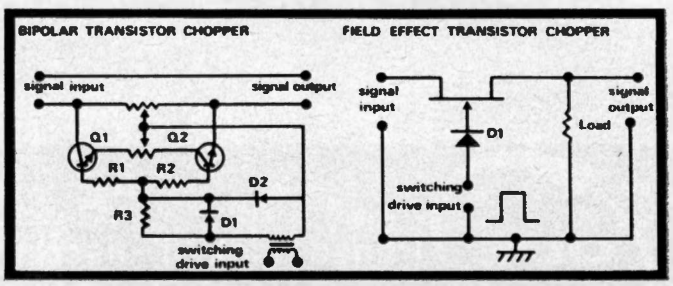
FETs Simplify Circuit Design



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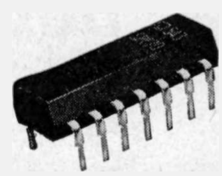

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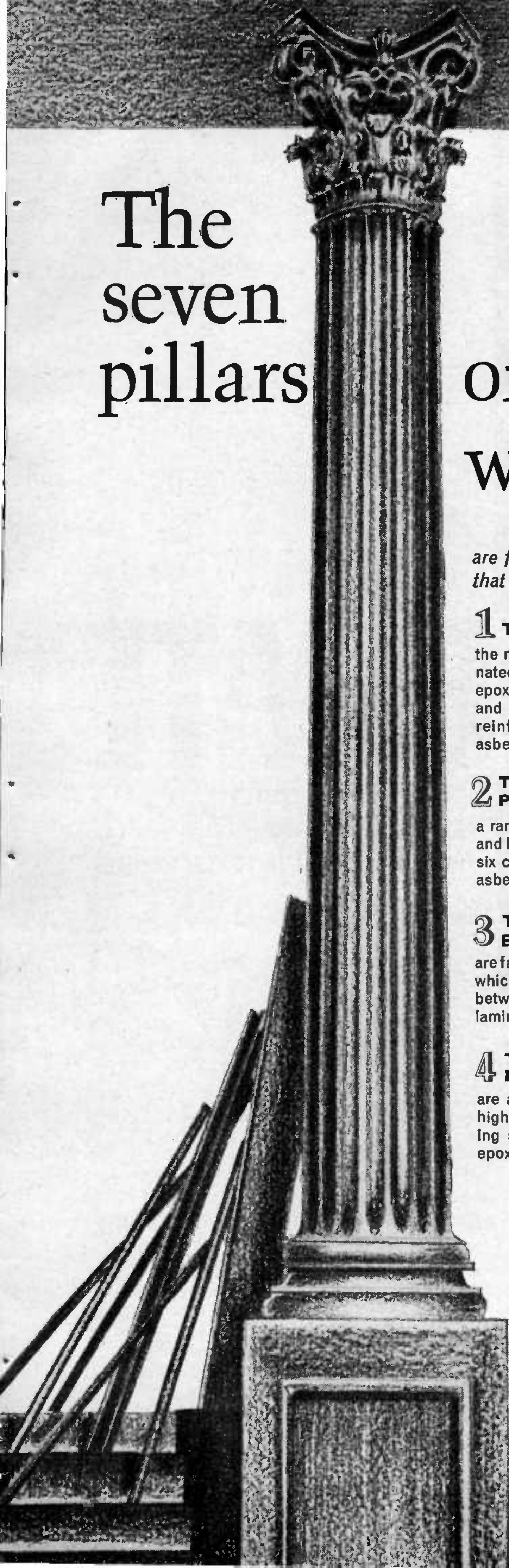

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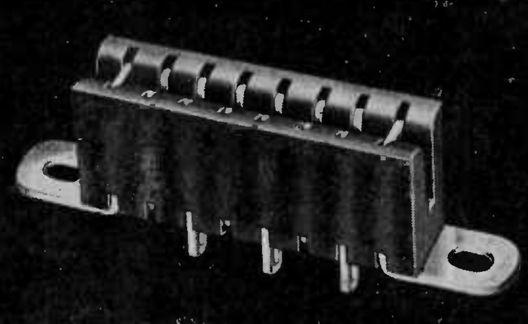

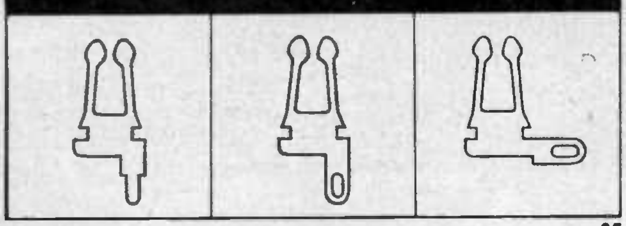

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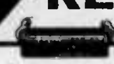

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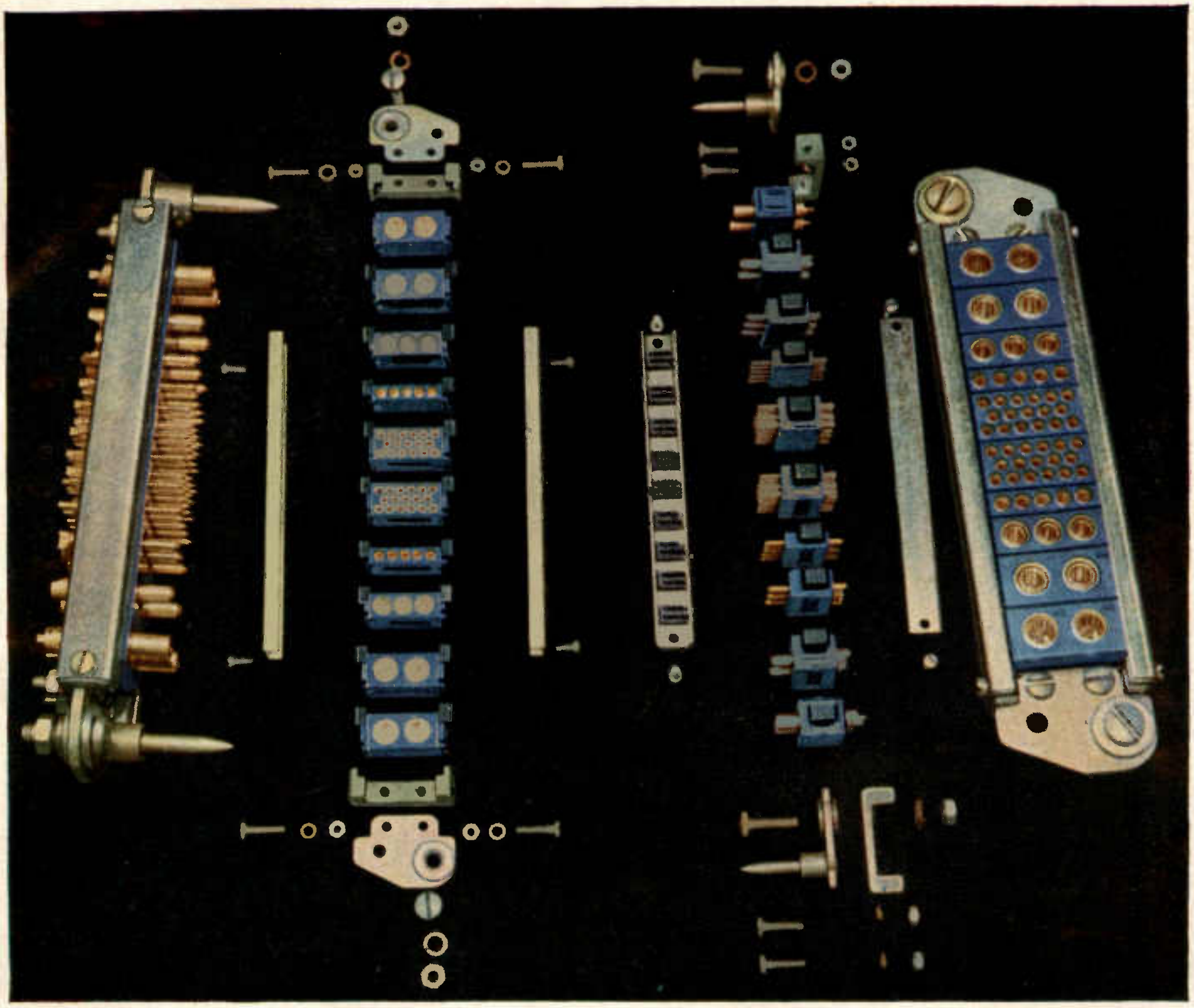

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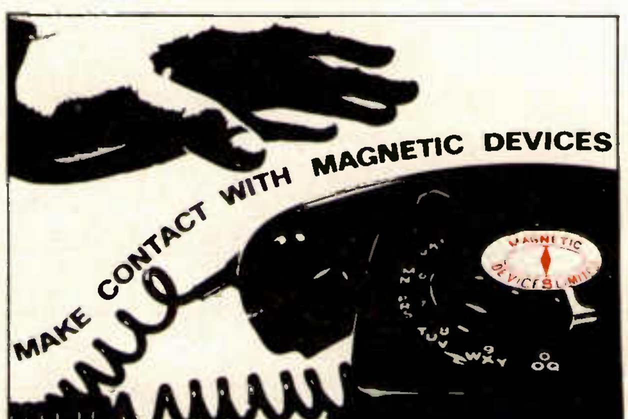

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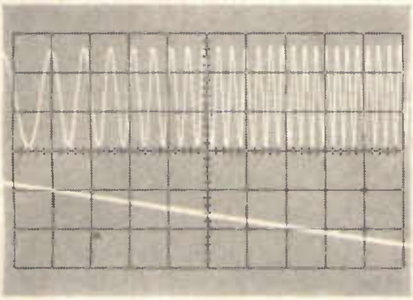
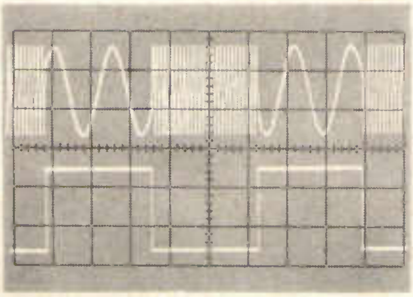
Programmable output frequency

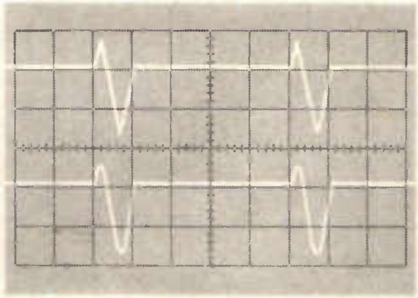


Frequency: upper trace, programming below

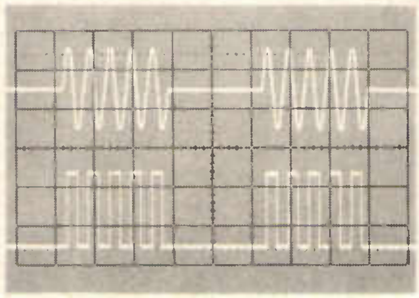


Frequency: upper trace, programming below

Single or multiple cycle bursts

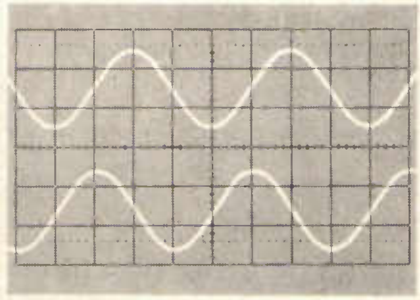


Single cycle—simultaneous sine and triangular outputs

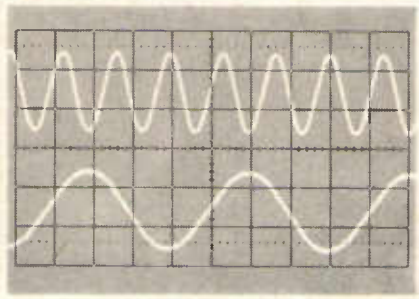


Multiple cycle—bursts of simultaneous sine and square waves

Phase locked output



Phase lock—3300 A output (upper trace locked to external fundamental)



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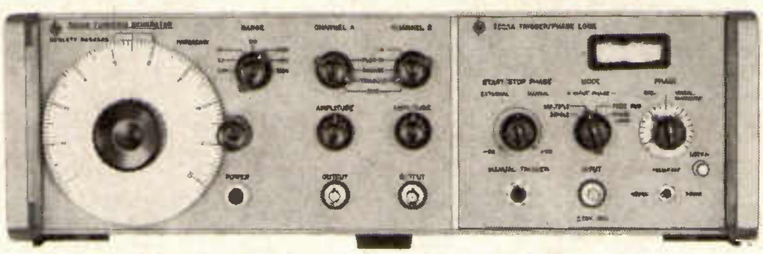

... in hp plug-in function generator

These actual trace photos illustrate just a few examples of the output versatility offered by the new HP 3300A plug-in function generator. The HP 3300A provides output frequencies from 0,01 Hz to 100 KHz in sine, triangular and square wave forms. Any two of these wave forms are available simultaneously at a pair of identical 600 Ω outputs. These outputs may be adjusted independently up to a maximum peak value of 15 V. Full data is available from your Hewlett-Packard distributor.

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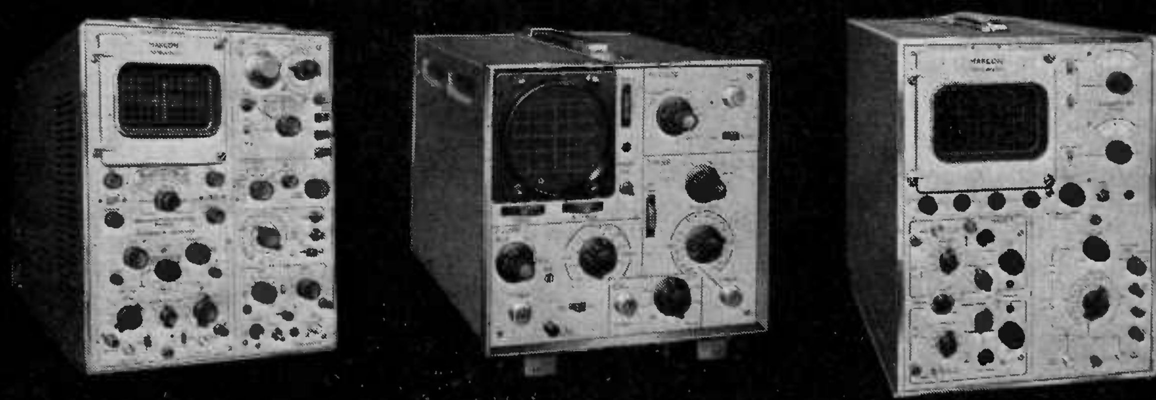

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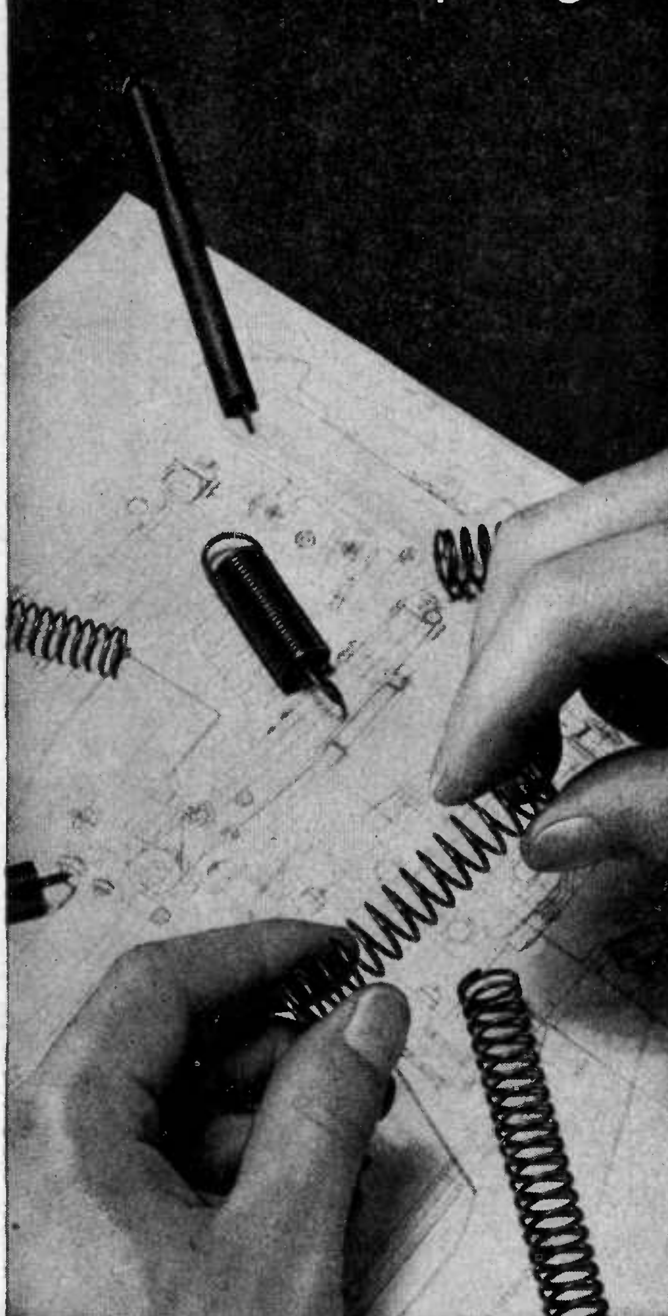

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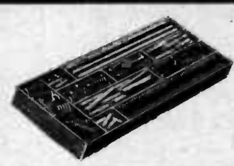

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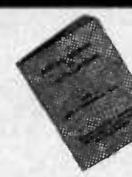

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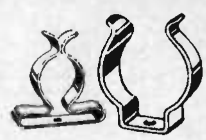

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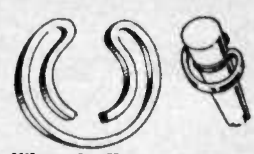

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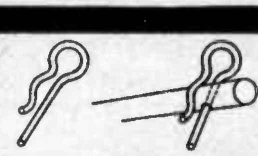

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Moral: If you count above Three, you need help. Get it from CI.

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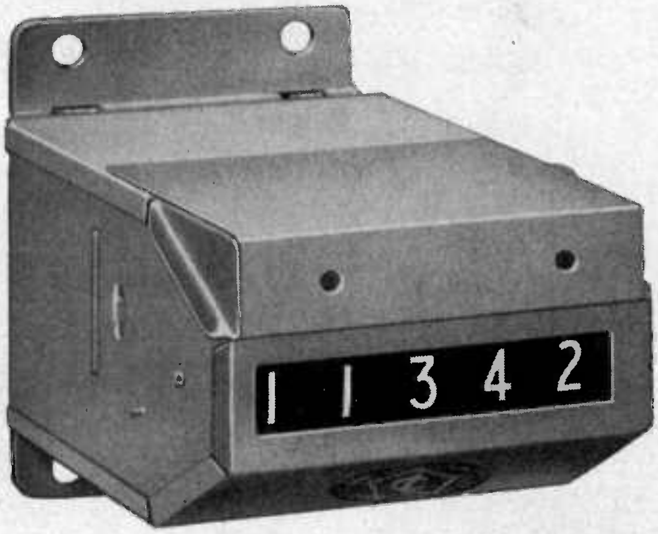

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✳Type 1209 (illustrated)—as 1200 but with flush mounting.
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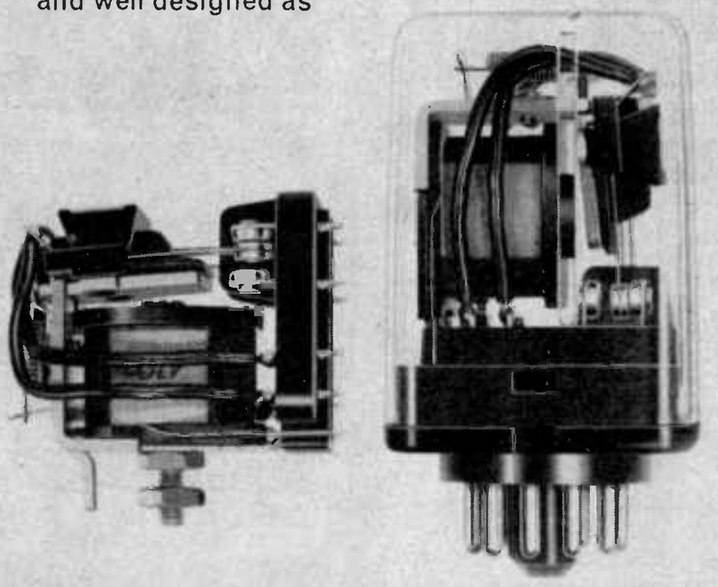

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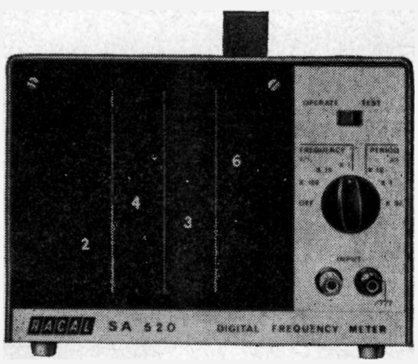

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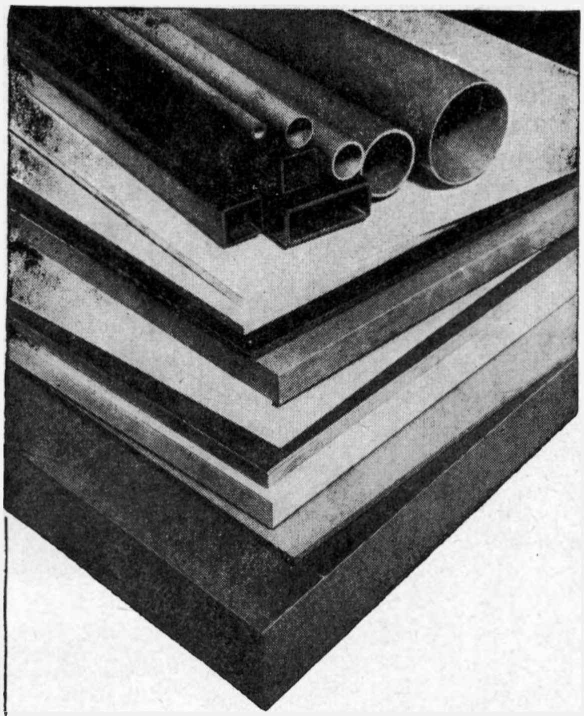

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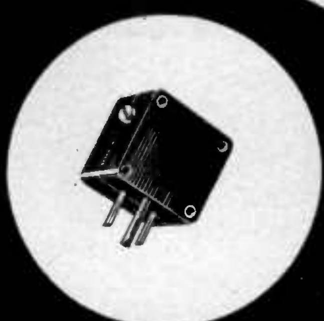

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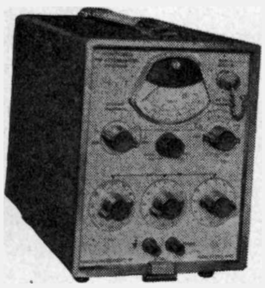

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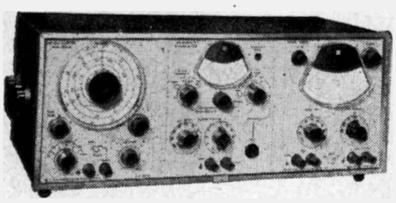

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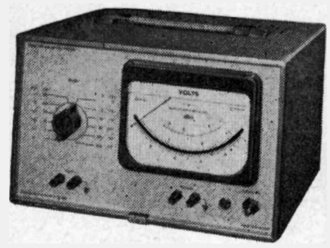

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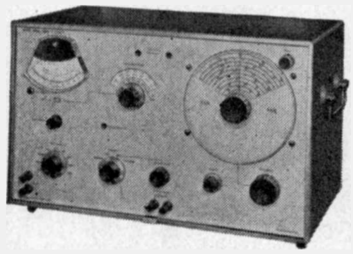

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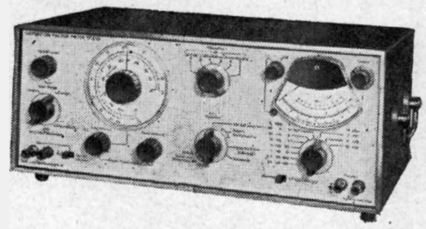

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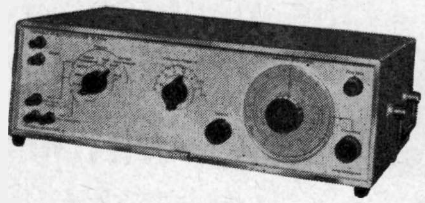

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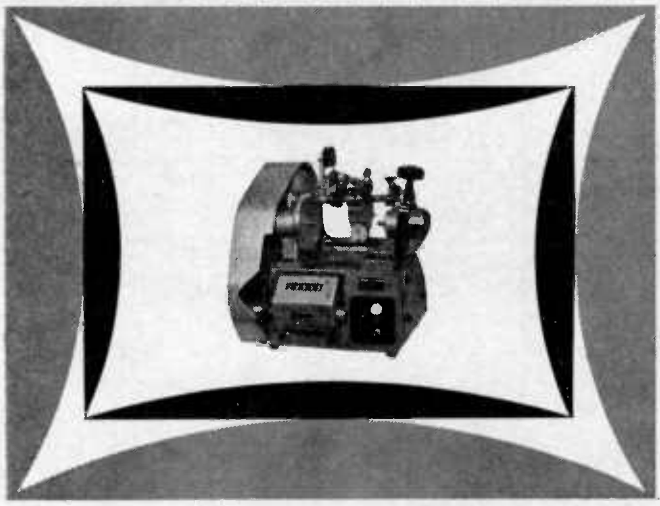

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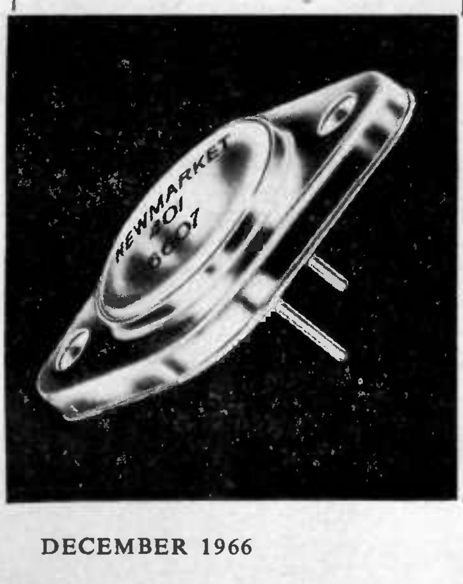

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Commentary

THE thirtieth anniversary of the opening of the world's first public high-definition television service was commemorated on 2 November. To mark the occasion a film, presented by Mullard Ltd in association with the BBC, received its first screening on 3 November. The film, entitled 'The Discovery of Television', set out to show, in the space of fifty minutes, the early history of television, and to answer the question, 'who invented television'?

As the film so aptly shows, it was not the work of any one man, but was rather the culmination of the struggles of many. The story may well be said to have begun over a hundred years ago when, in 1863, the Abbé Casselli transmitted a picture over the wire service running the 80 or so miles between Amiens and Paris, where the picture was reproduced on tin foil. Ten years later, engineers experimenting with submarine cables off the coast of Ireland discovered the photo-resistive properties of selenium and the possibilities of using this effect to 'see by electricity' soon became apparent. When, shortly after, Alexander Graham Bell invented the telephone, schemes to combine the two techniques in a 'seeing phone' were soon forthcoming.

Indeed, by the 1890's 'television' appeared to be not so far away, but further progress proved impossible until after Fleming had invented the thermionic valve. At this point it is interesting to recall that as early as 1897, the cathode-ray tube had been developed by the German, Karl Braun, although not, of course, with television in mind.

A quite remarkable forecast of the evolution of television over the next 25 years was made by A. A. Campbell-Swinton when he gave his Presidential Address to the Röntgen Society in 1911 and expanded on ideas put forward three years earlier in *Nature.*

Three years later, in 1914, Isaac Schoenberg, a mathematician, arrived in this country from Russia and went to work for the Marconi Company, of which he eventually became general manager. At about the same time John Logie Baird appeared on the scene, and these two men, Baird and Shoenberg, were to be the key figures in the race to get a practicable television system started.

Baird began experiments with television transmission in an attic in Hastings and his early apparatus included a Nipkow disk, a hat box, a fourpenny 'bulls eye' lens and a darning needle! His early pictures, transmitted across a room, caused little excitement but, in 1925, Gordon Selfridge, put Baird's equipment on show in his Oxford Street store in London. In 1926 Baird gave a demonstration to the Royal Institution, in 1927 he sent pictures from London to Glasgow by wire, while in 1928 he transmitted pictures by short-wave radio to the United States and to the liner *Berengaria* in mid-Atlantic. Baird was very keen to start a national television service and in 1928, after much campaigning by himself and his supporters, a demonstration was arranged especially for the BBC: as a result of this he was granted three, and later five, half hour periods a week of broadcasting outside normal hours. The first Baird transmissions were made on 30 September, 1929, but, as only one transmitter was available the picture was transmitted first, without sound, followed by the sound without picture: the 'programmes' were received on an estimated twenty-nine sets—all that were in operation at that time. In the next few years Baird improved his system and made a number of advances, but throughout, he clung tenaciously to a mechanical system and went so far as to tell an audience in America, "There is no hope for television by means of a cathode-ray tube."

Meanwhile considerable interest was being shown in electronic systems both in this country and in America. In the U.S.A. Vladima Zworykin was heading a powerful team at RCA and was probably leading the field at this time. In this country the Marconi Company and EMI (an amalgamation of the Gramophone Company and Columbia) had formed a joint company, The Marconi-EMI Television Company Ltd, and one of the most powerful research teams ever assembled by a commercial organization was set up at Hayes. It was headed by Isaac Schoenberg and included such men as Condliffe, McGee, White, Caines, Broadway, Browne and Blumein.

In 1933 the battle between Baird and EMI reached a peak and competitive demonstrations between the two systems were held. As a result the Postmaster General set up a committee under Lord Selsdon to advise him on the relative merits of the two systems. This committee decided that a duel should be held. Both systems were to be used alternately 'under strictly comparable conditions': the venue chosen for the transmitting station was Alexandra Palace.

In April 1936 the rivals started to move their equipment into Alexandra Palace and, at the last moment Schoenberg took a very brave decision—to transmit on 405 lines instead of the Selsdon Committee's minimum of 240 lines which Baird intended to use.

So it was that on 2 November, 1936, the first regular television service in the world came into being.

The duel was short lived and in February 1937 it was generally agreed that the results from the Baird system were distinctly inferior to those obtained with the EMI-Marconi system and the Baird system was dropped.

To summarize: although there is literally nothing of Baird's in the modern television system, it was he who made the early running and provided a great deal of stimulus. But it was Campbell-Swinton who proposed the theory, Zworykin who pioneered and patented the electronic system and Shoenberg and his colleagues who created the first national public service.

A Control System for a Molecular Vacuum Gauge

By R. G. Christian*, M.Eng.

An electronic control unit for use with a vibrating vane molecular vacuum gauge is described. The requirements of gauge control are discussed, the gauge sensing and driving arrangements being in the form of a differential capacitor.

The method of sensing used is that of measuring change in capacitance as the vane vibrates by means of a frequency modulation system which provides a voltage which is a function of vibration amplitude. Feedback of this voltage via suitable circuits provides the drive for the gauge. The oscillation amplitude is stabilized by means of a control circuit, the control voltage being a logarithmic function of pressure. This voltage is used to provide a logarithmic pressure scale, being indicated by a valve-voltmeter which, in addition, provides other facilities. Rapid build-up of vane amplitude is provided by means of a starting trigger.

Experimental results are given for air only, although the gauge has been used successfully for helium (mass 4) and hexafluoropropylene (mass 150) as well as with gases having intermediate molecular masses.

(Voir page 830 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 837)

AN oscillating vane molecular vacuum gauge has been developed in which the damping of the vane due to the gas is proportional to the pressure and the square-root of the molecular mass of the gas. In early types of viscosity gauge, the damping, and hence the pressure, were determined by measuring the time taken for the amplitude of oscillation to fall to, say, one half of its initial value. Since the times involved can be very large, some form of automatic control and direct pressure read-out is desirable. The control system described here maintains the vane in approximately constant amplitude and provides a direct continuous pressure indication at pressures below about 10^{-4} torr in air.

Gauge Considerations

The theory and construction of the gauge have been described in detail elsewhere[1] and it will be sufficient here to consider only those aspects which affect the control system. Briefly the gauge consists of a flat aluminium vane suspended from one edge by a short wire suspension so that it is free to oscillate in the manner of a pendulum. At the lower edge of the vane at right-angles to it is fixed a narrow capacitance plate, the vane being suspended so that the capacitance plate is located 1mm above a pair of fixed plates in the form of right-angled triangles. As the vane oscillates the capacitances between the vane and each fixed plate will vary linearly, one capacitance increasing while the other is decreasing, in the manner of a differential capacitor. One capacitance is used for sensing and the other for driving using the principle of the electrostatic voltmeter. The capacitance system is shown in Fig. 1 in which w is the width of the vane capacitance plate, its length being equal to the width of the fixed plates 2 q. The fixed plates have a length of 2 s and are bent to a radius of $(l + d)$ to maintain the separation d, the bottom of the vane oscillating at a radius of l.

The area of overlap determines the capacitance C between the plate and vane and will be

$$C = \frac{\varepsilon_0 wz}{d} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where ε_0 = permittivity of free space and z = length of overlap given by

$$z = q(1 + l\theta/s) \quad \ldots\ldots\ldots\ldots (2)$$

where θ = angular displacement of the vane from the vertical.

The rate of change of capacitance will be

$$dC/d\theta = (\varepsilon_0 w/d).(dz/d\theta) = \frac{\varepsilon_0 wlq}{sd} \quad \ldots\ldots (3)$$

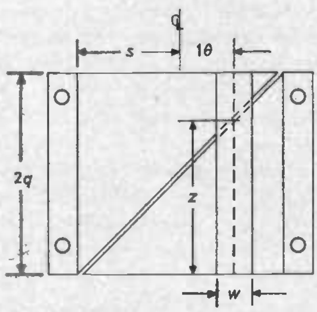


Fig. 1. Capacitance system

which is constant. The restoring torque due to gravity acting on the vane will, for a small displacement θ, be

$$T_0 = mgr\theta \quad \ldots\ldots\ldots\ldots (4)$$

where m = mass of vane, g = acceleration due to gravity, and r = radius of gyration of vane. By analogy with the electrostatic voltmeter the deflecting torque acting on the vane will be

$$T_D = \tfrac{1}{2}V^2(dC/d\theta) \quad \ldots\ldots\ldots\ldots (5)$$

where V = the steady deflecting voltage required between the vane and one fixed plate.

From equations (3), (4) and (5) the value of the direct voltage necessary to produce a steady deflexion of θ is obtained:

$$V = \sqrt{\frac{2mgr\theta sd}{\varepsilon_0 wlq}} \quad \ldots\ldots\ldots\ldots (6)$$

* *Liverpool College of Technology, formerly at the University of Liverpool.*

If the angle θ_o represents the initial deflexion of the vane at a time $t = 0$ then the energy stored in the system is

$$w_o = T_o\theta_o = mgr\theta_o^2 \quad \text{.........} \quad (7)$$

If the vane is now allowed to oscillate freely with no driving voltage, the amplitude will decay due to the damping of the gas and the suspension. After a time

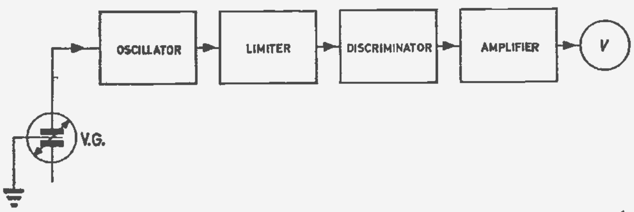


Fig. 2. Amplitude sensing unit

equal to the period of oscillation τ the amplitude will fall to

$$\theta_\tau = \theta_o e^{-a\tau} \quad \text{.........} \quad (8)$$

where a = decay constant due to the damping, and the stored energy will have fallen to

$$w_\tau = T_\tau\theta_\tau = mgr\theta_\tau^2 \quad \text{.........} \quad (9)$$

The loss of energy during this period will be

$$w_L = w_o - w_\tau = mgr(\theta_o^2 - \theta_o^2 e^{-2a\tau}) = w_o(1 - e^{-a\tau}) \quad .. \quad (10)$$

If the damping is small then $1 - e^{-2a\tau} \simeq 2a\tau$ hence, the energy loss per cycle is

$$w_L \simeq 2a\tau w_o \quad \text{.........} \quad (11)$$

For the vane to be driven at constant amplitude θ_o it will be necessary to make up this energy loss every cycle. If the voltage required to produce a steady deflexion θ_o is V_o then the energy stored would be proportional to V_o^2, thus the voltage required to maintain a constant amplitude must be

$$v = V_o\sqrt{2a\tau} \quad \text{.........} \quad (12)$$

which may be supplied in the form of a pulse once every cycle. By means of a suitable circuit such a pulse may be provided and will enable the vane to be driven up to full amplitude from standstill without the need for any initial direct voltage V_o. Since a is proportional to the pressure and square root of molecular mass of the gas, it follows that the drive voltage v is also proportional to these entities, hence a measure of v could be used as a measure of gas pressure.

For the gauge used the calculated dimensions were m = 96·8mg, r = 6·79cm, τ = 0·523sec, d = 1mm, $q = s$ = 2·5cm, w = 1cm, l = 10cm and θ_o = 0·15 radian hence using equation (6) yields a value for V_o of 1 475V. The decay constant a has a calculated value of 28·3 P where P = pressure of air in torr, hence for air at 10^{-5} torr the required drive voltage V from equation (12) is 24·9V. This is the voltage required to maintain oscillations in a system damped only by the gas. The stiffness of the suspension will contribute a constant damping and the drive voltage must make up the ensuing losses, so that equation (12) may be written as

$$v = V_o\sqrt{[2\tau(a + a_s)]} \quad \text{.........} \quad (13)$$

where a_s represents the decay constant due to the suspension alone. No numerical value of a_s could be calculated and it was not measured until the control system was in operation.

The Amplitude Sensing Circuit

Since much of the experimental work on the vacuum gauge involved decay measurements in order to assess the behaviour of the vane and to verify the theory, the amplitude sensing unit was originally constructed as shown in block diagram form in Fig. 2, the output being monitored on a c.r.o. Subsequently the output was fed to one of the units to be described later.

As the vane oscillates in simple harmonic motion and the rate of change of capacitance is constant, then the capacitance C between the vane and either plate will vary sinusoidally about a mean value C_o corresponding to the centre position of the vane so that

$$C = C_o + C_m \sin\omega t \quad \text{.........} \quad (14)$$

where C_m = maximum change of capacitance due to the displacement of the vane. C_m is proportional to the displacement θ and thus gives a measure of the amplitude of the oscillation. A method of measuring C_m is necessary and could take one of the forms:

(a) The capacitance could be used to modulate a carrier oscillator producing a frequency modulated signal with a deviation proportional to θ at a rate equal to the vane frequency ω, which is demodulated to give a voltage proportional to θ.

(b) The capacitance could be connected in series with a resistor and source of e.m.f. producing a variation in p.d. across the resistor which may be amplified giving a voltage proportional to θ.

Method (b) was rejected in view of the size of series resistor which would be required since C_m is of the order of 2pF, although this could be increased by means of a Miller integrator circuit. It was decided to use the first method, i.e. an f.m. carrier with demodulation.

Hurd and Corrin[2] have used such a system with the carrier oscillator operating at 10·7Mc/s. Since the change of capacitance is small and since for decay measurements on the gauge, the periods involved may be very long, it is possible for the drift in oscillator frequency to be comparable with the frequency deviation due to the gauge, so that as a result, the decay measurements would be unreliable. For the direct reading operation of the gauge the drift in oscillator frequency is of less importance, but since decay measurements were important, a stable oscillator was essential. Three possible methods of producing the carrier were considered:

(a) A stable low-frequency oscillator of the inductance-capacitance type with the vane gauge capacitance forming part of the oscillator tuned circuit.

(b) A crystal-controlled oscillator with an Armstrong phase modulator.

(c) A frequency-stabilized f.m. system with variable frequency as proposed by Ruston[3].

Method (b) was rejected since the modulating frequency would be too low (about 2c/s) and the system would be unsuitable[4]. Method (c) was thought to be unnnecessarily complex in view of the simplicity of (a) which was therefore chosen. Fig. 2 shows the block diagram of the sensing system the output being indicated by a voltmeter or c.r.o., although this monitoring point was moved later, as will be seen. The detailed circuit diagram is shown in Fig. 3.

The oscillator shown in Fig. 3 was originally described by Clapp[5] and Gouriet[6] and is essentially an inductance-capacitance version of the crystal oscillator. By making

L_T very large, C_T small, and both C_1 and C_2 large, the effects of drift in stray capacitances and valve parameters are minimized. The circuit may be considered as a Colpitts oscillator in which the tuning inductance is the effective combination of L_T and C_T in series and may easily be analysed from a generalized oscillator circuit[7] giving the conditions of oscillation as

$$\omega_0^2 = \frac{1}{L_T C_2} \quad [1 + (R/r_a) + (C_2/C_1) + (C_2/C_T)]$$

$$= \frac{g_m + 1/r_a + C_1/r_a C_T}{C_1(C_2 R + L_T/r_a)} \quad \ldots\ldots\ldots (15)$$

where ω_0 = oscillator frequency,
R = resistance of coil of inductance L_T
g_m = mutual conductance of valve.
r_a = anode slope resistance of valve.

A frequency of about 200kc/s was chosen for the carrier in view of the requirements of frequency stability and the operation of the demodulator, the output being taken from the cathode to minimize loading of the oscillator. In order to avoid discriminator alignment a pulse counter discriminator was used as the demodulator, the circuit including the limiter being an almost exact reproduction of one developed by Scroggie[8].

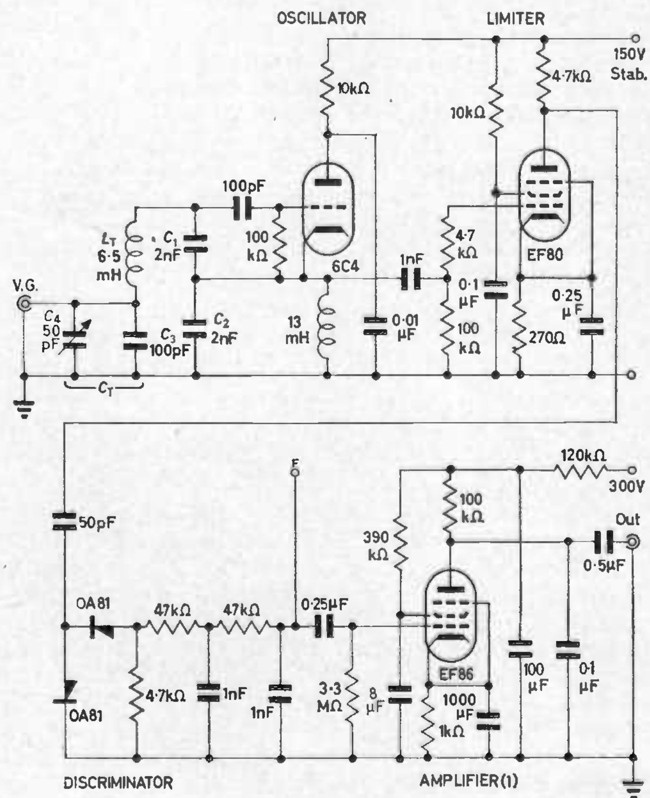
The deviation obtained in the oscillator frequency will be

$$\Delta f = (df/dC) \, . \, C_m = \frac{\omega_0 C'}{4\pi C_T^2} \, . \, C_m \quad \ldots\ldots (16)$$

where C' is the equivalent of C_1, C_2 and C_T in series, and C_T includes the gauge and lead capacitance. For the system used the maximum value of Δf is about 1 850c/s. The corresponding output from the discriminator should be about 30mV peak, according to Scroggie[8] and this was found to be so. This voltage is amplified by a standard pentode amplifier having a gain of about 100 and the output was observed on a c.r.o. in the early stages. Due to the fact that the amplifier had to operate at 2c/s, long *CR* coupling time-constants were necessary. These resulted in objectionable pick-up and amplification of the 50c/s mains frequency which was reduced by connecting large shunt capacitors throughout the circuit.

It should be mentioned that the original design of the unit was based on the use of transistors but in view of the large drive voltage required by the gauge at higher pressures it was realized that some valves would be required in the drive unit, since at 2c/s transformers would be very bulky. To avoid using a hybrid valve-transistor construction it was decided to use valves exclusively. By using high-voltage transistors and restricting the upper limit of pressure, the entire control system could be transistorized.

Fig. 3. Circuit of amplitude sensing unit



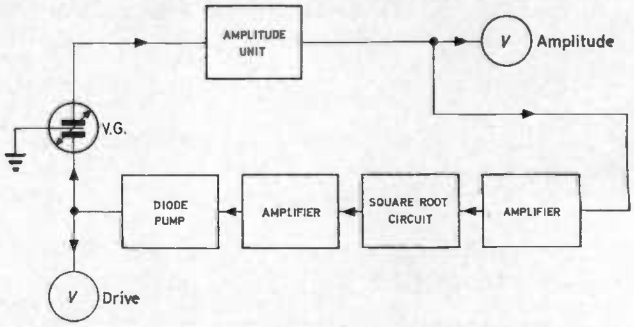


Fig. 4. Feedback system

The Feedback and Drive Circuit

It was desirable that the gauge and control unit should form an oscillatory system which would be self-starting and self-maintaining. It was intended that the amplitude of the vane should be adjusted manually by controlling the gain of an amplifier in the feedback loop. Pressure measurement was then to be carried out by setting the amplitude to a specified value and measuring the drive voltage which is proportional to the square root of pressure. This would have been satisfactory had the vane oscillated at a frequency of a few hundred cycles per second. Due to the very low vane frequency, i.e. 2c/s, the time-constants involved in the control system were excessively long and the idea had to be rejected in favour of an automatic system. In designing the system no account was taken of transient operation since it was assumed that the pressure being measured would change very slowly when change did occur.

The complete feedback loop is shown as a block diagram in Fig. 4 and in this form was intended for manual adjustment. Let k_1, k_2, k_3 etc. be the appropriate transfer constants for the various stages in the loop. If the sensing capacitance changes by ΔC due to movement of the vane, then the frequency deviation in the oscillator will be $\Delta f = k_1 \, . \, \Delta C$.

The output from the sensing unit is $V_1 = k_2 \Delta f$ which is fed into the drive unit. A square-root circuit gives an output of $V_2^2 = k_3 V_1$ and is necessary to make the system linear. The voltage V_2 is applied to the driving capacitance of the gauge producing a torque $T_D = k_4 V_2^2$. The deflexion is $\theta = k_5 T$ producing a change in capacitance $\Delta C = k_6 \theta$. The condition for oscillation is thus

$$k_1k_2k_3k_4k_5k_6 = 1 \quad \text{.............} \quad (17)$$

which is met only if the square-root circuit is included. A relaxation oscillation is possible without this circuit in the loop but it was thought preferable to make the system linear. An alternative method is to shape the fixed capacitor plates of the gauge so that $dC/d\theta \propto 1/\sqrt{C}$ and equation (2) becomes

$$z = q(1 + 1{\cdot}815\, l\, \theta/s)^{2/3} \quad \text{..........} \quad (18)$$

but since amplitude measurements were required this idea was rejected. The square-root circuit consists of two diodes connected back-to-back in series with a resistor, the output being taken across the diodes. The output is approximately proportional to the square root of the input voltage. More complex circuits are possible[9] but were not thought to be necessary. A large loss of voltage occurs in the circuit which must be compensated for by the addition of an amplifier stage preceding the square root circuit.

The output from the square-root circuit is fed to the drive amplifier via a control potentiometer RV_1 so that the loop gain may be adjusted. The amplifier drives a pump-diode rectifier circuit which provides uni-directional pulses to the gauge. The maximum peak pulse voltage is of the order of 150V which, as already mentioned, was the reason for using valves instead of transistors. The uni-directional pulse is essential since the deflecting torque does not reverse direction if the driving voltage is reversed. Furthermore the driving pulse must be applied while $dC/d\theta$ is positive since the capacitance tends to increase when a voltage is applied. Because of this, it is necessary to ensure that the correct phase relationship holds in the feedback loop.

The system so far described operated satisfactorily for initial experiments but the time delay observed when RV_1 was adjusted was excessive and it was impossible to use the system as intended. For this reason an automatic amplitude control circuit was added. The circuit diagram of the feedback unit is shown in Fig. 5 and includes the modification to the drive amplifier necessitated by adding the amplitude control and a starting trigger.

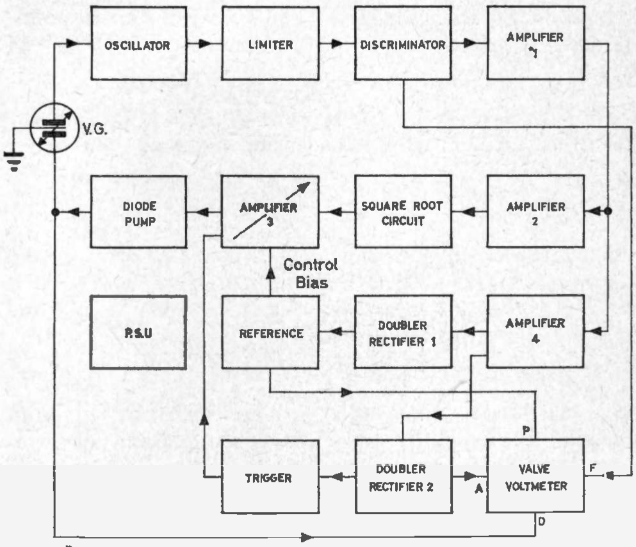


Fig. 6. Arrangement of complete control system

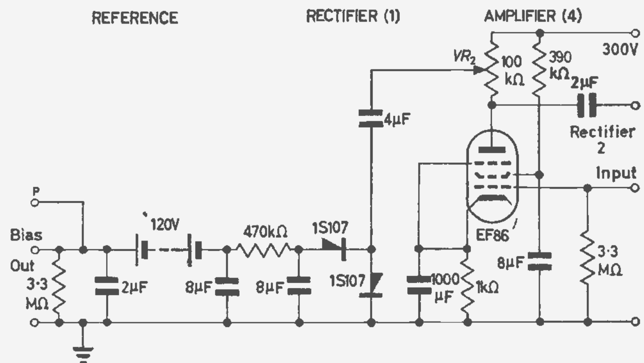


Fig. 7. Amplitude control unit

The Amplitude Control Circuit

The addition of this circuit effectively converts the system from open-loop to closed-loop control. Fig. 6 shows a block diagram of the entire gauge control system from which it will be seen that the output from the sensing unit is fed via an amplifier (Amplifier 4) to a voltage-doubler rectifier which produces a negative direct voltage. This voltage is compared with a reference voltage and the negative difference voltage is used to control the gain of the drive amplifier, which uses a variable-mu valve.

Originally the reference circuit consisted of a cold-cathode tube together with a resistance comparator network, but this was found to be ineffective since the rectifier became reverse biased. It was therefore replaced by a 120V battery in series with the rectifier output as shown in Fig. 7. The output from Amplifier 4 is adjustable by means of RV_2, the correct setting being such that at the highest pressure, the amplifier output is about 120V and zero control bias is applied to the drive amplifier. Thus the latter operates at maximum gain at the highest pressure. By means of both RV_1 and RV_2 the system may be adjusted at its highest and lowest pressures and will work within that range.

The vane amplitude will have its greatest value at the lowest pressure P_1 and will decrease to a value $\theta\,(1 - x)$ at the highest pressure P_2 due to the

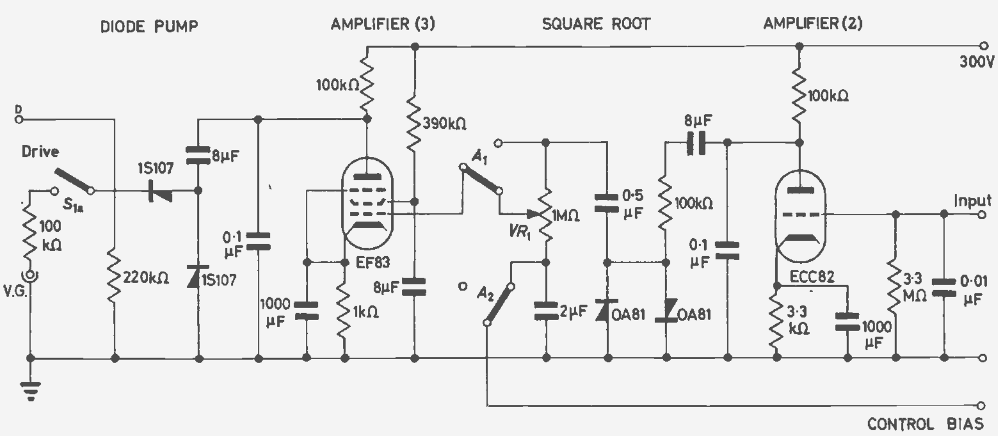


Fig. 5. Feedback and drive unit

increase in gas damping. If V_1 and V_2 are the corresponding drive voltages then

$$\theta = k_4 k_5' V_1^2 \text{ and } \theta(1-x) = k_4 k_5 V_2^2$$

where $k_5 \propto 1/(P_2+P_o)$ and $k' \propto 1/(P_1+P_o)$, P_o being the pressure equivalent of the suspension damping, whence

$$1 - x = \frac{P_1 + P_o}{P_2 + P_o} \cdot (V_2^2/V_1^2) \qquad (19)$$

When the pressure is P_1 the output from the sensing unit is v but this falls to $v\,(1-x)$ at a pressure P_2 so that if the gain of the drive amplifier is A_1 and A_2 at low and high pressures respectively then

$$V_1 \propto A_1 \sqrt{v} \text{ and } V_2 \propto A_2 \sqrt{[v(1-x)]}$$

the square-root being taken by the square-root circuit preceding the amplifier. Substituting in equation (19) gives

$$A_1/A_2 = \sqrt{\frac{P_1 + P_o}{P_2 + P_o}} \qquad (20)$$

If the control voltage applied to the grid of the variable-mu valve is V_g which is zero for a gain of A_2, then

$$A_1 = A_2 \exp(bV_g) \qquad (21)$$

where b is a constant applicable to the valve used. Substituting for A_1/A_2 in equation (20) from equation (21) gives

$$V_g = (1/2b) \ln \frac{P_1 + P_o}{P_2 + P_o} \qquad (22)$$

This voltage is derived as the algebraic sum of the reference voltage V_R and the output from the doubler-rectifier and amplifier which will be Mv where M is the gain of the circuit. Thus $V_g = V_R - Mv$ and if $V_R = Mv(1+x)$ at pressure P_1 it follows that $V_g = -Mvx$. Since V_g is fixed by equation (22) and v is fixed by the sensing circuit for a given vane amplitude, then the proportional decrease in vane amplitude will be inversely proportional to M, which should therefore be as large as possible if x is to be small. Typical values in the system were $b = 0{\cdot}1$, $v = 1{\cdot}5$V, $M = 200$, $P_2 = 10^{-4}$, $P_1 = 10^{-6}$ and $P_o = 3{\cdot}8 \times 10^{-6}$ giving a value of x of about 5 per cent, V_g being about -15V at the lowest pressure.

Since the variation of V_g with pressure covers a wider range than the drive voltage, and since this variation is logarithmic, it was decided to use V_g as a measure of pressure.

Starting Trigger, Valve-Voltmeter

The time taken for the vane to reach full amplitude from rest was inconveniently long so that it was necessary to include a rapid starting device. Rapid build-up of vane amplitude is achieved by means of a relay which removes the control bias and by-passes RV_1 so that the drive amplifier operates at maximum gain. When the amplitude reaches a predetermined level, the relay is released via a Schmitt trigger circuit[10], the control bias is applied and RV_1 is operative. The amplitude level is derived from Amplifier 4 via a doubler-rectifier, the release level being set by RV_3. Circuit details are shown in Fig. 8. The relay system also operates when the pressure is too high, the system tending to hunt.

A d.c. valve-voltmeter is incorporated in the control unit and is shown in detail in Fig. 9. Reference to Fig. 6 shows the application of the meter to various parts of the system. Since the voltmeter is of the balanced bridge type, the first position Z of the function switch enables the zero to be adjusted. The second position F connects the meter to read the d.c. level at the output of the discriminator and is used for setting the carrier frequency. This is done by adjusting the variable capacitor C_4 in the oscillator circuit and is only necessary if the lead from the oscillator to the gauge is altered in length.

In the A position the valve-voltmeter reads a voltage which is proportional to the vane amplitude, the most convenient point being the input to RV_3. For amplitude decay measurements the drive is removed by means of a switch S_{1a}-S_{1b} which also removes the h.t. supply to the relay circuit, so preventing the drive from becoming excessive since it would be possible at very low pressures for the vane to be maintained via the capacitance of the switch, S_{1a}.

In the F position the valve-voltmeter reads the control bias V_g which has been shown to be a function of pressure and may therefore be calibrated directly in terms of pressure against some suitable standard. No range switch-

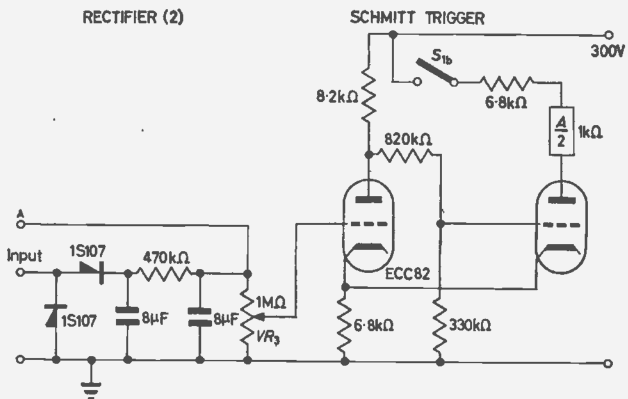


Fig. 8. Starting trigger

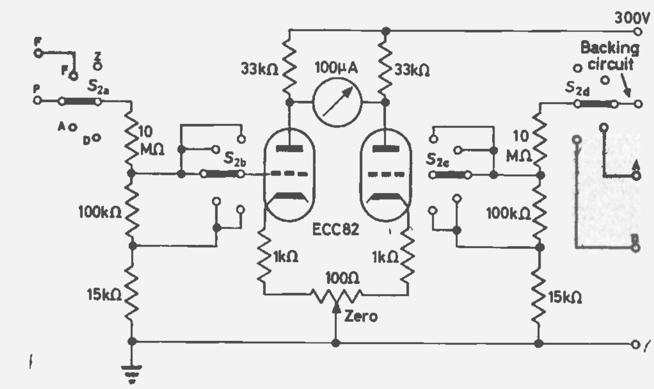
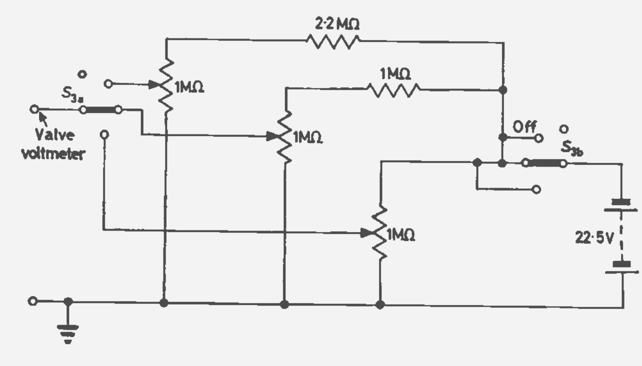


Fig. 9. Valve-voltmeter

Fig. 10. Backing-off circuit



ing, other than the one fixed range shown, was used in this position, but it would be necessary in a final design. Such range switching could be arranged in conjunction with the backing-off circuit shown in Fig. 10 so that several pressure scales could be incorporated. The *D* position of the function switch measures the drive voltage as a pulse, the meter following the pulse with time. No attempt was made to transform this into a steady reading since it was not used to indicate pressure.

Experimental Results

Most of the experimental work was concerned with decrement measurements on the gauge itself, and with the calibration system and will be described elsewhere[11]. The gases with which the gauge was used ranged from helium (mass 4) to hexafluoropropylene (mass 150), although only the results for air (mass 29) which are typical are shown here. Fig. 11 shows the variation of control voltage V_g with pressure over a range of about 100 to 1, the pressure in this case being measured by means of an ionization gauge. Since the pressure scale is logarithmic the graph is approximately a straight line over most of the range. The departure from linearity, observed also with other gases, may well be due to a variation in the constant b in equation (21) with grid voltage, being a characteristic of the variable-mu valve. Considerable departure from linearity, observed when the controls RV_1 and RV_2 were incorrectly adjusted. In fact, in the case of Fig. 12 it could not be claimed that these controls had been set at their optimum positions.

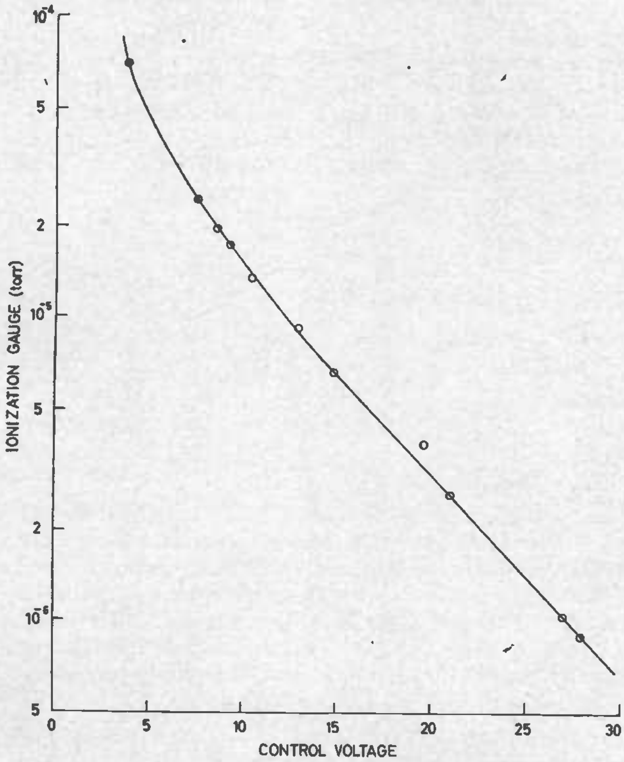
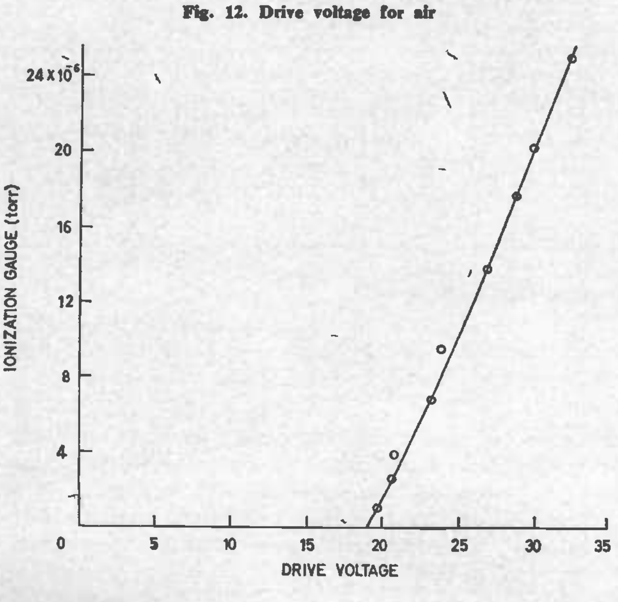


Fig. 11. Control voltage for air with log pressure

Fig. 12. Drive voltage for air



The graph of drive voltage against ion gauge reading is shown in Fig. 12, the pressure scale in this case being linear. At zero pressure a drive voltage of 19V peak is necessary to maintain the vane due to the suspension stiffness. The graph appears to have the general shape expected since drive voltage is proportional to square root of pressure although numerical values differ from those expected. This discrepancy is possibly due to the inertia of the moving-coil meter used in the valve-voltmeter, so that the meter was unable to follow the voltage, the error becoming greater as the voltage increased.

The system shows a tendency to hunt when subjected to a sudden large change in pressure or when the pressure is too high for the amplitude to be maintained, i.e. outside the range of the instrument. Since the gauge was developed for use in a vacuum plant where the pressure may be expected to be reasonably steady and within the range of the instrument, the hunting was not investigated further.

The use of two controlled stages in the drive amplifier would increase the pressure range of the system to about 10^4 to 1 so that the lower limit of pressure will be a function of the molecular gauge itself and not of the control system as at present. It is possible of course to adjust RV_1 and RV_2 to enable the lower limit to be reached at the expense of a lower maximum pressure, the range being about 100 to 1. A final version is being developed which will contain two controlled drive stages together with multi-range valve-voltmeter and backing-off circuit but due to the stiffness of the vane suspension, the lower limit of pressure may well be no better than 10^{-7}.

Acknowledgments

Thanks are due to Dr. J. H. Leck, for his assistance and encouragement in connexion with this work, to Prof. J. D. Craggs for provision of facilities at Liverpool University and to the Liverpool Education Authority for leave of absence and financial assistance.

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Cryostat Temperature Measurement from 0·1°K to 20°K Using a Wein Bridge Oscillator

By P. R. Adby*, B.Sc.

The oscillators described are a modification of the conventional Wein type in which the power dissipated in the resistors of the phase shift network is very small. One of the resistors is temperature dependent and may be placed in a cryostat at temperatures as low as 0·1°K without significantly affecting the thermal conditions. The oscillator frequency is related to temperature and may be measured by means of a counter.

(Voir page 830 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 837)

THE use of carbon composition resistors for the measurement of very low temperature has been an established technique for some time. The relationship between resistance and temperature for certain Speer and Allen-Bradley carbon resistors has been investigated[1,2]. The measurement of resistance has normally been carried out by an a.c. Wheatstone bridge technique[3] with very low power dissipation in the resistance thermometer. The bridge method has the advantage of precision but is relatively slow in use since continuous manual balance is necessary to measure temperature.

An alternative method for measurement of resistance is possible by incorporating the resistance into one arm of the phase shift network of a Wein bridge oscillator. The relationship between resistance and temperature is rather complex, therefore the additional one between resistance and frequency is no real disadvantage in this application. Two transistorized oscillators have been designed to work with very low power in the phase shift network and are, therefore, suitable for very low temperature measurement. A valve circuit has been described[4] using this principle.

Oscillator Circuit for 0·1°K to 4°K

The circuit diagram of the oscillator is given in Fig. 1. The basic oscillator is similar to the Mullard circuit[5] in which a current connected phase shift network would normally be connected between the collector of VT_8 and the base of VT_7. The output is stabilized to about 1 to 1·5V r.m.s. by negative feedback via the thermistor.

In the modified circuit the output at the collector of VT_8 is attenuated by a factor of 1 000 and applied to a voltage connected phase shift network. The thermometer resistor is connected at the input terminal and appears across part of the Wein network. The voltage across this resistor varies from approximately 1mV at $R = 220\Omega$ to 2mV at high resistance. The power dissipated in R therefore varies from about $0{\cdot}5 \times 10^{-2}\mu$W at higher temperatures to about $0{\cdot}5 \times 10^{-3}\mu$W at lower temperatures. The oscillator loop is completed by the wide band amplifier consisting of transistors VT_1 to VT_6. The voltage gain to the collector of VT_5 is set to about 1 000 by means of the preset potentiometer and the emitter-follower VT_6 gives a current output to the base of VT_7 via a 1kΩ series resistor.

The oscillator frequency is continuously monitored by a digital counter connected to the collector of VT_8. Various thermometer resistors may be examined easily by means of a switch selecting one of several input sockets. A stabilized power supply of 30V at 50mA is required.

The frequency of oscillation is given by:

$$f = \frac{1}{2\pi\sqrt{(C_1C_2R_1R_2)}} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

where $C_1 = C_2 = 0{\cdot}022\mu$F

$R_1 = 270 + 43 = 313\Omega$

R_2 is the temperature dependent resistor in parallel with 22kΩ.

Therefore:

$$f = 406/\sqrt{R_2}\,\text{kc/s} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

Oscillator Performance

The formula given in equation (2) was used to construct Fig. 2 showing the relation between oscillator frequency and the thermometer resistance. The curve obtained from the prototype circuit shows reasonable agreement between calculated and observed results over the resistance range 200Ω to 10kΩ. The error at high frequencies appears to be caused by bandwidth limitations and at low frequencies by partial failure of the amplitude stabilization.

The frequency stability of the oscillator was tested at various values of resistance for ambient temperature and power supply variations. The results are summarized in Table 1.

TABLE 1
Frequency Stability

THERMOMETER RESISTANCE (Ω)	AMBIENT TEMPERATURE EFFECTS	H. T. VOLTAGE EFFECTS
390	8 in 10^4/°C	1 in 10^3/Volt
2·2k	4 in 10^4/°C	1 in 10^3/Volt
6·8k	3 in 10^4/°C	1 in 10^3/Volt

Thermometer Performance

Fig. 3 shows typical curves of resistance against temperature for some of the resistors suitable for use with this oscillator. More complete information is given elsewhere[1,2].

It can be seen that the relationship between resistance and temperature is somewhat complex. Also variations between resistors of one type are not negligible. It is therefore normal to calibrate each individual resistor-cryostat system and to periodically recalibrate to check the long-term stability.

* *University of Sussex.*

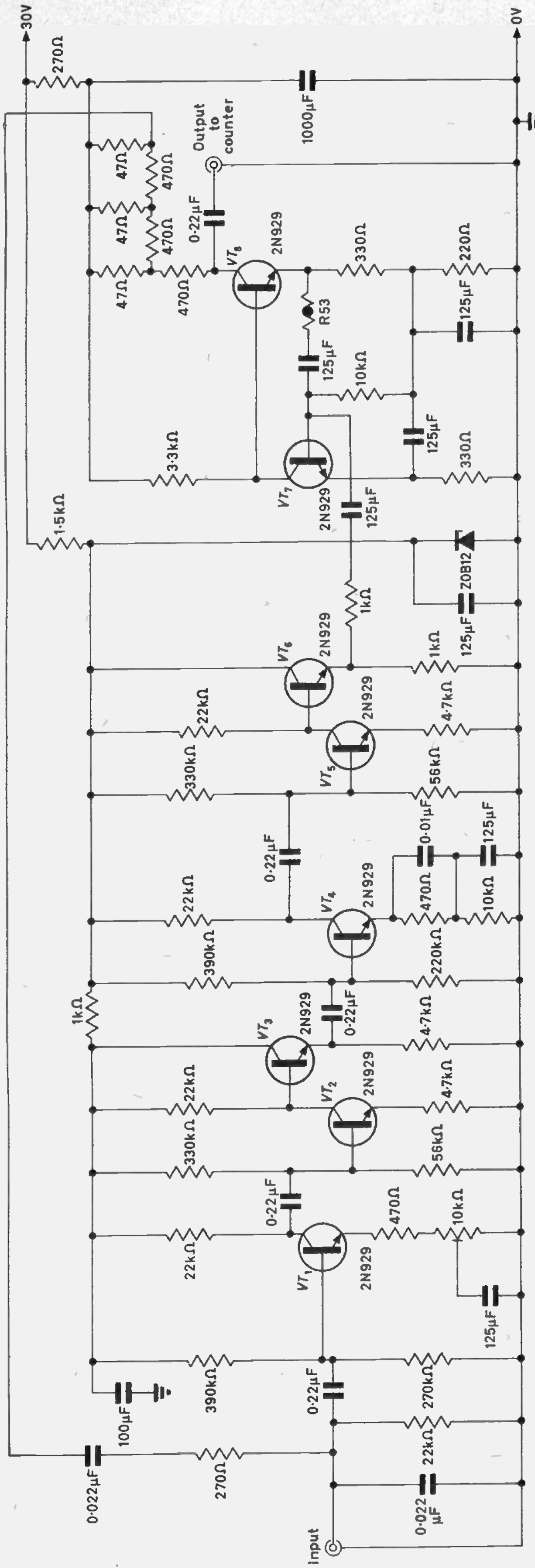


Fig. 1. 4kc/s to 25kc/s oscillator for 0·1°K to 4°K

The resistor chosen for the temperature range 0·1°K to 4°K was the Speer, grade 1 002, 220Ω nominal, ½W; since the range of resistance obtained is about 400Ω to 6kΩ. This range is within the working limits of the oscillator. For other temperature ranges a suitable resistor would be chosen in a similar way.

Fig. 4 was obtained for the resistor chosen showing the calibration curve relating frequency and temperature. The accuracy of the thermometer may be assessed from the calibration curve in conjunction with the performance

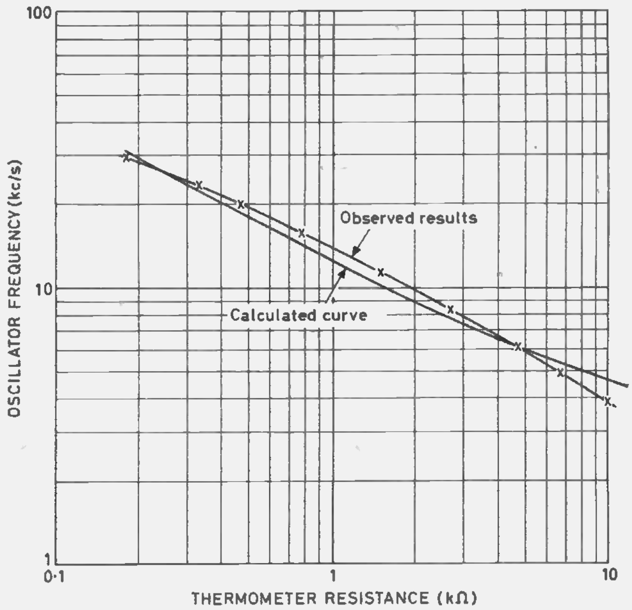
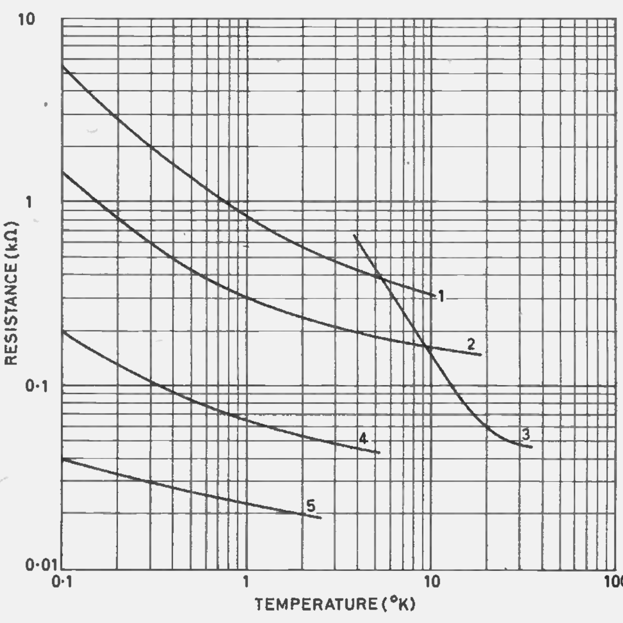


Fig. 2. Relation between oscillator frequency and thermometer resistance

Fig. 3. Typical resistance thermometer characteristics for certain Speer and Allen Bradley resistors

(1) Speer 220Ω 0·5W grade 1002
(2) Speer 100Ω 0·5W grade 1002
(3) Allen Bradley 47Ω 0·125W
(4) Speer 27Ω 0·5W grade 1002
(5) Speer 10Ω 0·5W grade 1025



data given in Table 1. Under normal laboratory conditions the accuracy obtained should be from about 1 per cent at 4°K to about 0·5 per cent at 0·1°K.

Oscillator Circuit for 4°K to 20°K

The accuracy of the previous circuit was obtained without special precautions for very high frequency stability. It was therefore felt that improvement to the design could be obtained. A requirement for a second circuit for 4°K to 20°K using the Allen-Bradley 47Ω ½W resistor enabled some improvements to be incorporated.

Since the lowest temperature to be measured was 4°K more power could be dissipated in the thermometer resistor than in the previous circuit and a dissipation of 0·01μW at 4°K was chosen. The curve for the Allen-Bradley resistor in Fig. 3 shows that the resistance range expected is from 60Ω to 600Ω. A frequency range of about 200kc/s

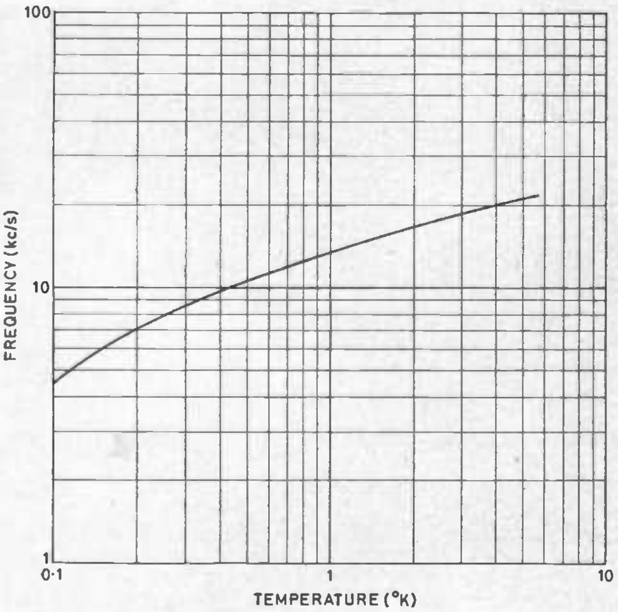
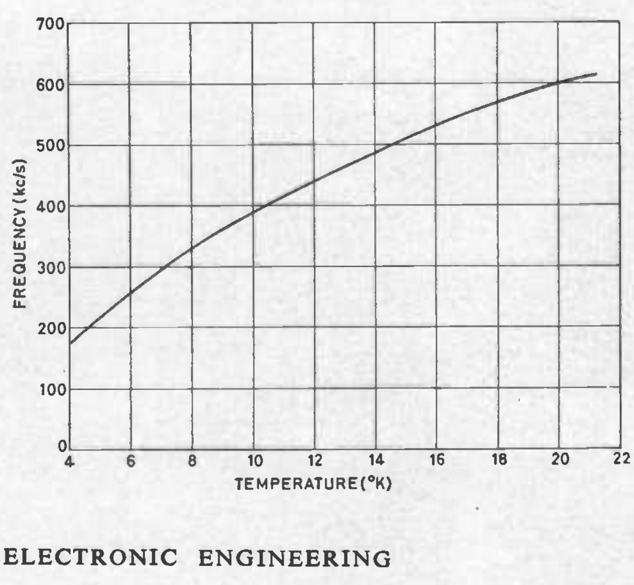


Fig. 4. Calibration curve for 0·1°K to 4°K
(Speer 220Ω 0·5W grade 1002)

Fig. 6. Calibration curve for 4°K to 20°K
(Allen Bradley 47Ω 0·5W)



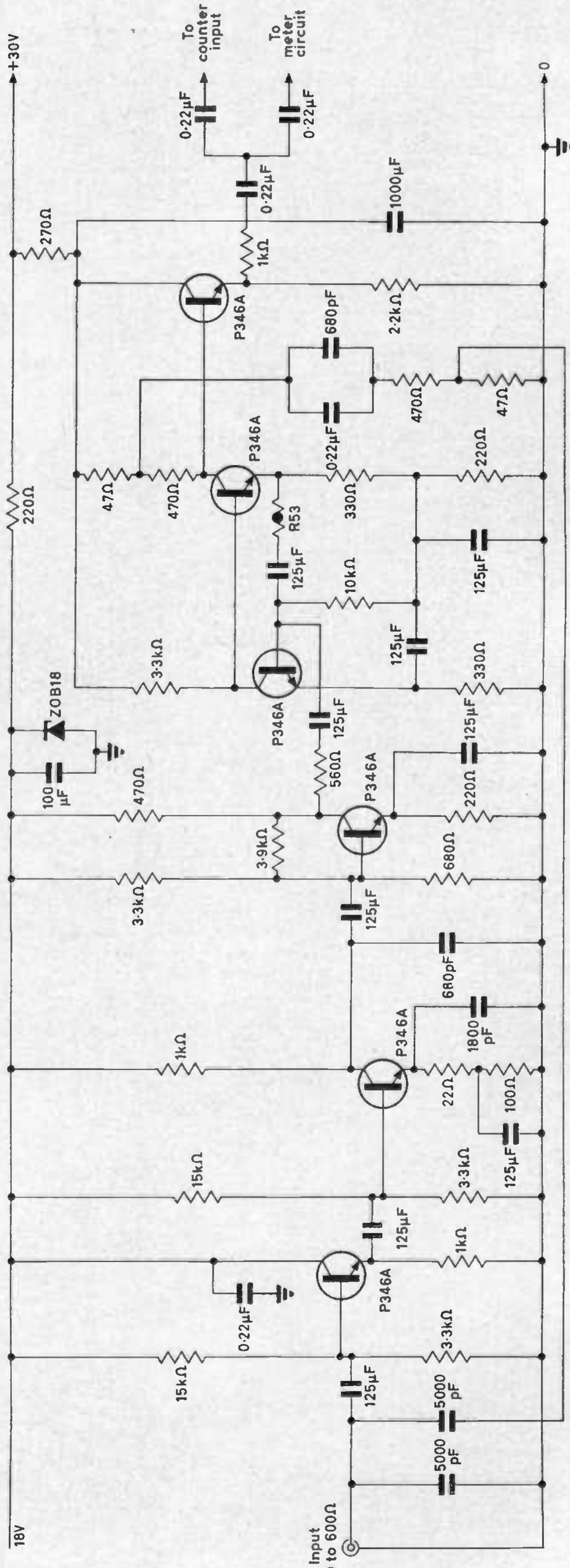


Fig. 5. 200kc/s to 700kc/s oscillator for 4°K to 20°K

TABLE 2
Frequency Stability

THERMOMETER RESISTANCE (Ω)	AMBIENT TEMPERATURE EFFECTS	H. T. VOLTAGE EFFECTS
56	1 in 10^5/°C	1 in 10^4/Volt
180	3 in 10^5/°C	3 in 10^4/Volt
560	1 in $10/^4$°C	1 in 10^3/Volt

to 700kc/s was therefore selected giving improved resolution in a short time.

From these considerations the oscillator was redesigned to use an improved amplifier with, a gain of 100, a bandwidth of several megacycles per second, and an input

assessed from the performance data given in Table 2 in conjunction with graph 4. Under normal laboratory conditions the accuracy obtained should be from about 0·01 per cent at 20°K to 0·1 per cent at 5°K. The circuit has been optimized at 20°K where resistance changes least with temperature in order to improve accuracy at the expense of accuracy at 4°K.

Conclusion

The main advantage of the oscillator technique is that measurement of frequency may be carried out automatically over a period of 0·1sec or 1sec using a low-speed (1Mc/s) counter. Permanent results are easily obtained if required by adding a printer. In contrast the a.c. bridge is rather slow and requires continuous attention. A.C. bridges must be used however where high

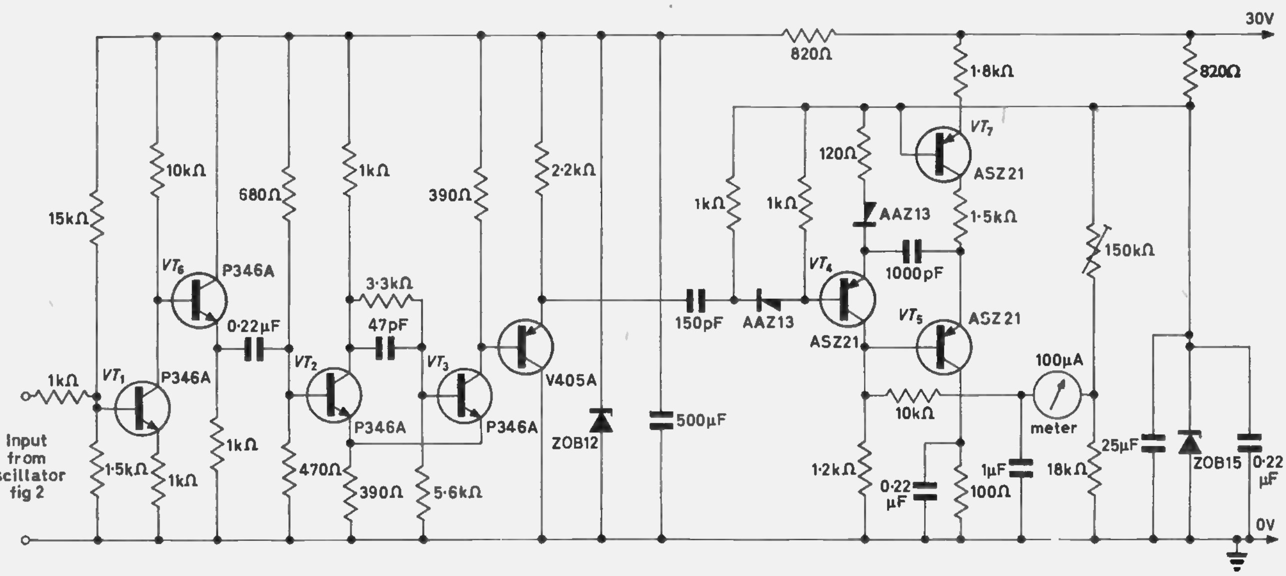


Fig. 7. Meter circuit for 200kc/s to 700kc/s

impedance of about 2kΩ. The circuit diagram of the oscillator is given in Fig. 5 and a calibration curve is given in Fig. 6.

The performance of the oscillator is summarized in Table 2.

The improved performance has been obtained by attention to the amplifier input impedance which appears in shunt across the thermometer resistance, and by improving the amplifier bandwidth in comparison with the oscillation frequency. Loading of the output is, under normal conditions, constant but an emitter-follower has been added.

A further facility added to this thermometer is a meter readout. The circuit diagram is shown in Fig. 7. The output from the oscillator is amplified by transistor VT_1, and squared by the Schmitt trigger VT_2 and VT_3. The output from the squarer triggers an emitter coupled monostable multivibrator which gives a fixed output pulse length of just under 1μsec. The mean d.c. level at the collector of VT_5 is measured by the meter which is calibrated directly in degrees kelvin.

The meter scale is slightly non-linear over the range 4°K to 20°K and the accuracy of 5 per cent is limited by the pulse width stability and reading accuracy. The accuracy of measurement using the counter may be

accuracy is required (e.g. 1 in 10^5 for some specific heat measurements).

It is hoped that further work on circuits of this type will lead to:

(1) A multi-range thermometer with meter read-out accurate to about 1 per cent over the range 0·05°K to 150°K.

(2) A transistorized precision thermometer accurate to 1 in 10^5.

(3) A circuit with cryostat temperature stabilization based on frequency comparison.

Acknowledgment

The author would like to thank Dr. M. G. Richards for valuable discussions and for assessment of the thermometer at low temperatures.

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Electronic Mode-Control of Operational Amplifiers

By A. D. Bond*, B.Sc., Ph.D., and P. L. Neely*, B.Sc.

This article discusses the design of the equipment necessary to extend existing analogue computer facilities to include the repetitive/iterative mode of operation.

The basis of the design is a six diode bridge in a feedback switching arrangement.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 837)

THE modern repetitive analogue computer, which is capable of storage and logical decision, is essentially a hybrid machine in that its mode of operation is controlled by a digital logic system[1]. The basic link between the digital and analogue parts of the system is the electronic switch which operates the resetting of integrators, mode switching of amplifiers and the gating of signal paths.

High speed semiconductor switches enable iterative computations[2] to be carried out at rates varying from seconds per cycle to kilocycles per second[3,4,5]. This versatility opens up possibilities for the use of a special purpose iterative analogue computer as an on-line controller for the dynamic optimization of real time systems.

The article discusses the design of the equipment necessary to extend existing analogue computer facilities to include the repetitive/iterative mode of operation. High speed hill-climbing adaptive loops can thus be implemented to solve boundary value problems. (See Appendix (A)).

Mode Switching of Operational Amplifiers

Fig. 1 shows the conventional relay implementation of a mode switched amplifier. The various modes of operation are listed in Table 1 against the corresponding combinations of contacts A and B and feedback element Z_0. The problem considered in this section is that of replacing relay contacts with high speed solid state switches and choosing the best configurations for economy coupled with sufficient accuracy.

Table 2 compares the more important parameters involved in switch design for three different types of electronic switch, considering all three to have similar switching speeds.

Diode bridge switches are the cheapest, can handle larger signal levels, and have better switching spike characteristics[6]. The principal disadvantage is the requirement to provide anti-phase control pulses. However, the six diode bridge[7] eliminates the necessity for absolutely symmetrical pulses to the switch, and a further modification[2] (Fig. 2) reduces the leakage errors at the expense of two more diodes.

Substitution of solid state switches for contacts A and B involves consideration of the problem of direct switching and feedback switching.

Direct Switching

An electronic switch, replacing contacts A in Fig. 1, will tend to introduce severe errors in the output due to offset, unbalance, etc., since the switching takes place at the amplifier virtual earth. These errors can be reduced by placing the switch on the input side of resistor R_1. (See Appendix (B)). However, any attenuation through the switch will now be significant and, more serious, the control pulses must be greater than the maximum signal level otherwise limiting in the 'on' condition and breakthrough in the 'off' condition will occur. For a ±100V computer operating with switching times of a few microseconds, this condition will impose stringent requirements on the control pulse generator.

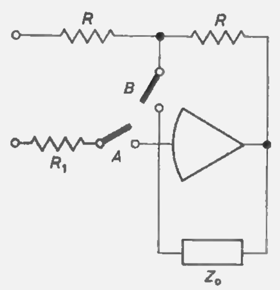


Fig. 1. Mode-switched amplifier

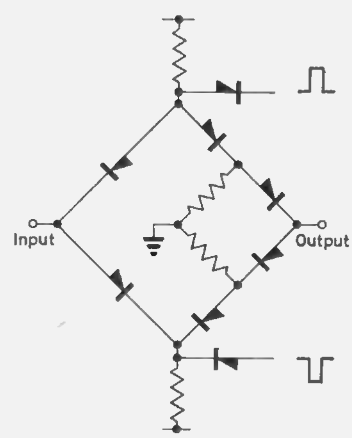


Fig. 2. Modified six diode bridge switch

TABLE 1

	$Z_0=C_0$			$Z_0=R_0$	
Contacts A	Open	Open	Closed	Open	Closed
Contacts B	Closed	Open	Open	Closed	Open
Amplifier Mode	Track or Reset	Hold	Integrate or Compute	Summer 1	Summer 2
				Change over Switch	

TABLE 2

	FORM OF CONTROL PULSE	FORWARD RESISTANCE	BACK RESISTANCE OR LEAKAGE	OFFSET VOLTAGE IN 'ON' STATE
Diode Bridge	Antiphase Symmetrical	$<50\Omega$	<10nA	Balanced to Zero
Transistor	Single	$<50\Omega$	<10nA	<5mV
F.E.T.	Single	$100\Omega \to 5k\Omega$	$>10^9\Omega$	Very small

* *The Queen's University of Belfast.*

For virtual earth switching, diode limiters at the switch input will prevent signal breakthrough in the 'off' condition.[8]

However, direct switching can be avoided in the reset and compute modes by using a buffer or driver amplifier in series with the feedback switch B.

Feedback Switching

A mode switched amplifier used exclusively in reset and compute becomes the track-hold unit of an iterative analogue computer when there is no integrator input present. This unit performs the function of sample and hold and provides the analogue computer with a storage element.

If B is a simple electronic switch, then the reset/track sample time-constant is given by $(R/2)\ C_o$. This can be excessive; for example, using typical components with a compute time of 1msec the reset time-constant is 50μsec. Consequently a driver amplifier, acting as an impedance transformer in series with the switch, is required to reduce the reset time-constant[2,9], typically by a factor of 10^3.

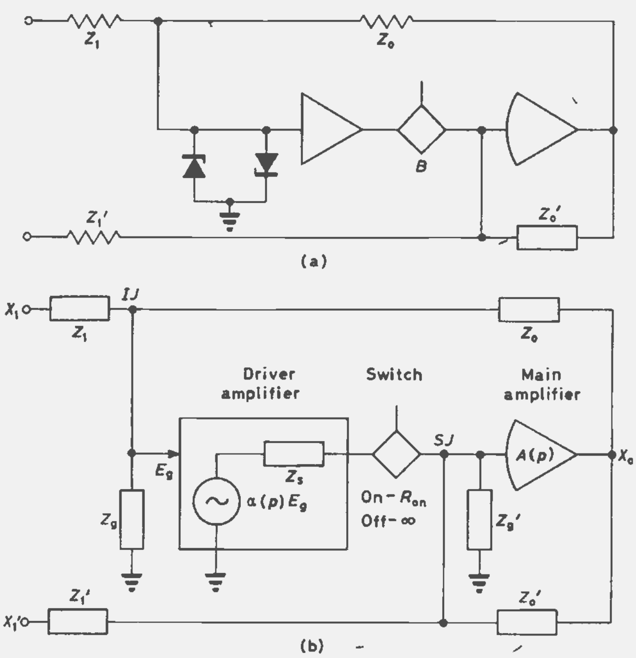


Fig. 3 (a). Electronic mode-switched amplifier
(b). Equivalent circuit of electronic mode-switched amplifier

With reference to Fig. 3(a) the function of the driver amplifier is to effectively short-circuit the input and feedback currents through Z_1' and Z_0' respectively, ensuring that the computing components become virtually inoperative relative to the track/reset components.

Thus a single switch B with a driver amplifier will operate the computing amplifier in the reset and compute modes. As shown subsequently, the error can be maintained between 0·1 per cent and 1 per cent over a wide range of frequencies, using simple, inexpensive equipment. If a system integrator is required to operate a reset-hold-compute cycle, then a direct switch A must also be used for the hold mode.

Feedback switching takes place between the initial-condition summing junction and the main amplifier summing junction and hence the switch is not required to handle peak signal variations. Again breakthrough, and also paralysis of the driver amplifier in the 'off' state, are prevented by diode limiters at the driver input (see Fig. 3(a)).

Unbalance and offset effects are reduced by feedback action. One disadvantage of feedback mode-switching is that the inclusion of a driver amplifier designed to have a very low output impedance introduces the possibility of instability[2], during the track period.

As is shown in the next section, the necessary bandwidth exceeds that required purely from computing considerations.

Electronic Mode Control Unit

As noted previously, the unit consists of:

(a) Computing Amplifier
(b) Electronic Switch
(c) Switch Driver Amplifier
(d) Control Pulse Generator

A unit comprising these four elements will operate from an arbitrary train of control pulses in the following modes.

Track-Hold	for a track/hold unit
Reset-Compute	for an integrator
Sum 1-Sum 2	for a changeover switch

Analysis[10] has shown that overall analogue computing accuracy is governed by the bandwidth of the computing amplifier. Thus design of the elements in a mode-control unit is based on the performance of the computing amplifier.

Loop Analysis

Consider a linear feedback loop represented by the signal flow graph of Fig. 4(a).

The overall output C is given by:

$$C = W_r \frac{A\beta}{1 - A\beta} \cdot r + W_d \cdot \frac{A\beta}{1 - A\beta} \cdot d$$

or:

$$C = W_r \frac{(-1)}{1 + 1/T} \cdot r + W_d \cdot \frac{(-1)}{1 + 1/T} \cdot d \quad \ldots . (1)$$

where C = overall output
r = reference input
d = disturbance referred to input
W_r = desired transfer function
W_d = disturbance transfer function
$T = -A\beta$ = return ratio[11]

Now the desired input/output relationship is:

$$C_d = -W_r \cdot r$$

Hence the error $\epsilon = C_d - C$ consists of two components:

(1) Difference between desired and actual transfer functions:

$$\epsilon_r = W_r\left(-1 + \frac{1}{1 + 1/T}\right) = W_r \cdot \frac{(-1)}{1 + T}$$

(2) Output due to an unwanted disturbance:

$$\epsilon_d = W_d \cdot \frac{(-1)}{1 + 1/T}$$

It is clear that $\left.\begin{array}{l}|\epsilon_r| \propto W_r\,(1/T) \\ |\epsilon_d| \propto W_d\end{array}\right\}$ for $T \gg 1$

At the same time to ensure stability of the loop, the Bode plot of T must have adequate gain and phase margins.

The equivalent circuit of a mode-control unit (Fig. 3(b)) has been analysed[8] and the result for the case when the switch is 'on' is shown in the form of a flow-graph in Fig. 4(b). A simple reduction leads to the diagram of Fig. 4(c) which is seen to be of the same form as Fig. 4(a). It is important to realize that during the track-reset period, when the switch is 'on', the computer signal X_1' represents a disturbance input which must be rejected.

Comparing Figs. 4(a) and 4(c)

$$T = -\frac{\alpha(p)\,Z_0'\,A_{on}\beta_{on}}{Z_s + R_{on}} \quad \ldots\ldots\ldots\ldots\ (2)$$

$$W_r = Z_0/Z_1$$

$$W_d = \frac{Z_s + R_{on}}{\alpha(p)\,Z_1'\,\beta_{on}}$$

where $\beta_{on} = (1 + (Z_0/Z_1) + (Z_0/Z_g)^{-1}$ and is the feedback factor for the track input X_1.

$$\beta_1 = \left(1 + (Z_0'/Z_1') + (Z_0'/Z_g') + \frac{Z_0'}{Z_s + R_{on}}\right)^{-1}$$

and is the feedback factor for the computer X_1'.

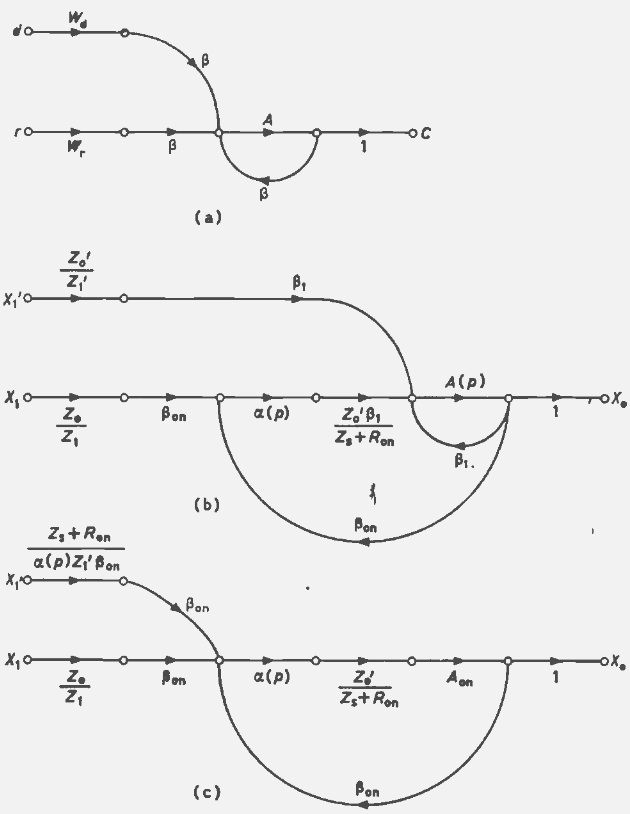
$$A_{on} = \frac{A(p)\,\beta_1}{1 - A(p)\,\beta_1}$$

Note: Z_g and Z_g' represent the input impedances of the driver and computing amplifiers respectively and should be large compared with Z_0', Z_1', Z_0, Z_1'.

An upper limit of the values of Z_0', Z_1', Z_0, Z_1 is set by the required speed of operation[2] while a lower limit is set partly by the maximum power available from the main amplifier but principally by the requirement that the series combination of driver amplifier source impedance Z_s and switch 'on' resistance R_{on} shall be very much smaller than the computing impedances Z_0', Z_1' and the track/reset impedances Z_0, Z_1. This latter condition, by increasing T and reducing W_d, serves to reduce both error terms ϵ_r and ϵ_d.

Fig. 4 (a). Flow graph of linear feedback loop

(b) and (c). Flow graphs of electronic mode-switched amplifier



Typical values gives the ratio $\beta_{on}/\beta_1 \simeq 100$ when $\beta_{on} = 0{\cdot}5$, indicating that the amplifier has a very small feedback factor and is virtually working open-loop.

Hence the frequency response of the return ratio T approaches the open-loop response of the main amplifier and the tracking error ϵ_r is maintained with the same order of accuracy as normal dynamic errors in computation. Although $\alpha(p)$ appears in both expressions for T and W_d, it does not affect the magnitude of either quantity at computing frequencies, since stable operation requires that the driver amplifier be wide band with unity voltage gain.

Stability

When the switch is 'on', the stability of the driver amplifier—computing amplifier loop is determined by examining the behaviour of the return ratio T as a function of frequency. Note again that for small values of $(Z_s + R_{on})$, $(< 50\Omega)$, the computing amplifier is virtually operating open-loop. Consequently, as shown below, the frequency variation of the return ratio T is approximately determined by the cascade connexion of the open loop computing amplifier and the unity gain driver amplifier.

$$T = -\frac{\alpha(p)\,Z_0'\,A_{on}\beta_{on}}{Z_s + R_{on}}$$

Now with $Z_0 = Z_1 = R \ll Z_g$,

β_{on} is independent of frequency.

Hence:

$$T \propto \frac{\alpha(p)\,Z_0'\,A_{on}}{Z_s + R_{on}}$$

And with $(Z_s + R_{on}) \ll Z_1', Z_g'$

$$\beta_1 \simeq \left(1 + \frac{Z_0'}{Z_s + R_{on}}\right)^{-1}$$

Put:

$$G(p) = \frac{Z_0'}{Z_s + R_{on}} \cdot A_{on}$$

where $G(p)$ is the transfer function of the computing amplifier with feedback Z_0' and input $(Z_s + R_{on})$.

When $(Z_s + R_{on}) \ll Z_0'$, $G(p)$ approximates to the open-loop transfer function of the main amplifier, $A_0/(1 + p\tau_1)$.

If $\alpha(p)$ is considered to have a second order frequency response characteristic, then since:

$$T = \alpha(p)\,G(p)$$

the bandwidth of $\alpha(p)$ must be greater than, or equal to, the gain-bandwidth product of the main computing amplifier, for stability of the loop. It is evident that if $\alpha(p) > 1$, the required bandwidth is increased.

The driver amplifier bandwidth required from computing accuracy and tracking accuracy considerations will be smaller than the figure arrived at from stability considerations for the following reasons:

(a) The driver amplifier only operates in the compute mode as a summing unit (changeover switch mode in Table 1).

(b) The tracking rate is limited by the ability of the mode-control unit to supply charging current to the integrator or hold capacitor[8].

Design and Performance of the Electronic Mode-Control Unit

As has been stated, the problem considered here is to provide existing computing amplifiers with electronic mode switching facilities. An overall accuracy figure of 1 per cent is considered sufficient since the units are intended to operate in an iterative fashion at frequencies of the order of 1kc/s. This high speed of operation will serve to minimize any tendency for errors to accumulate during an iterative computation.

The computer specifications which have to be met, can be stated as:

(a) Maximum signal level ±60V

(b) Maximum current output from the main amplifier 4mA.

(c) Maximum compute/reset frequency of 1kc/s

(d) Maximum sample or reset time of 100μsec

(e) Main amplifier open-loop gain-bandwidth product of 4·5Mc/s

(f) Tracking bandwidth of 10kc/s

(g) Overall accuracy of 1 per cent.

Switch

The choice of switch has been indicated already. In this case the six diode bridge switch is chosen for economy linked with speed and sufficient accuracy.

The control pulse generator is a Schmitt trigger circuit[2], both elements being capable of delivering 4mA, which is the rate limit introduced by the main amplifier.

Driver Amplifier

In order to meet the criteria for maximizing the return ratio T, the following specifications are laid down for the driver amplifier.

(a) Input impedance $\simeq 1\text{M}\Omega$ ($\gg Z_0 = Z_1 = Z_1' = 100\text{k}\Omega$)

(b) Output impedance $\simeq 1\Omega$ ($\ll R_{on} = 50\Omega$)

(c) Zero d.c. offset.

(d) Unity voltage gain with no phase inversion.

The circuit designed is shown in Fig. 5 and measured parameters are:

Input impedance $Z_g \simeq 2\text{M}\Omega$ in parallel with 15pF

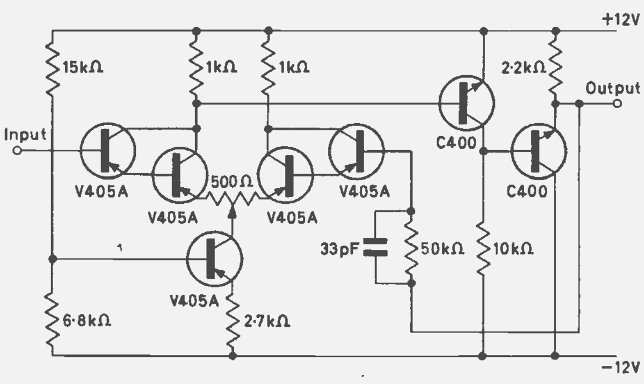
Output impedance $Z_s \simeq 1\Omega$

Bandwidth $\simeq$ 5Mc/s.

Performance of the Mode-Control Unit

A hold capacitor of 1 000pF is required to meet the specification of 10kc/s tracking bandwidth in conjunction with 60V peak value and 4mA charging current.

Fig. 5. Driver amplifier circuit



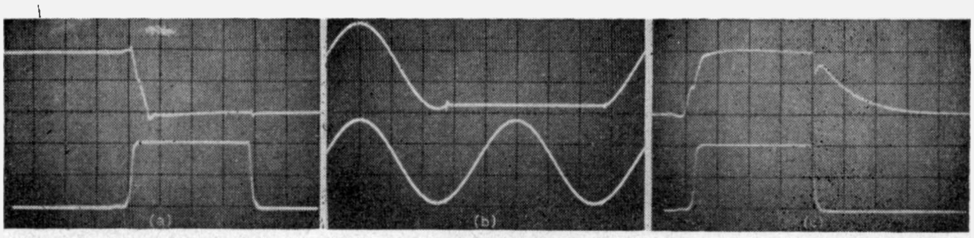


Fig. 6. Oscillographs showing performance of mode-control unit

(a) *Upper: Integrator reset 5V/cm : 5µsec/cm Lower: Control pulse ±10V 10V/cm : 5µsec/cm*

(b) *Upper: Track hold output Track period 800µsec Hold period 200µsec 5V/cm : 20µsec/cm Lower: Track/Hold input 10kc/s Sine wave 5V/cm : 20µsec/cm*

(c) *Upper: Output as s.p.d.t. switch 5V/cm : 5µsec/cm Lower: Control pulse ±10V 10V/cm : 5µsec/cm*

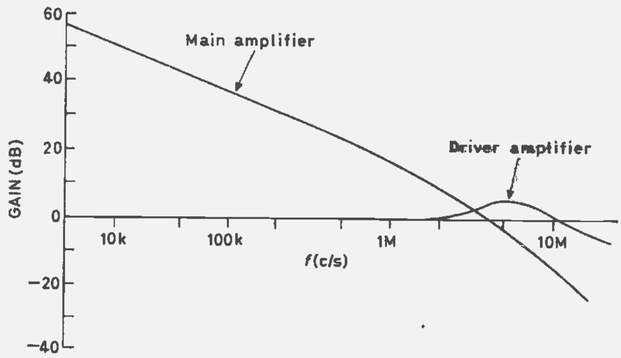


Fig. 7. Frequency response of main and driver amplifier

Using the open-loop frequency response plot for the computing amplifier, $A(p)$, shown in Fig. 7, the tracking error at 10kc/s is given by:

$$|\epsilon_r| = \frac{W_r}{1+T} = 1/139 = 0.72 \text{ per cent}$$

while the error due to a computer input X_1' is given by:

$$|\epsilon_d| = W_d \frac{T}{1+T} = \frac{0.001 \times 138}{139} \simeq 0.1 \text{ per cent.}$$

Oscillographs of the unit operating as a track-hold unit and as a single pole changeover switch are shown in Fig. 6.

Holding accuracy is determined by leakage through the switch and offset and drift within the main amplifier. Since the latter is chopper stabilized, the equivalent error current is principally switch leakage. Leakage currents of less than 10nA are typical for good quality diodes, contributing a maximum error of 0·02 per cent of full scale for a 1msec hold period.

Frequency response plots for the open-loop main amplifier together with the driver amplifier are drawn in Fig. 7. The actual stability margin is improved by the phase advance effect as the output impedance of the driver amplifier increases slightly with frequency.

Acknowledgments

The authors wish to thank Professor E. H. Frost-Smith for the facilities provided and also the Ministry of Education for Northern Ireland for the grant provided to one of them (P.L.N.).

Appendix

(A) Iteration Schemes Using Direct and Feedback Mode-Switching

Two analogue computer schemes are shown in Figs. 8

and 9, both of which generate the recurrence relationship for computing the initial condition in a two-point boundary values problem[1,5]:

$$X_{n+1}(0) = X_n(0) + \Delta(X_n)\ \mathrm{sgn}\ \delta$$

where $X_n(0)$ = initial condition at start of n^{th} iteration cycle.

$\Delta(X_n)$ = iteration step calculated during the n^{th} iteration cycle.

δ = perturbation amplitude.

The scheme of Fig. 8 is more economical in its use of computing amplifiers but the inputs to integrator I are gated through series switches which can lead to large cumulative errors.

The scheme of Fig. 9, on the other hand, uses no direct integration or switching so that errors are confined to tracking errors.

(B) Effect of Unbalance in Direct Switching of an Operational Amplifier

(i) When switching takes place at the virtual earth as in Fig. 10(a), the six diode bridge switch is replaced by its equivalent circuit and symmetry applied[7], Fig. 10(b).

E_u = unbalance in supply voltage when the switch is 'on'

R_f = forward resistance of the diode

R_c = series feed resistor for supply voltage.

$$E = \frac{R_f/2\,(R_1 + (R_f/2))}{R_f\,(R_1 + (R_f/2)) + R_c\,(R_1 + R_f)}\ .\ E_u$$

$$e_o = -E\,.\,(2R_o/E_f)$$

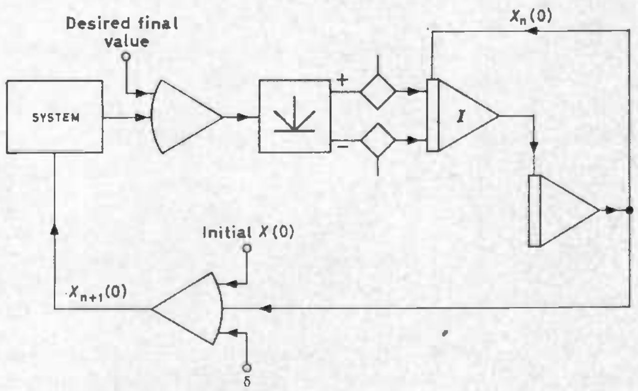
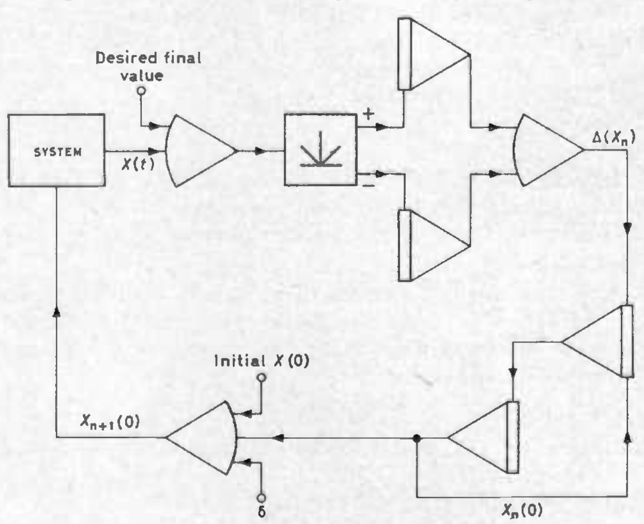


Fig. 8. Iteration scheme for solution of two-point boundary problem

Fig. 9. Iteration scheme for two-point boundary value problem



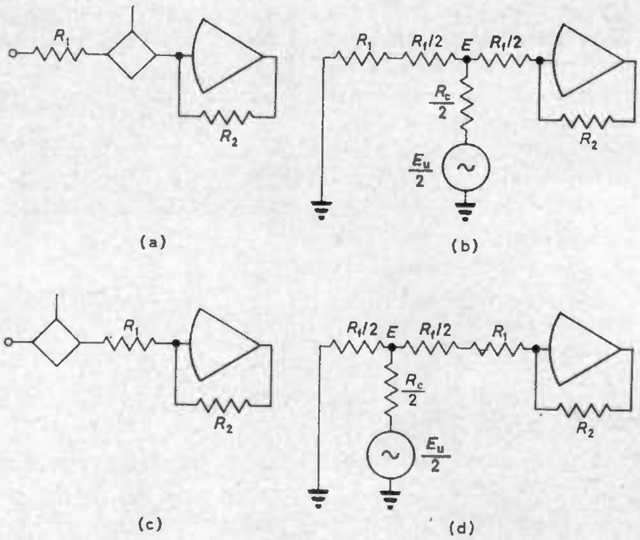


Fig. 10. Computer schematics and equivalent circuits for direct electronic switching

$$\therefore\ e_o = -\frac{R_2\,(R_1 + (R_f/2))}{R_f\,(R_1 + (R_f/2)) + R_2\,(R_1 + R_f)}\ .\ E_u$$

Typical values: $R_1 = R_2 = 100\text{k}\Omega$

$R_f = 200\Omega,\ R_o = 1{\cdot}8\text{k}\Omega$

$$\therefore\ e_o/E_u \simeq -\frac{R_2}{R_f + R_o}$$

giving $e_o/E_u = -50$

i.e.: the effect of slight unbalance of the bridge supply voltages is severe when switching takes place at the virtual earth.

(ii) When switching takes place before the input resistor R_1 as in Fig. 10(c), the switch is again replaced by its equivalent circuit, Fig. 10(d).

$$E = \frac{R_f/2\,(R_1 + (R_f/2))}{R_f\,(R_1 + (R_f/2)) + R_o\,(R_1 + R_f)}\ .\ E_u$$

$$e_o = -E\frac{R}{(R_1 + (R_f/2))}$$

$$\therefore\ e_o = -\frac{R_2\,(R_f/2)}{R_f\,(R_1 + (R_f/2)) + R_o\,(R_1 + R_f)}\ .\ E_u$$

Again $R_1, R_2 \gg R_f, R_o$ and using values as in (i)

$$e_o/E_u = -(R_o/R_1)\ .\ \frac{R_f}{R_f + 2R_o}$$

$$\therefore\ e_o/E_u = -0{\cdot}049$$

It is evident that unbalance effects are slight in this case, although the switch voltage gain becomes significant and can be < 1 if $R_o \ngtr R_f$.

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The Generation of Triangularly Pulse-Width Modulated Waves

By J. F. Young*, C.G.I.A., A.M.I.E.E., A.M.I.E.R.E.

For practical investigations of the performance of pulse-width-modulated amplifying systems, it is useful to have a source of pulses the width of which varies linearly or triangularly with time. Such a source can be produced by beating together two rectangular waves having slightly different frequencies, using either an AND *or an* EXCLUSIVE-OR *logic circuit as the mixer.*

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 837)

THE popularity of pulse-width-modulated systems has increased during recent years, partly because of the high efficiencies obtainable, with linearity, in amplifying systems. In the investigation of such systems it is helpful to be able to consider separately the modulator and demodulator sections. For the latter purpose it is useful to have available a source of rectangular waves having a width which varies linearly with time. So that repetitive testing with oscillographic display can be accomplished, the source should preferably provide a wave having a width which varies triangularly with time.

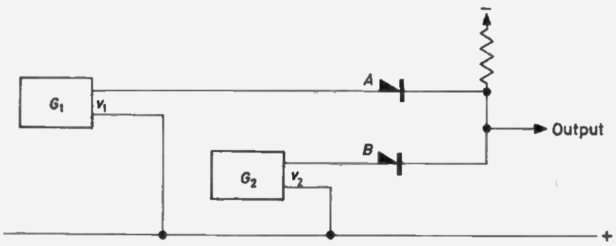

Fig. 1. Use of diode AND gate

One method of production of such a triangularly pulse-width-modulated wave is by use of a conventional pulse-width-modulator taking its input from a triangular-wave generator. Inherently in this method the linearity of the modulator and the triangular-wave generator must both be considered. This is unfortunate if the object is merely to investigate the linearity of the pulse amplifier and demodulator. A more direct method of production of triangularly pulse-width-modulated waves might therefore be of interest.

One simple method of production of a form of triangularly p.w.m. wave is by taking the logical product of two rectangular waves, for example with a simple AND gate as shown in Fig. 1. Here rectangular wave generators G_1 and G_2 produce outputs having identical amplitudes but slightly different repetition frequencies. The fact that a negative-signal AND gate is used in Fig. 1 has no significance beyond the fact that this is the form which has been used in practical work.

Taking a unidirectional rectangular wave A having angular frequency ω_1, amplitude V and unity mark-to-space ratio, the Fourier expression is:

$$A = v_1 = (V/2)+(2V/\pi)(\sin\omega_1 t + (1/3)\sin 3\omega_1 t+(1/5)\sin 5\omega_1 t+\ldots) \quad \ldots\ldots (1)$$

A second wave having the same amplitude V but a different angular frequency ω_2 is:

$$B = v_2 = (V/2)+(2V/\pi)(\sin\omega_2 t + (1/3)\sin 3\omega_2 t+(1/5)\sin 5\omega_2 t+\ldots) \quad \ldots\ldots (2)$$

Now A and B can only have the instantaneous values V or 0, so that the instantaneous product of the two waves can be expressed by the following table:

A	B	Product	AND
0	0	0	0
V	0	0	0
0	V	0	0
V	V	V^2	1

By comparison of this table with the truth table for a logical AND gate, it can be seen that an AND gate can be used to generate a signal proportional to the instantaneous product of the two rectangular waves. Expressed in Fourier form, the product is obtained by multiplying equations (1) and (2) term by term, giving:

$$\begin{aligned} A.B = v_1 v_2 = (V^2/4) & \\ +(V^2/\pi)[&\sin\omega_1 t+(1/3)\sin 3\omega_1 t+(1/5)\sin 5\omega_1 t+(1/7)\sin 7\omega_1 t+\ldots \\ &\sin\omega_2 t+(1/3)\sin 3\omega_2 t+(1/5)\sin 5\omega_2 t+(1/7)\sin 7\omega_2 t+\ldots] \\ +(4V^2/\pi^2)[&\sin\omega_1 t\,.\,\sin\omega_2 t+(1/3)\sin\omega_1 t\,.\,\sin 3\omega_2 t+(1/5)\sin\omega_1 t\,.\,\sin 5\omega_2 t+\ldots \\ &+(1/3)\sin 3\omega_1 t\,.\,\sin\omega_2 t+(1/9)\sin 3\omega_1 t\,.\,\sin 3\omega_2 t+(1/15)\sin 3\omega_1 t\,.\,\sin 5\omega_2 t+\ldots \\ &+(1/5)\sin 5\omega_1 t\,.\,\sin\omega_1 t+(1/15)\sin 5\omega_1 t\,.\,\sin 3\omega_2 t+(1/25)\sin 5\omega_1 t\,.\,\sin 5\omega_2 t+\ldots \\ &+\ldots\ldots] \quad \ldots\ldots\ldots\ldots (3) \end{aligned}$$

Here, the first term is a constant while the second term contains the sum of the two original rectangular waves. These two terms could be eliminated completely by a.c. coupling the two original rectangular waves by some form of high-pass filter, for example capacitive- or transformer-coupling. This would eliminate the $V/2$ term from both A and B. However, the construction of the instantaneous product in the form of a truth table then shows that a simple AND gate could not be used to obtain the product, though an inverted EXCLUSIVE-OR circuit could be used:

A	B	PRODUCT	EXCLUSIVE-OR
$-(V/2)$	$-(V/2)$	$+(V^2/4)$	0
$+(V/2)$	$-(V/2)$	$-(V^2/4)$	1
$-(V/2)$	$+(V/2)$	$-(V^2/4)$	1
$+(V/2)$	$+(V/2)$	$+(V^2/4)$	0

This will be mentioned again later.

The final bracket of equation (3) contains terms which

* *The University of Aston.*

can be expanded by making use of the form:

$$\sin \omega_a t \, . \, \sin \omega_b t = \tfrac{1}{2} \cos (\omega_a - \omega_b)t - \tfrac{1}{2} \cos (\omega_a + \omega_b)t$$

When this is done, it can be seen that the diagonal (underlined) terms contain the expansion:

$$2V^2/\pi^2 \left[\cos(\omega_1 - \omega_2)t + (1/9)\cos 3(\omega_1 - \omega_2)t + (1/25)\cos 5(\omega_1 - \omega_2)t + \ \ldots\right]$$

This is the Fourier expansion of a triangular wave with repetition angular frequency $(\omega_1 - \omega^2)$. Thus the product wave contains a 'beat-frequency' triangular wave. Similarly, it also contains a 'sum-frequency' triangular wave, as well as the non-diagonal terms.

Thus a simple AND gate can be used to produce a wave having varying pulse-width and containing a low frequency triangular wave. If a three-input AND gate is used instead, a more complex form of output, containing two beat-frequency triangular waves is obtained. If the third rectangular-wave input is made equal in frequency but opposite in phase to one of the other inputs, the AND gate can produce no output. Thus it is possible to analyse a wave such as that above by applying it to one input of a two-input AND gate, the other input of which is obtained from a rectangular-wave generator. The frequency of the rectangular-wave generator is then adjusted until the output of the AND gate is zero. In this way an experimental form of rectangular-wave Fourier analysis is possible. This is one way of obtaining a very selective determination of the presence of a particular frequency of rectangular wave in the original waveform.

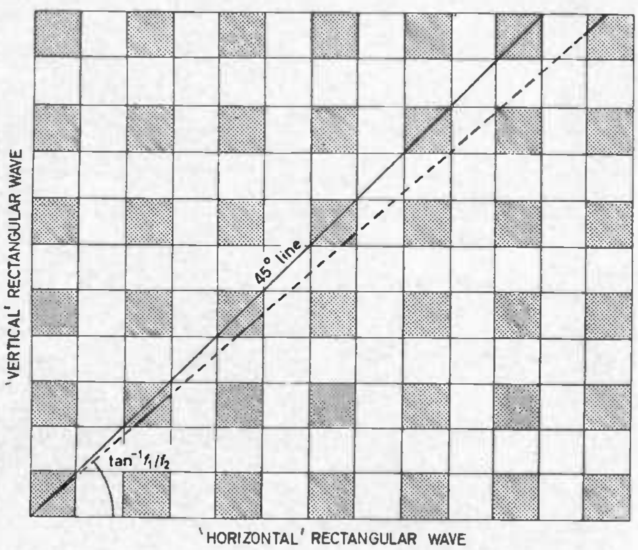

Fig. 2. Chessboard derivation of waveshape

One way of considering the actual shape of a wave such as that discussed is by means of a chess-board pattern such as that of Fig. 2. Here the two input rectangular waves, each of unit amplitude, are represented, one by the horizontal axis and the other by the vertical axis. Vertical and horizontal lines on the plane represent the changeover points of the two rectangular waves. Shaded areas on the plane indicate regions where both inputs are simultaneously positive. On such a diagram, the progression of time can be represented by a sloping line as shown. If the two rectangular waves have the same frequency and phase (i.e. they are identical), then the sloping line must be at 45° and it must pass through the origin. To indicate a phase difference between the two waves, the sloping line must be moved vertically. If there is a frequency difference between the two waves, then the slope of the line must be changed from 45° to $\tan^{-1} f_1/f_2$, where f_1 and f_2 are the two frequencies.

On such a diagram, points where the sloping line crosses a shaded region correspond to times when both input waves are simultaneously positive, whereas in the unshaded regions one or both of the waves is negative. Consequently, points where the sloping line crosses a shaded region correspond to times when a positive signal AND gate, having the two waves as inputs, would give a positive output.

Careful examination of the chessboard form of diagram shows how the pulse-width modulated wave is built up when two rectangular waves are applied to an AND gate. In the first half-cycle of the beat-frequency triangular wave, the leading edges of the modulated pulses are produced by the vertical waveform in this example. During the same half-cycle, the trailing edges of the modulated pulses are produced by the horizontal waveform. During the next half-cycle the leading edges of the modulated pulses are produced by the horizontal waveform, while the trailing edges are produced by the vertical rectangular wave. On such a diagram, the effect of a change of the relative frequencies—corresponding to a change of slope of the straight line—can be seen easily.

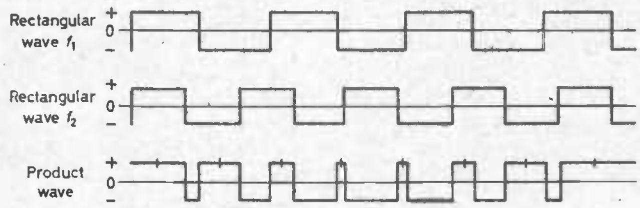

Fig. 3. Product of two rectangular waves

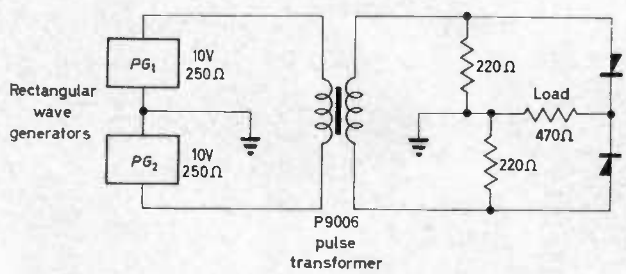

Fig. 4. Use of full wave rectifier

Although the width of pulses produced in the manner described changes uniformly with time in each half cycle, the resulting wave would not always be regarded as being a true pulse-width modulated wave. The reason for this is that the centres of the pulses are not always evenly spaced in time. This can be seen in Fig. 2. Uneven spacing occurs at the end of each half-cycle, coinciding with the changeover of control of leading and trailing edges from one input waveform to the other. In some applications this fact will be unimportant, but in others it will be an undesirable feature.

The possibility of using an EXCLUSIVE-OR circuit has been mentioned earlier. In Fig. 3, two rectangular waves of differing frequencies are shown. Both waves are symmetrical about zero voltage, so that the mean value is zero and neither has any d.c. content. If each wave has unit amplitude, the instantaneous product of two waves is as shown. When both waves simultaneously have the same sign, the instantaneous product is positive, but when the signs differ the product is negative. It will be noted that the product wave consists of a triangularly pulse-width modulated wave, the centres of the pulses being evenly spaced in time.

This product wave could be obtained using some device giving an instantaneous multiplication. However, because of the nature of the waves involved, this would be difficult to accomplish. A somewhat easier method of obtaining the product can be seen with reference to the instantaneous sum of the two waves, also drawn in Fig. 3. It will be seen that simple full-wave rectification of this sum-wave will give the required product wave, plus a d.c. component.

Basically, full-wave rectification of the sum gives the product because it gives a positive output only when both inputs are positive or both are negative. When only one input is positive and the other is negative, no output is obtained. Thus this method in fact gives the inverse of the EXCLUSIVE-OR mentioned earlier, and therefore it gives a product wave. A circuit which has been used to give a full-wave rectified pulse wave of this type is given in Fig. 4.

An alternative approach would be to apply the two rectangular waves to a logical circuit such as that of Fig. 5, which can produce an EXCLUSIVE-OR function directly. This can easily be built, for example using standard static switching units[2].

One use of the techniques described here is to make startlingly obvious the deficiencies of the simple non-clamped integrator method of demodulation of pulse-width modulated waves. In the simple integrator method

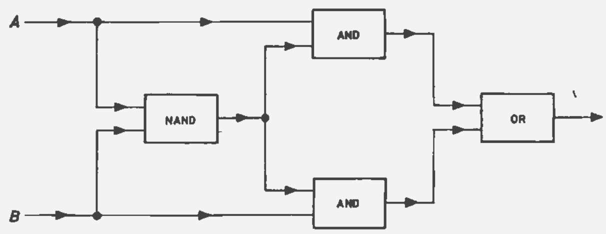

Fig. 5. EXCLUSIVE-OR logic

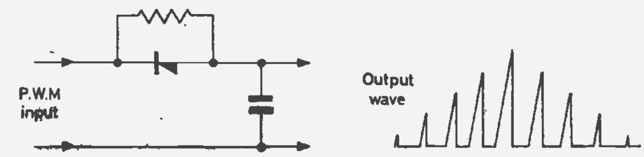

Fig. 6. Simple clamped demodulator

the high frequencies present in the triangular wave are eliminated by the integrator. However, a simple clamped integrator such as that of Fig. 6 can reproduce the triangular wave in the form of varying pulse heights. Such a simple demodulation circuit cannot always be regarded as satisfactory because pulse overlap can cause spikes to appear in the output wave.

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A Compensated Transistor D.C. Stabilizer

By M. Pacák*

A stabilizer circuit is described enabling high grade stabilization of direct current or voltage adjustable over a wide range, with a long-term stability of $\pm 1{\cdot}5 \times 10^{-4}$ in eight hours, the short-term stability $\pm 2{\cdot}5 \times 10^{-6}$ within several minutes. The circuit involves the use of positive feedback and is compensated for the main disturbing effects.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

THE general concept of transistor regulators or stabilizers and of valve ones is basically the same, differing only in matters such as working voltages and currents. Consequently, the same precautions are to be taken and similar results are obtainable with both types. However, the specific transistor properties should be respected, especially their need for small signal source resistance and their steep temperature dependence. On the other hand, transistor slow noise is substantially smaller in comparison with valves, offering favourable properties to the transistor regulator: the possibility of the use of smaller reference voltage and/or higher precision attainable, smaller output voltage or its wider adjustability, better power efficiency, etc.

In Fig. 1 a stabilizer circuit of simple design and rather good properties is presented. With simple modification concerning the reference part of the device, the circuit may function as a voltage or current stabilizer. The stabilized output value may easily be adjusted over a wide range, between 1·5 and several tens of volts or from practically zero to several amperes according to the power elements used, with the relative stability remaining the same and being better than 1 : 10 000. With some precautions, over one hundred volts may be stabilized with a common germanium power transistor.

The stabilizer circuit consists of the following parts: reference circuit, error amplifier, control element and supplying sources. The reference circuit involves a reference source, for which a small 1·5V cell was used, and a voltage divider R_1, R_2 for the voltage stabilization, or a reference resistor R_I for the current stabilizer function. The underlined resistance values indicate wirewound resistors with as low as practicable temperature coefficient.

There is only one voltage amplifying stage in the error amplifier symmetrically arranged with two identical transistors mounted closely one to another in a metal block to keep their temperatures identical. As the stabilizer output voltage may change appreciably when the stabilized value is adjusted, the amplifier is supplied by a separate source V_b, pre-stabilized by a Zener diode D_b and connected to the main circuit by a 1 + 1kΩ divider. The following cascade, VT_2, VT_2' and VT_2'', controls the

* *Institute of Physical Chemistry CSAV, Prague.*

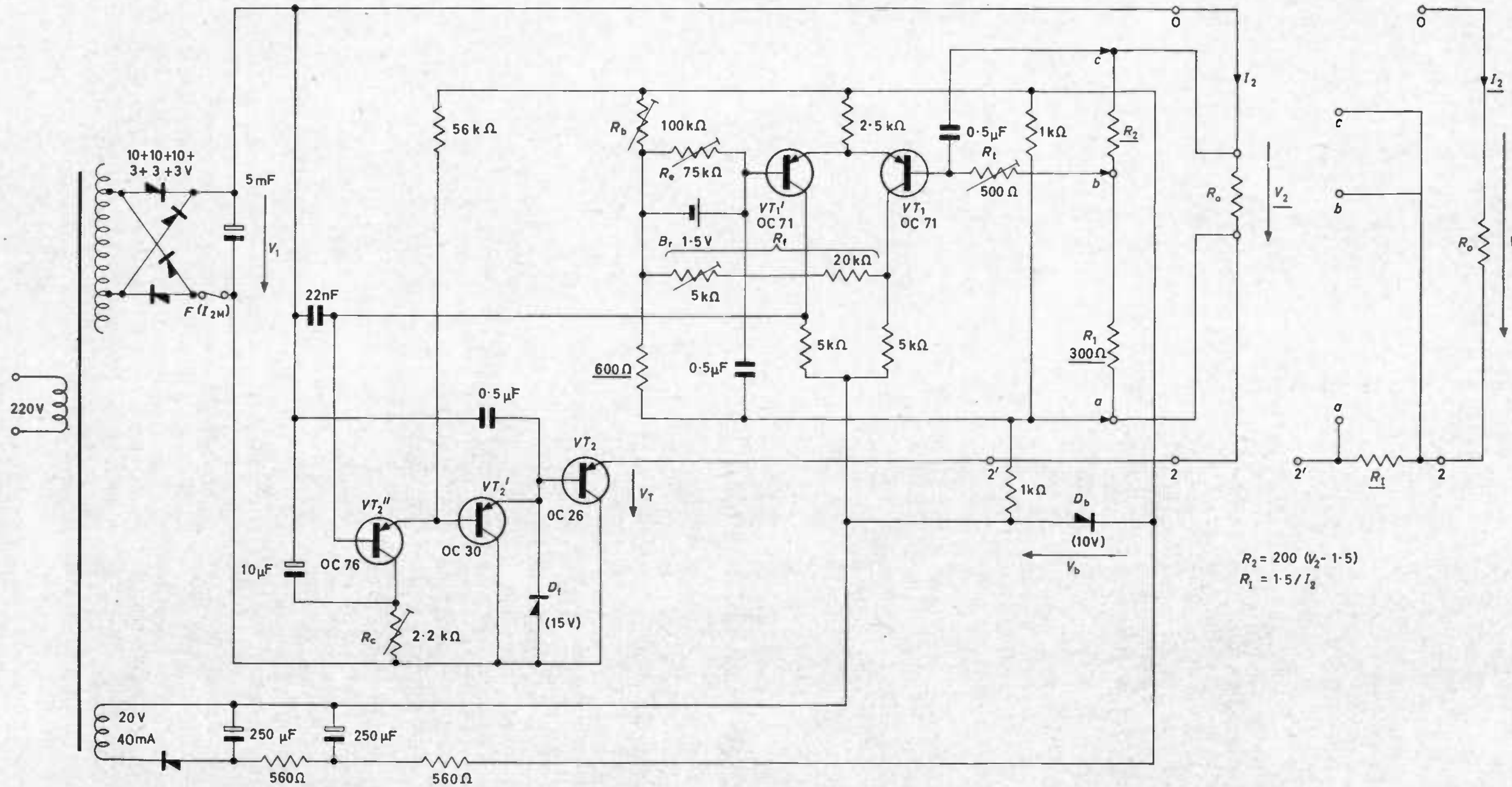

Fig. 1. The voltage or current stabilizer

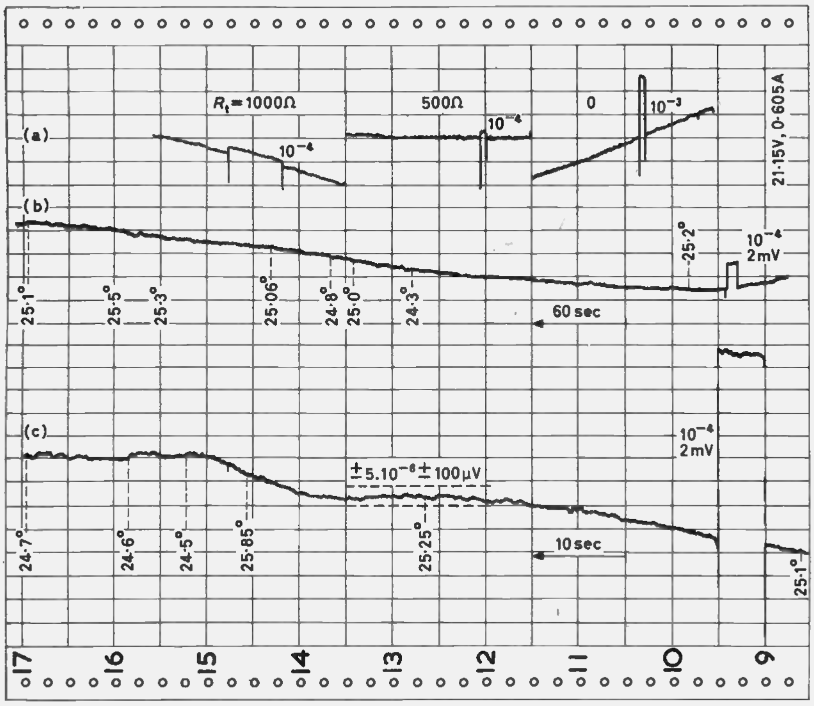

Fig. 2. Stabilizer performance

(a) Run-in period, stabilized voltage/time record for three values of temperature compensating resistance
(b) Long term stability, recorded for 8 hours
(c) As (b) with increased sensitivity and paper speed

output value according to the amplified error signal, and the cascade power capacity, together with the mains transformer and rectifier, limits the stabilizer output current and power. The capacitors, 22nF and 0·5µF ensure the dynamic stability and fast stabilizer action.

For attaining good overall stabilization, several compensating arrangements were adopted. The resistors R_f close the small positive feedback loop in the amplifier, raising substantially its amplification and therefore the regulation quality[1]. The following procedure may be used for adjustment: the stabilized value, voltage or current, is measured by the compensating method, which shows very distinctly any variation. The ballast resistor R_o, set to the preferred output power value, is then changed by some 20 per cent in any direction and at the same time the variable part of R_f is adjusted till the arising stabilized value variations decrease or vanish.

The resistor R_t connected in the base lead of the transistor VT_1 enables one to find a ratio of amplifier input resistances at which the temperature variations, acting equally on both symmetrically arranged transistors, cancel each other and do not cause any change of the stabilized value. For the adjustment a starting period may be used with the instrument temperature rising substantially. The resistor R_t is then changed to obtain only a short and indistinct run-in stabilized value variation. The correct value of R_t is influenced a little by the divider R_1, R_2 and therefore the compensation may be exactly set only for one chosen value of the stabilized voltage.

The resistor R_b introduces into the amplifier a small correcting signal proportional to the amplifier supply voltage fluctuations, remaining in spite of the pre-stabilization. Its adjustment may easily be done by introducing an artificial disturbance of V_b. — The resistor R_e, bypassing the reference cell, discharges it by some 20µA and compensates the charging effect of the transistor base current, which otherwise causes irregular change of reference voltage which reproduces itself proportionally in the stabilized value.

After proper adjustment of positive feedback loop gain the stabilizer static parameters approach closely to their ideal values, e.g. the output resistance of the voltage stabilizer tends to zero within a limit of one milliohm or so. However, the usual copper leads to the load add some 20mΩ for each yard of length. To prevent this stabilizer quality degradation, one may connect the reference circuit not as usually into the stabilizer circuit, but as shown in Fig. 1, through separate 'potential' leads to the load terminals. In this way the lead resistance is completely eliminated by the regulating action.

For voltage stabilization the output voltage V_2 is adjusted to the desired value by means of the resistor R_2. If a single value of V_2 is desired, the wire-wound fixed resistor R_2 should be used, giving the best precision attainable. If a ready and continuous adjustment of the stabilized voltage is desired, a precision variable resistor should be used as R_2, enabling a direct proportional change of V_2. For stepwise (digital) adjustment a resistance decade may be employed. If a wider output voltage adjustability is desired or if its fixed value is greater than some 20V, the input voltage V_1 must be changed to ensure the power transistor voltage V_T remains in proper limits. A Zener diode D_t connected between the power transistor collector and base opens the controlling cascade when its voltage overpasses the safe value.

Should the circuit function as a current stabilizer, a reference resistor R_I, which also may be fixed, bypassed by a potentiometer or replaced by a decade, is used instead of the voltage divider. In both applications a fuse in the main circuit protects the instrument against current overload.

The stabilizing properties of the circuit just described may be judged from the original voltage-time record in Fig. 2. A voltage of about 20V was stabilized, the loading current being 0·6A and its variations were recorded by a compensator millivoltmeter. First, the temperature compensation is demonstrated by means of the resistor R_t, enabling the adjustment of rising or falling run-in curve. In the second part, the long term stability of the circuit is demonstrated, showing a slow, monotonous drift of 6mV for 8 hours, giving the relative limits of $\pm 1{\cdot}5 \times 10^{-4}$ for this time. In the third part of the record the meter sensitivity and paper speed were raised ten times and the short-term variations became visible being at most 100µV i.e. $2{\cdot}5 \times 10^{-6}$ for at least 5min. It should be mentioned, that as long as the reference voltage remains unchanged during the stabilized value adjustment, the relative stability given above should remain constant too for any value set, provided the proper instrument function. On the contrary, if some loss in precision may be allowed, a smaller reference voltage may be employed enabling effective stabilization of small voltages of the order of 0·1V or so.

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A Variable Gain Amplifier

By S. Ghosh*

An amplifier is described the gain of which can be varied over a range of about 60dB by changing one resistance value only, without appreciably altering frequency characteristic, input and output impedances and loop gain of the amplifier.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

THE gain of an amplifier can be varied by:

(a) Controlling some intrinsic parameter of the active device (e.g. emitter resistance of a transistor).

(b) Introducing a variable loss network somewhere in the amplifier.

(c) Changing the feedback in the amplifier.

The variable gain amplifier described here is based on principle (c) above. The negative feedback of the amplifier is varied by changing one resistor only in such a way that the gain varies linearly with the resistance. However, the loop gain does not change appreciably so that frequency response, distortion margin, input and output impedances are practically unaffected by variation of amplifier gain.

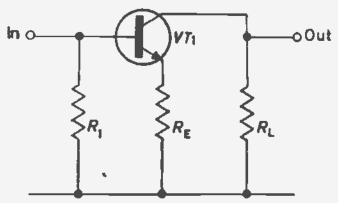


Fig. 1. Amplifier with emitter degeneration
R_E = external emitter resistance
R_L = effective collector load

Principle of Operation

Fig. 1 shows the basic a.c. circuit of a single stage amplifier with emitter degeneration. If the current gain is large (which can be achieved by using a Darlington pair) the voltage gain of the amplifier is approximately given by

$$A_v = \frac{R_L}{R_E + r_e} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

where r_e is intrinsic emitter resistance.

The gain of this stage will vary inversely with R_E provided $R_E \gg r_e$. However, this simple configuration suffers from the disadvantage that R_L has to be large for high gain, and this limits its high frequency response.

A better arrangement is the feedback pair shown in Fig. 2.

The voltage gain of this amplifier is approximately given by

$$A_v = \frac{R_F + R_E}{R_E} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

Therefore the gain can be controlled by varying R_E provided that $R_F \gg R_E$. As however R_E is increased gain asymptotes towards unity.

This difficulty can be overcome by replacing the first stage of the amplifier by a cascode arrangement shown in Fig. 3.

In the Appendix, the gain of this amplifier, assuming large feedback has been approximately shown to be

$$A_v = R_F/R_E \quad \ldots\ldots\ldots\ldots\ldots (3)$$

Therefore the gain of the amplifier decreases linearly as R_E is increased, or gain in decibels decreases directly as log R_E. There are practical limits to which R_E can be increased or decreased. As R_E is increased, the input impedance looking into XX becomes large and signal input is shunted by base collector impedance of VT_1. On the other hand as R_E becomes very small the loop gain starts dropping and the approximate relationship of equation (3) does not hold.

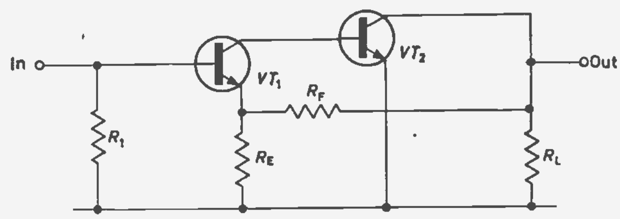


Fig. 2. Feedback pair
R_F = feedback resistance

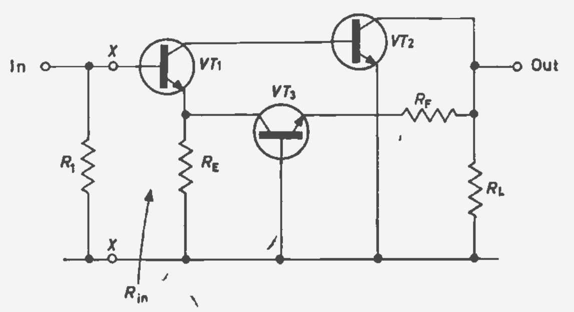


Fig. 3. Modified feedback pair

Brief Circuit Description

Fig. 4 shows the practical circuit of the amplifier.

VT_2 in Fig. 3 has been replaced by a pair of transistors for large loop gain. Also pnp transistors have been used so that the 10kΩ feedback resistor provides both a.c. and d.c. feedback. The choice of transistors and component values is dictated by the frequency range of interest and gain required. The high frequency response of the amplifier can be shaped by C_1 or other networks in its place. Low frequency performance is determined by capacitors C_2 and C_3.

For large values of R_E, the shunting effect of the output capacitance of VT_3 becomes significant and may cause a high frequency peak in the frequency response. For large bandwidth, therefore, VT_3 should be chosen so that its output capacitance is small.

* *Standard Telephones & Cables Ltd.*

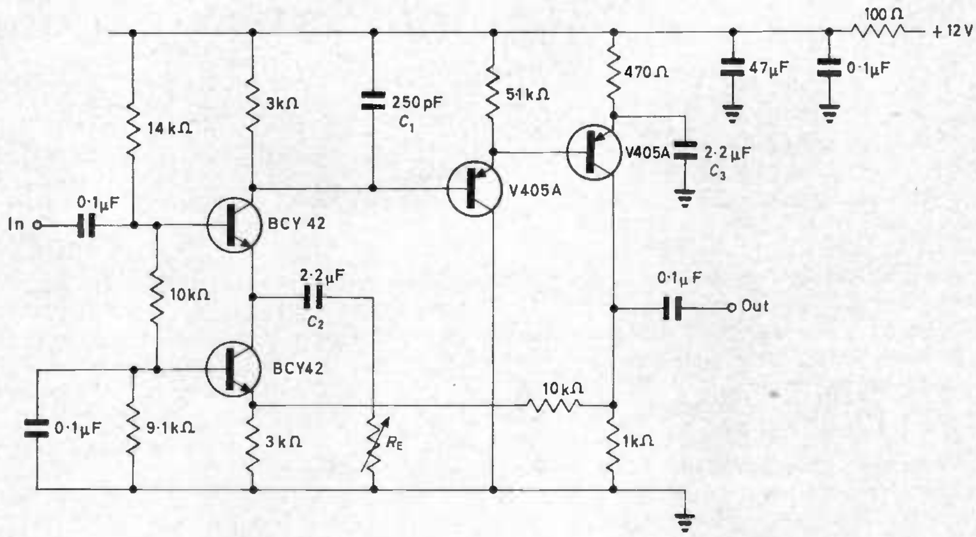


Fig. 4. Variable gain amplifier

Performance

(a) *Gain Variation*—As R_E is increased from 12Ω to 20kΩ gain of the amplifier is reduced from +58dB to −6dB almost exactly according to the relation given by equation (3).

(b) *Frequency Response*—The high frequency cut-off (3dB point) of the amplifier occurs near 5Mc/s.

(c) *Maximum Output*—2·5V peak.

(d) *Input Impedance*—Over the range of gain variation as in (a), the input impedance varies from 2kΩ to 5·8kΩ.

(e) *Output Impedance*—Over the same range of gain variation, the output impedance changes from 300Ω to 100Ω.

(f) *Distortion*—With maximum output both second order and third order harmonic margins are better than 40dB.

(g) *Noise*—Effective noise power at the input, when the gain of the amplifier is maximum is about −124dBm. This gets worse by about 10dB under minimum gain condition.

Acknowledgment

The author wishes to thank the management of Standard Telephones & Cables Ltd for permission to publish this article.

Appendix

Analysis of the Amplifier Shown in Figs. 3 and 5

In this approximate analysis the transistors are assumed to be simple amplifying devices, and the effects of their intrinsic parameters are ignored.

$$V_{in} = R_E(i_f + i_{c1})$$

$$i_f = i_{c2}\frac{R_L}{R_F + R_L} = A_i\, i_{c1}\frac{R_L}{R_F + R_L}$$

where A_i = effective current gain from collector of VT_1 to collector of VT_2

$$\text{Therefore, } V_{in} = R_E i_{c1}\left(1 + A_i\frac{R_L}{R_F + R_L}\right) \quad \ldots\ldots\ (4)$$

$$\text{Also}\quad V_o = A_i i_{c1}\frac{R_F R_L}{R_F + R_L} \quad \ldots\ldots\ (5)$$

Hence voltage gain $A_v = V_o/V_{in} \simeq R_F/R_E$ if A_i is large .... (6)

From equation (4), the input impedance R_{in} looking into terminals XX can be estimated.

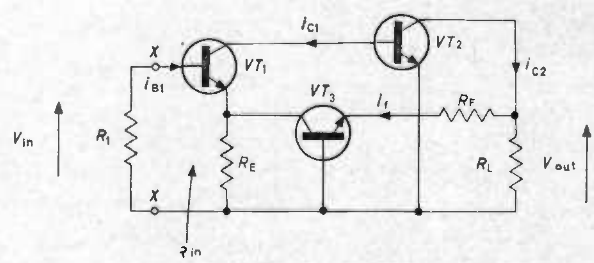


Fig. 5. Circuit for analysis

$$R_{in} = V_{in}/i_{B1} = R_E\beta_1\left(1 + A_i\frac{R_L}{R_F + R_L}\right) = R_E\beta_1(1 + A_i') \quad \ldots\ (7)$$

where β_1 = common emitter current gain of VT_1 and A_i' = loop current gain with input short-circuited.

To estimate output impedance, R_L is replaced by a voltage source V_1 and V_{in} is replaced by its appropriate source resistance.

With this arrangement,

$$i_f = V/R_F$$

$$i_{c1} = Ki_f$$

where $K = \frac{R_E}{R_E + Z}$, Z being the impedance looking into the emitter of VT_1.

and $i_{c2} = A_iK(V/R_F)$.

Output impedance is therefore

$$R_o = \frac{V}{i_f + i_{c2}} = \frac{R_F}{1 + A_iK} = \frac{R_F}{1 + A_i''}$$

where A_i'' = loop current gain with output open circuit.

Design for a Parallel Binary Adder

By J. B. Earnshaw* and P. M. Fenwick*

This article describes the basic design of a parallel binary adder with high speed carry propagation. The design is based on the use of a two-level transistor-diode feedback-type logic circuit which has a delay approaching 1nsec per logic level. Initial measurements indicate that the sum generation time for words containing 16 bits is approximately 50nsec.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

IN any form of parallel adder the speed with which the carry digit is propagated through the adder becomes the dominant factor in determining the overall time required to complete a sum. This is because it is necessary, in the worst case, for the carry digit to propagate from the least to the most significant stage of the adder before the final sum output is correct.

Clearly the major problem in the design of a parallel adder is that of reducing the overall carry propagation delay. One solution to this problem is to effectively shorten the length of the carry-path by using rather sophisticated 'carry-skip' or 'carry-look-ahead' logic schemes[1,2]. A more direct solution, initially employed by Kilburn *et al.*[3] and extended somewhat herein, is to reduce the number of logical operations to be performed on the carry digit and thereby decrease the actual carry propagation delay in each stage of the adder.

Rules For Addition

If s_k and c_k are respectively the sum and carry digits from the k^{th} stage of an adder when the three inputs to this stage are x_k and y_k, the k^{th} digits of the two words to be added, and c_{k-1}, the carry digit from the $(k-1)^{th}$ stage ... the truth table for addition is as set out in Table 1.

TABLE 1
Truth table for (a) binary addition, (b) associated functions

x_k	y_k	c_{k-1}	s_k	c_k	$x_k \equiv y_k$	$x_k \not\equiv y_k$	$x_k . y_k$	$\bar{x}_k . \bar{y}_k$
0	0	0	0	0	1	0	0	1
1	0	0	1	0	0	1	0	0
0	1	0	1	0	0	1	0	0
1	1	0	0	1	1	0	1	0
0	0	1	1	0	1	0	0	1
1	0	1	0	1	0	1	0	0
0	1	1	0	1	0	1	0	0
1	1	1	1	1	1	0	1	0

(a) (columns x_k to c_k) (b) (columns $x_k \equiv y_k$ to $\bar{x}_k . \bar{y}_k$)

The Kilburn rules for addition, which minimize the logical operations in the carry-path, are based on the 'equivalence' and 'non-equivalence' relations between x_k and y_k. They are presented below in the form most suitable for implementation by the Kilburn adder, viz.:

(a) When $x_k \equiv y_k$: sum ... $s_k = c_{k-1}$
carry ... $c_k = 0$ if $\bar{x}_k . \bar{y}_k = 1$
$c_k = 1$ if $x_k . y_k = 1$

(b) When $x_k \not\equiv y_k$: sum ... $s_k = \bar{c}_{k-1}$
carry ... $c_k = c_{k-1}$

* *University of Auckland, New Zealand.*

The schematic diagram of a single stage of a parallel adder based on the Kilburn rules is given in Fig. 1. The switches which determine the sum and carry outputs are closed by appropriate logical functions generated from the x and y digits particular to this stage. Because the control of these switches is independent of the nature of the incoming carry signal, corresponding switches throughout all stages of the adder can be operated simultaneously in a time referred to as the 'set' time, t_s.

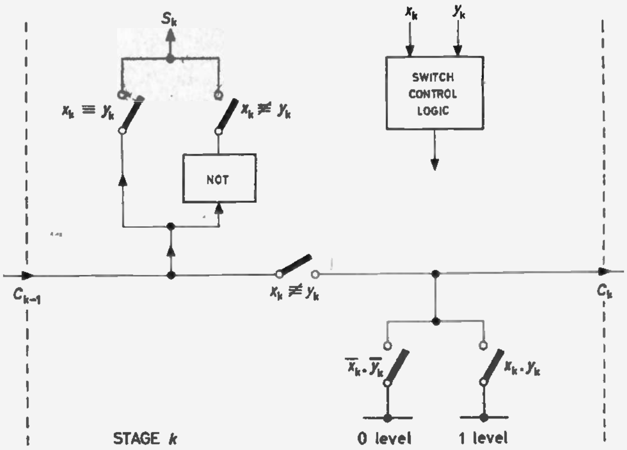


Fig. 1. Kilburn parallel adder—stage k

Thus the time to generate the correct sum output at the n^{th} stage, in the worst case, is:

$$t_s + (n-1)\,\delta_c + \delta_s \quad \ldots\ldots\ldots\ldots\ (1)$$

where δ_c is the carry propagation delay per stage and δ_s is the delay in the sum output.

The carry-path in the Kilburn adder uses saturated transistors in a uniquely devised non-inverting mode of operation. For a carry-path using more conventional gate-invertor coupling techniques, the Kilburn rules require reformulation, because the carry signal is propagated as carry and NOT-carry by alternate stages of the adder.

Expressed in Boolean form, the Kilburn rules for addition at stage r, say, become:

$$\text{sum} \ldots s_r = (x_r \equiv y_r) . c_{r-1} + (x_r \not\equiv y_r) . \bar{c}_{r-1} \quad \ldots\ (2)$$

$$\text{carry} \ldots c_r = (x_r \not\equiv y_r) . c_{r-1} + x_r . y_r \quad \ldots\ (3)$$

Dealing with the sum relationship first, equation (2) may be rewritten as:

$$s_r = \overline{\overline{(x_r \equiv y_r) . c_{r-1} + (x_r \not\equiv y_r) . \bar{c}_{r-1}}}$$

$$= \overline{\overline{\{(x_r \equiv y_r) . c_{r-1}\}} . \overline{\{(x_r \not\equiv y_r) . \bar{c}_{r-1}\}}}$$

$$= \overline{\{(x_r \not\equiv y_r) + \bar{c}_{r-1}\} . \{(x_r \equiv y_r) + c_{r-1}\}} \quad \ldots\ (4)$$

The r.h.s. of equation (4) is in a form suitable for implementation by a NOT(OR-AND) circuit arrangement when the functions $(x_r \equiv y_r)$ and $(x_r \not\equiv y_r)$ are already available.

The carry relationship in equation (3) may be rewritten as:

$$c_r = \overline{\overline{(x_r \not\equiv y_r) \cdot c_{r-1} + (x_r \cdot y_r)}}$$
$$= \overline{\overline{\{(x_r \not\equiv y_r) \cdot c_{r-1}\}} \cdot \overline{\{x_r \cdot y_r\}}}$$
$$= \overline{\{(x_r \equiv y_r) + \bar{c}_{r-1}\} \cdot \{\bar{x}_r + \bar{y}_r\}} \quad \ldots\ldots\ldots\ldots (5)$$

This also is in a form suitable for NOT(OR-AND) circuit implementation.

Inverting both sides of equation (3) the NOT carry relationship is:

$$\bar{c}_r = \overline{(x_r \not\equiv y_r) \cdot c_{r-1} + (x_r \cdot y_r)} \quad \ldots\ldots\ldots\ldots (6)$$

In this case a NOT(AND-OR) circuit is required.

Logic Circuits

A high speed transistor-diode feedback-type logic circuit, which possesses the necessary two-level logic capability to implement the functions in equations (4), (5) and (6), has been developed and is described in detail elsewhere[4-6]. This logic circuit consists of a current-voltage transfer-type invertor preceded by an appropriate current steering gate. Fig. 2(a) shows the invertor which is a current driven saturated-mode circuit employing non-linear current feedback to limit the collector voltage excursions. The circuits of four alternative forms of current-steering diode gates are shown in Figs. 2(b) to 2(e), and two of these gates perform two-level logic. Various arrangements of the logic circuit are shown symbolically in Fig. 3, and the diode gates used in these arrangements correspond directly with those shown in Fig. 2. The basic delay of the logic circuit is nominally 2nsec and this is increased by approximately 0·5nsec for each of the fan-out circuits. Thus with the two-level circuits the delay per logical decision approaches 1nsec.

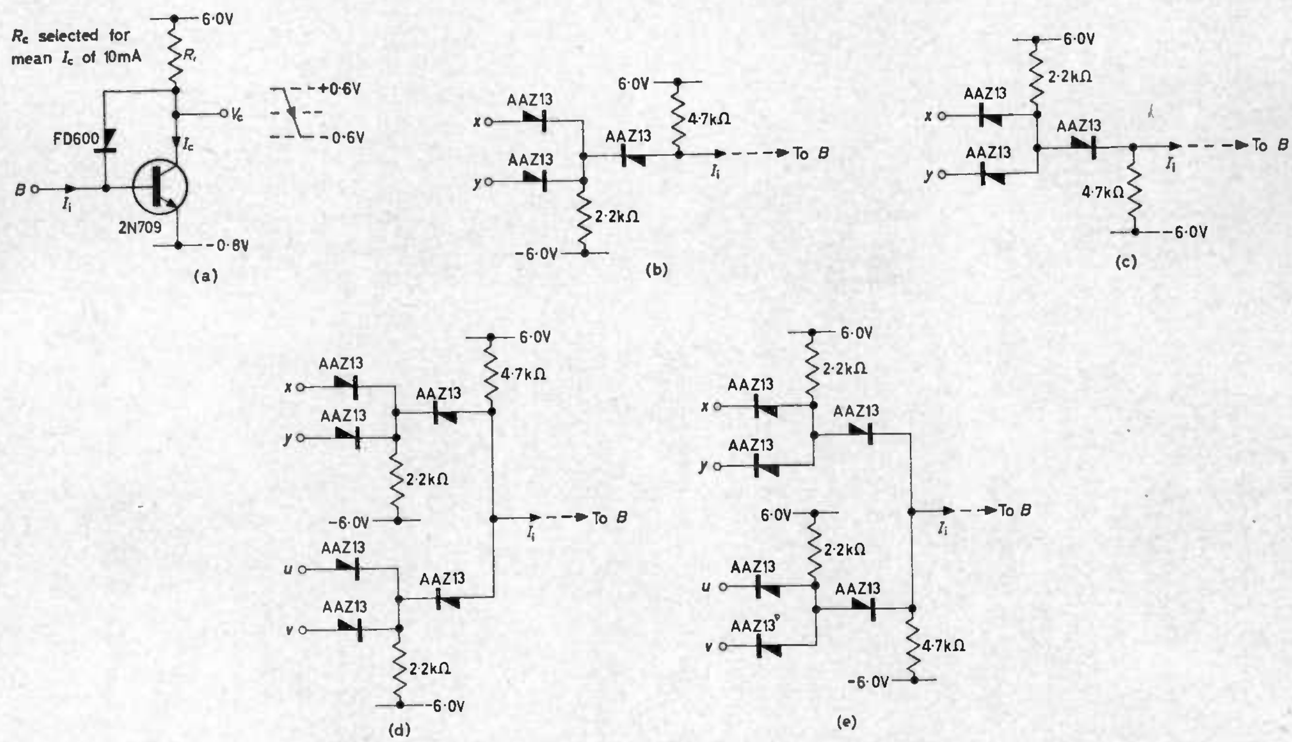


Fig. 2. Transistor-diode logic circuits

(a) Transfer type invertor. (b) Basic n-gate. (c) Basic p-gate. (d) Two level n-gate (e) Two level p-gate.

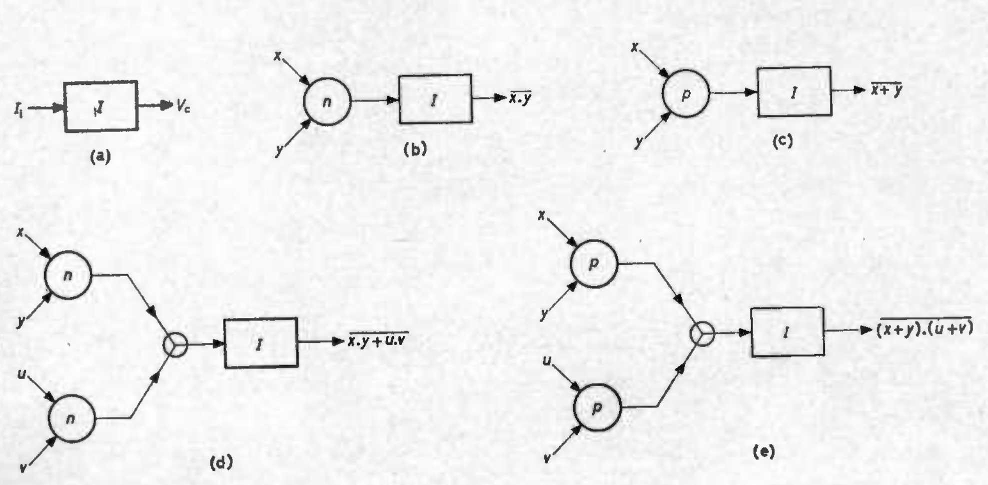
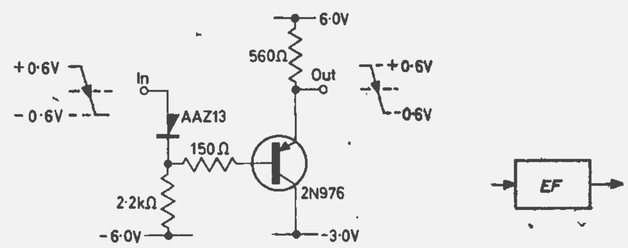


Fig. 3. Symbols for logic circuits

(a) Transfer type invertor. (b) NAND. (c) NOR. (d) NOT(AND-OR). (e) NOT(OR-AND)

(Indicated logical operations are valid only for negative logic)



(a) (b)

Fig. 4. Emitter follower
(a) Circuit. (b) Symbol

The emitter-follower in Fig. 4, which can be used to reduce the transient loading on the logic circuit, introduces a signal delay of less than 1nsec for fan-outs of up to 4.

Parallel Adder Design

The circuit arrangement in Fig. 5 shows stages k and $k+1$ of a parallel adder which has been designed to implement the logical statements in equations (4), (5) and (6). It makes extensive use of the two-level logic circuits described above and operates with negative logic.

The carry-path in stage k consists of a NOT(AND-OR) logic circuit which is based on equation (6) with r replaced by k. The input carry is c_{k-1} and the output carry is $\overline{c_k}$. In stage $k+1$ the carry-path consists of a NOT(OR-AND) logic circuit in accordance with equation (5) when r is replaced by $k+1$. The input carry for this stage is $\overline{c_k}$ and the output carry is c_{k+1}. Because the carry signal is generated by alternate stages as NOT-carry and carry respectively, it is necessary to design the adder in pairs of stages as indicated.

The sum generation in both stages k and $k+1$ is performed by NOT(OR-AND) logic circuits based on equation (4) with r replaced by k and $k+1$ as appropriate. Although the sum circuits in the two stages appear somewhat similar, the gate control functions are in fact interchanged due to the difference in the input carry signals. Emitter-followers are used to reduce the transient loading on the carry-path invertors.

Generation of the gate control functions ($x_r \not\equiv y_r$) and ($x_r \equiv y_r$) is accomplished readily using the appropriate two-level logic circuit when $\overline{x_r}$ and $\overline{y_r}$ are available in addition to x_r and y_r. For example, with suitable inputs, the NOT(AND-OR) circuit generates the function:

$$\overline{(x_r \, . y_r) + (\overline{x_r} \, . \, \overline{y_r})} = (x_r \not\equiv y_r)$$

Similarly the NOT(OR-AND) circuit generates the function ($x_r \equiv y_r$).

As a direct result of the simplicity of the gating requirements and speed of the transistor-diode logic circuits the setting and delay times for the adder are:

$$t_s \simeq 3{\cdot}5\text{nsec}, \ \delta_0 \simeq 2{\cdot}5\text{nsec}, \ \delta_s \simeq 8\text{nsec}.$$

Consequently, for words containing 16 bits, the sum generation time, in accordance with equation (1), is approximately 50nsec.

This adder has been designed to operate in conjunction with a 16 word × 16 bit tunnel diode-transistor memory[7] which has a non-destructive read cycle-time of 25nsec, and a write cycle-time of 50nsec. The prototype of this memory has been described previously[4].

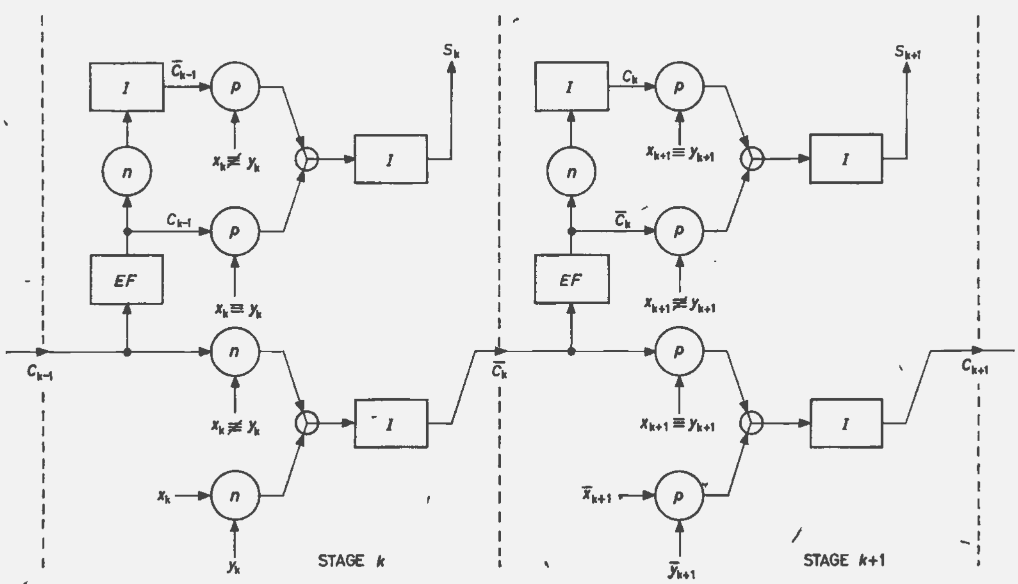


Fig. 5. Two successive stages of a parallel adder based on the transistor-diode logic circuit and using negative logic

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Worst Case Design of a Pulse Driver

By P. F. Jones*

A worst case design procedure for total excursion components is described in which the limiting values for each component are calculated directly from the limiting values of the initial performance parameters. Preferred value components are then chosen which lie within these calculated limits. The method formulates a set of inequalities from the performance parameters and solves these simultaneously to produce a second set which specify the limiting values for each individual component.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

THE existence of total excursion resistors, in which the total variation of value is known, makes more rigorous design methods possible. Instead of calculating single centre values, in this design, absolute limits for each component have been calculated directly from the limiting values of the parameters specifying the final performance. Components have then been chosen which lie within these calculated limits. The advantages of such a method are that it shows the range of values which may be used for a given component and leaves one free to choose a preferred value. In addition, these limits are worst case ones so that for any value within them the required performance will be achieved under all conditions. Furthermore, in choosing values one can see more clearly the effect of a choice from its closeness, or otherwise, to the calculated limits.

In general, the approach seems fundamentally more appropriate to the nature of design problems; in which the behaviour of a circuit and the components in it are both defined by limits rather than precise spot values.

The particular design described arose out of a need in Nimrod, the 7GeV proton synchroton, to send trigger pulses long distances over 100Ω lines. A simple circuit block was designed for manufacturing in quantity for general use in this capacity. Input levels and supply voltage were made compatible with the Mullard 100kc/s range of circuit blocks.

The performance specified was as follows:

Output

Pulse length:	10μsec
Frequency:	0 to 1kc/s
Rise time of front edge:	≯ 1μsec
Height:	+ 9 ± 1V
Load:	100Ω to ∞

Power Supply

12V ± 3%

Input

Height for full output:	≮ 4·8V
Threshold for no output:	2·1V

Circuit

The circuit configuration chosen is shown in Fig. 1. The input pulse to the base of VT_1 carries the emitter up, producing enough collector current in VT_1 to bottom VT_2. The standing potential at the emitter of VT_2 determines the height of the output pulse, while that at the emitter of VT_1 determines the threshold below which no output appears at all. C_1 smooths the current pulses drawn through VT_2 into a steady current proportional to the duty cycle. The values of R_1, R_2 and R_3 must be chosen correctly and to do this, the limits for their values were calculated from the performance limits set above.

* *Rutherford High Energy Laboratory.*

Theory

In Fig. 2.

V_1: emitter potential of VT_2

V_2: quiescent emitter potential of VT_1

V_s: supply voltage

R_1, R_2, R_3: nominal values

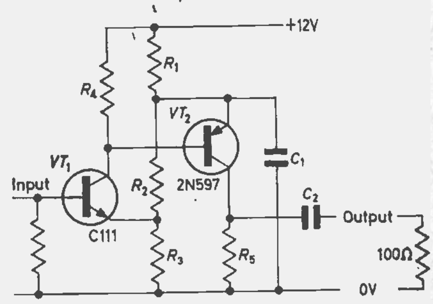
I : smoothed current pulses

I_p: VT_1 emitter current due to pulse

V_p: VT_1 emitter potential during input pulse

f: total excursion of R_1, R_2, R_3 from the nominal value (expressed as a fraction of the nominal value)

Now the output pulse height must be between the required limits. This means that V_1 must lie between an upper limit $V_{1(u)}$ and lower one $V_{1(L)}$.



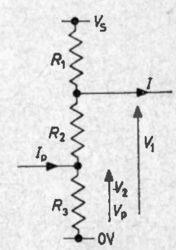


Fig. 1 (left). The circuit employed

Fig. 2 (right). Circuit for analysis

Analysing the circuit in Fig. 2:

$$V_1 = \frac{R_2 + R_3}{R_1 + R_2 + R_3} \cdot V_s - \frac{IR_1(R_2 + R_3)}{R_1 + R_2 + R_3}$$

where R_1, R_2 and R_3 are the nominal values.

V_1 is at a maximum when I is zero, R_1 is at the lower end of its tolerance band, and R_2, and R_3 are at the upper end. This maximum value must not exceed the upper limit $V_{1(u)}$.

Hence:

$$V_{1(u)} > \frac{(R_2 + R_3)(1 + f)V_{s(u)}}{R_1(1 - f) + (R_2 + R_3)(1 + f)} \quad \ldots\ldots (1)$$

V_1 is at a minimum when I is its maximum value $I_{(u)}$, R_1 is at its upper tolerance limit and R_2, R_3 at their lower limit.

Hence:

$$V_{1(L)} < \frac{(R_2+R_3)(1-f)V_{s(L)}}{R_1(1-f)+(R_2+R_3)(1-f)} - \frac{I_{(u)}R_1(R_2+R_3)(1+f)(1-f)}{R_1(1+f)+(R_2+R_3)(1-f)}$$

$$\therefore\ V_{1(L)} < \frac{(R_2+R_3)(1-f)V_{s(L)} - I_{(u)}R_1(R_2+R_3)(1+f)(1-f)}{R_1(1+f)+(R_2+R_3)(1-f)} \quad \ldots\ldots (2)$$

The next condition is that below a certain threshold there should be no output, i.e. VT_1 should not turn on. In other words the quiescent emitter potential of VT_1, VT_2 must always be above a certain level, $V_{2(L)}$.

Analysing the circuit in Fig. 2:

$$V_2 = \frac{R_3(V_s - IR_1)}{R_1 + R_2 + R_3} \quad \ldots\ldots (3)$$

The minimum value of V_2 occurs when R_1, R_2 are at the upper ends of their ranges, I is at its maximum and R_3 and V_s are at the lower ends of their ranges.

$$V_{2(L)} < \frac{R_3(1-f)\left[V_{s(L)} - I_{(u)}R_1(1+f)\right]}{(R_1+R_2)(1+f) + R_3(1-f)} \quad \ldots\ldots (4)$$

Finally there is the condition for full output with a certain minimum input. The base current of VT_2 is determined by the collector current of VT_1 which is approximately equal to its emitter current. Thus a certain minimum emitter current $I_{p(L)}$ must flow in VT_1, to ensure a bottomed output transistor VT_2. Considering I_p as the result of the difference between the quiescent and pulse emitter potentials across the resistance of R_2 and R_3 in parallel:

$$I_p = (V_p - V_2)\frac{R_2 + R_3}{R_2R_3}$$

and substituting for V_2 from equation (3):

$$I_p = \left[V_p - \frac{R_3(V_s - IR_1)}{R_1 + R_2 + R_3}\right]\frac{R_2 + R_3}{R_2R_3}$$

The minimum value of I_p occurs when R_1 is at the lower end of its range, R_3 at the upper, I is zero, and V_s is at its upper end. One cannot deduce the effect of R_2 because of the complexity of the function. However, since the variation is small it can be assumed that there is no peak in the middle of the variation. This means that the worst case will be either with R_2 at its maximum or at its minimum. Both cases will have to be examined.

$$I_{p(L)} < \left(V_p - \frac{R_3V_{s(u)}(1+f)}{R_1(1-f)+R_2(1\pm f)+R_3(1+f)}\right)\frac{R_2(1\pm f)+R_3(1+f)}{R_2R_3(1\pm f)(1+f)}$$

After a little re-arrangement and writing $1+f = 1/(1-f)$ and $(1-f)^2 = 1-2f$, assuming $f \ll 1$, the two conditions therefore become:

$$I_{p(L)} < \left(V_p - \frac{R_3V_{s(u)}}{R_1(1-2f)+R_3+R_2}\right)\frac{R_2+R_3}{R_3R_2(1+f)} \quad \ldots (5(a))$$

$$I_{p(L)} < \left(V_p - \frac{R_3V_{s(u)}}{R_1(1-2f)+R_3+R_2(1-2f)}\right)\frac{R_2+R_3}{R_3R_2(1-f)} \quad \ldots\ldots (5(b))$$

The set of inequalities (1) (2) (4) (5) can be transposed and solved to yield a second set of inequalities which give the limits for R_1, R_2 and R_3.

Transposing (1) and writing $1+f = 1/(1-f)$ and $(1-f)^2 = 1-2f$ assuming $f \ll 1$:

$$R_2 + R_3 < \frac{V_{1(u)}R_1(1-2f)}{V_{s(u)} - V_{1(u)}} \quad \ldots\ldots (6)$$

Similarly from (2):

$$R_2 + R_3 > \frac{V_{1(L)}R_1(1+2f)}{V_{s(L)} - V_{1(L)} - I_{(u)}R_1} \quad \ldots\ldots (7)$$

Therefore, from (6) and (7):

$$\frac{V_{1(u)}R_1(1-2f)}{V_{s(u)} - V_{1(u)}} > \frac{V_{1(L)}R_1(1+2f)}{V_{s(L)} - V_{1(L)} - I_{(u)}R_1}$$

Rearranging:

$$R_1 < \frac{V_{s(L)} - V_{1(L)}}{I_{(u)}} - \frac{V_{1(L)}(1+4f)(V_{s(u)} - V_{1(u)})}{I_{(u)}V_{1(u)}} \quad \ldots\ldots (8)$$

This is the upper limit for the nominal value of R_1. A value can be chosen and substituted into (6) and (7) to yield the upper and lower limits for $R_2 + R_3$.

Now transposing (4):

$$R_3 > \frac{V_{2(L)}(R_1+R_2)(1+2f)}{V_{s(L)} - V_{2(L)} - I_{(u)}R_1(1+f)} \quad \ldots\ldots (9)$$

and from (6)

$$R_3 < \frac{V_{1(u)}R_1(1-2f)}{V_{s(u)} - V_{1(u)}} - R_2$$

Therefore, from the above two expressions:

$$\frac{V_{2(L)}(R_1+R_2)(1+2f)}{V_{s(L)} - V_{2(L)} - I_{(u)}R_1(1+f)} < \frac{V_{1(u)}R_1(1-2f)}{V_{s(u)} - V_{1(u)}} - R_2$$

Rearranging:

$$R_2 < R_1 . \frac{V_{1(u)}(1-2f)\left[V_{s(L)} - V_{2(L)} - I_{(u)}R_1(1+f)\right] - V_{2(L)}(V_{s(u)} - V_{1(u)})(1+2f)}{\left[V_{s(L)} - V_{2(L)} + V_{2(L)}(1+2f) - I_{(u)}R_1(1+f)\right]\left[V_{s(u)} - V_{1(u)}\right]} \quad \ldots\ldots (10)$$

This is the upper limit for the nominal value of R_2.

Inequality (5) will not yield to the above treatment because of its quadratic nature and must be dealt with by trial and error.

Practical Design

The following values are to be substituted into the equations:

$V_{s(L)}$	: 11·6V	$V_{s(u)}$	: 12·4V
$V_{1(L)}$	: 9·5V	$V_{1(u)}$	: 11·0V
$I_{(u)}$	: 1mA	f	: 0·03
$V_{2(L)}$	: 1·3V	$V_{p(L)}$	: 4·0V
I_p	: 3·6mA		

The limits for V_1 are the result of taking into account the bottoming voltage of VT_2, and $V_{2(L)}$ and $V_{p(L)}$ the result of taking account of the base-emitter voltage of VT_1.

Substituting into (8) gives:

$$R_1 < 750\Omega$$

A value of 680Ω was chosen and substituted into (6), (7) and (10) to yield the following limits:

(a) $4·8\text{k}\Omega < R_2 + R_3 < 5\text{k}\Omega$ (6) and (7)

(b) $R_2 < 4·1\text{k}\Omega$ (10)

Choosing $R_2 = 3·9\text{k}\Omega$ and substituting into (9) gives:

$$R_3 > 0·84\text{k}\Omega$$

$R_3 = 1\text{k}\Omega$ was chosen. Evaluation of 5(a)) and (5(b)) gives 2·14 and 2·1 respectively. Since no other preferred values for R_2 and R_3 satisfy the conditions (a) and (b) a lower value of R_1 was chosen -560Ω. This gave the following conditions which permitted a wider range of values for R_2 and R_3.

(a) $3{\cdot}65\text{k}\Omega < R_2+R_3 < 4{\cdot}12\text{k}\Omega$

(b) $R_2 < 3{\cdot}4\text{k}\Omega$

Since a high input impedance was preferred, a value near the upper limit was chosen for R_2 : $3{\cdot}3\text{k}\Omega$

Substitution into (9) gives:

$$R_3 > 0{\cdot}59\text{k}\Omega$$

$R_3 = 680\Omega$ was chosen.

Evaluation of (5(a)) and (5(b)) gives 3·6 and 3·74mA respectively. This current determines the rise time of the output pulse and since a higher margin was preferred a value of 560Ω was finally chosen for R_3, giving a current of 4·8mA in the worst case.

Transistor Circuit Modules

Latest addition to the Mullard Ltd range of transistor circuit modules is a compact audio amplifier unit which gives a nominal output of 4W into a 12Ω loudspeaker. The module is currently being supplied to set makers for incorporation in mains-powered record players and other audio equipment.

Designed for operation from a 24V supply, the circuit employs a complementary push-pull output pair (AC128/176) mounted on an integral heat sink to allow operation up to 50°C. A thermistor is used in the circuit to achieve bias stabilization and the adoption of d.c. coupling between all stages ensures stability of the circuit against variations of voltage and temperature.

The module, type LP1162, has a typical input sensitivity of 85mV and can therefore be fed from the majority of gramophone pick-ups. Alternatively, the advantages of a high-value diode load can be realized when the module is used as part of a radio receiver.

Provision has been made for the incorporation of top cut and bass boost control circuits and suitable connexion points are provided.

This range of Mullard transistor circuit modules for a.m./f.m. record players and other audio equipment offers manufacturers the advantages of faster, easier production and greater freedom in cabinet styling.

Other advantages gained by using the modules are a saving in storage space, simpler stock control, and greater ease of servicing.

Brief details of other modules in the range, and which are shown in the accompanying illustration are as follows:

(Left to right, top row)

LP1164 Fully screened a.m./f.m. i.f. amplifier with integral a.m. mixer stage. Suitable for use in mains powered radio equipment and 12 volt car radios.

LP1167 This fully-screened module is for use with a 14 volt power supply and gives the necessary separation of the left and right channel information contained in the stereophonic signal. It is suitable for use with most types of ratio detector or similar circuits. The module has negligible insertion loss, very low noise and distortion, and may be left in circuit during normal monophonic reception.

LP1165 Fully screened a.m./f.m. i.f. amplifier with integral a.m. mixer stage. Suitable for battery powered radio receivers and 6 volt car radios.

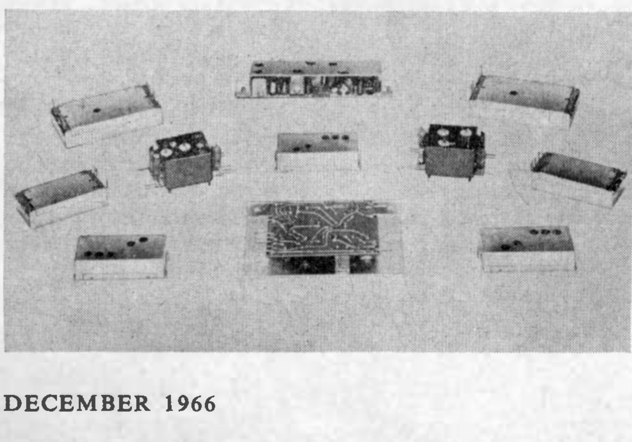
(Left to right, middle row)

LP1156 Fully-screened single-tuned i.f. amplifier and mixer stage for use in a.m. short, medium and long wave receivers.

LP1153 500mW audio amplifier designed for use in battery portable receivers and inexpensive record players.

LP1158 Fully-screened single-tuned i.f. amplifier and mixer oscillator stage for reception of medium and long wave signals.

The transistor circuit modules



Especially suitable for use with the Mullard AC1033 ganged tuning capacitor.

LP1172 1 watt audio amplifier designed for use in inexpensive record players and similar applications.

LP1159 Fully screened double-tuned i.f. amplifier and mixer stage for use in a.m. short, medium and long wave band receivers. Particularly suitable for high-performance receivers.

(Left to right, bottom row)

LP1166 Fully screened single-tuned i.f. amplifier and mixer stage for use in a.m. short, medium and long wave receivers. Particularly suitable for car radios.

LP1162 4 watt audio amplifier designed for use in mains powered record players and similar applications and described above.

LP1169 Fully screened single tuned i.f. amplifier and mixer stage for use in a.m. short, medium and long wave band receivers. Particularly suitable for car radios.

The New Leafield Radio Station

The most highly automated radio transmitting station in Britain—possibly the most advanced of its kind in the world—was recently inaugurated at Leafield, Oxfordshire.

This new radio station, which makes a significant contribution to the improvement of international communications, became generally operational in the spring of this year. Built and equipped by Post Office engineers, with the co-operation of British industry, at a cost of over £1M, it replaces the old Leafield Radio Station which, in its early years, played a pioneering role in the history of radio telegraph communications.

The building on an extended site of 287 acres adopts a three-wing layout for the transmitter hall. A two-storey design with transmitter cubicles at first-floor level incorporates a ground floor carrying all engineering and domestic supplies. A separate apparatus room houses the low-power equipment and a central control console. Offices, stores and welfare accommodation are grouped in a fourth wing.

The transmitters are of the self-tuning, self-loading type, employing conventional tuned-amplifier stages with motorized drives of variable capacitors and inductors controlled by phase discriminators. As appropriate tuning signals are applied to the transmitter, tuning and loading is completed automatically within two minutes using only five motor-drives. The wide use of solid state (silicon diode) h.t. rectifiers in place of mercury vapour types greatly improves reliability.

Twelve 30kW transmitters, for fixed service operation, are in two self-contained groups in two of the wings. A third wing contains the six 85kW transmitters which are normally used for press broadcast traffic.

Carrier generation is carried out by synthesizers which, controlled from a master-oscillator, maintain the radiated carrier frequencies accurate to one part in 10^7. The association of a synthesizer with each transmitter is an added factor in securing flexibility. Each synthesizer frequency is selected remotely according to a prearranged programme by motorized switches. Fifty frequencies are programmed and selection of any one initiates the automatic tuning processes in the transmitter. For the first time at a commercial station, the crystals and their transistor-maintaining oscillators of the 100kc/s master-oscillator system are sunk in sealed containers down shafts 30ft below ground as an alternative to using conventional ovens. This introduces conditions almost ideal for precision oscillators.

Concentric tiered rhombics for the fixed services, wideband log-periodic aerials for the press services and standby purposes, and triple-tiered rhombics erected on 300ft stayed masts have been introduced. The log-periodic aerial is a comparative newcomer in the h.f. field and is used by the Post Office for the first time at Leafield.

A Binary-Quinary Decade Counter Using Resistance Logic

By R. Parshad* and S. P. Suri*

This article presents a binary-quinary decade counter using resistance logic. The present circuit uses fewer transistors than the conventional. The digital read-out is simple and economical in the use of components. Bi-directional counting can also easily be achieved with the circuit.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

CONVENTIONAL counters using four bistable stages and employing various decade conversion techniques[1,2,3,4,5] have been made in the past. All these methods of decimal counting perform an operation equivalent to internally advancing the count by six during the application of ten input pulses.

The present logic counter using binary-quinary code is an attempt towards using fewer components to achieve decimal counting. The use of binary-quinary code for the counter ensures more economical read-out logic and the counting is straightforward.

Principle

The counter discussed is based on the binary-quinary self-checking code used in a computer. This code has the uniqueness of having two one's in each digit as is shown in Table 1. A decimal counter based on this code uses a binary and a quinary which in fact is a five bit register occupying the same bit position for two successive input pulses and registering the next higher count only when the binary changes its state from '1' to '0'. It is evident that the decimal counting proceeds in 0102468 code instead of 1242 binary code as in a conventional b.c.d. counter.

Circuit Description and Operation

Fig. 1 shows the complete binary-quinary counter. Transistors VT_1, VT_2 constitute the conventional flip-flop and the quinary is built using transistors VT_3 to VT_7. Each quinary element is essentially a resistor-transistor NOR logic unit[6] consisting of a resistor OR gate (R_1 and four resistors R) followed by an invertor stage. If any of the inputs to the OR gate resistance is at '1', the base current which the invertor transistor draws through it is sufficient to bottom it with the logic '0' output. When all the inputs are '0', the output of the r.t.l. is logic '1'.

Now in the counter operation the quinary has to be arranged such that when one of its transistors is non conducting, the remaining four should be conducting. For this purpose, each quinary element has the four inputs of the OR circuit described above connected through resistors to the collectors of each of the other quinary elements. Again, each quinary element in conjunction with the common bistable element has associated with it one of the diode AND gates (G_1, G_2 - - - G_5) comprised of diodes D_1, D_2 and resistor R_3.

The mode of operation of the counter is as follows:

To start with, let the binary be in its '0' state (VT_1 non-conducting and VT_2 conducting) and the off transistor VT_3 holding the rest of the quinary transistors 'on'. At count one, with the transition of the binary from '0' to '1' state, the output of the diode AND gate G_1 is at a lower level of approximately 6V. At the second input pulse, the binary changes from '1' to '0' state and the output voltage of gate G_1 jumps to a higher level of approximately 0V. The positive going pulse thus obtained through C triggers VT_4 of the quinary to the '1' state. Due to the mutual coupling of the quinary transistors by the OR circuit mentioned earlier VT_3 at this instant becomes conducting. Fig. 2 depicts the switching cycle of the counter.

TABLE 1

DECIMAL	BIT WEIGHT 0	1	0	2	4	6	8
0	1	0	1	0	0	0	0
1	0	1	1	0	0	0	0
2	1	0	0	1	0	0	0
3	0	1	0	1	0	0	0
4	1	0	0	0	1	0	0
5	0	1	0	0	1	0	0
6	1	0	0	0	0	1	0
7	0	1	0	0	0	1	0
8	1	0	0	0	0	0	1
9	0	1	0	0	0	0	1

Thus each transistor in the quinary is sequentially triggered at the even numbered input pulses and the whole system works as a decade counter according to the assigned binary quinary weights.

Reversible Counting

The circuit discussed above can readily be used for reverse counting. For this purpose it is merely necessary to use another set of diode AND gates so arranged that the output of any AND gate associated with a binary transistor routes the signal to its preceding transistor, instead of to the succeeding one as required for forward counting.

Digital Read-Out

The digital read-out used in the counter employs filament

* *National Physical Laboratory, New Delhi, India.*

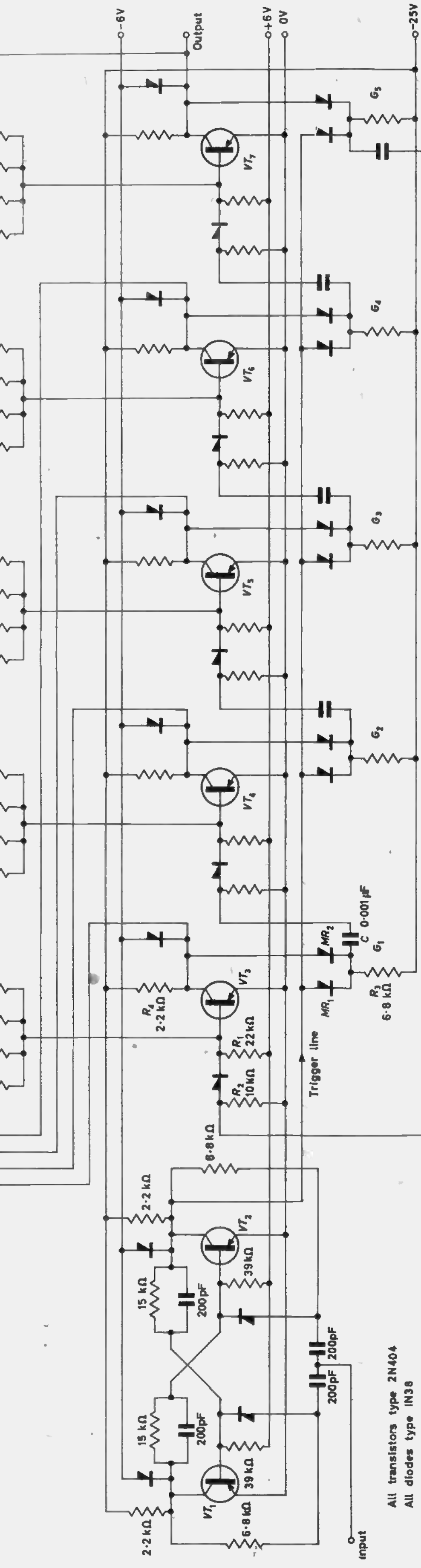


Fig. 1. Binary-quinary decade counter

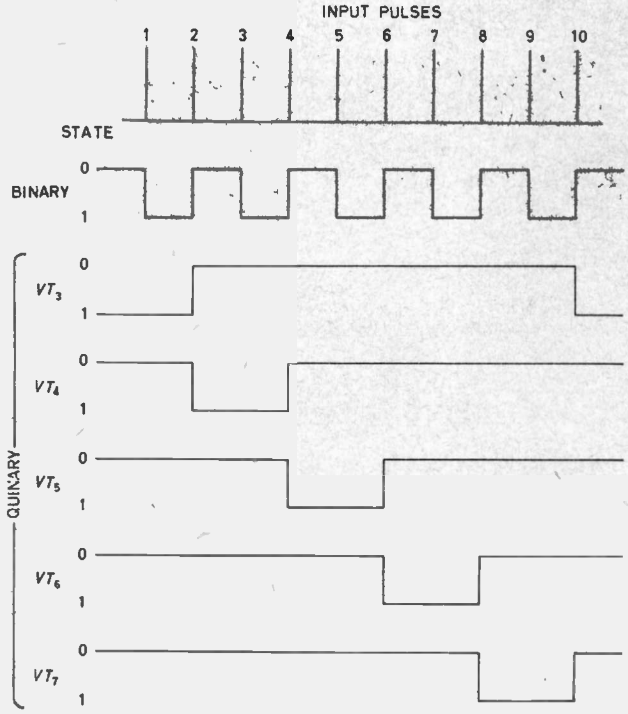


Fig. 2. Waveforms of counter

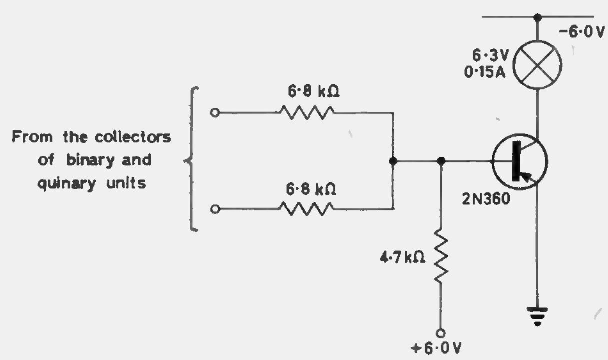


Fig. 3. Digital read-out circuit

lamps and is simple and economical, avoiding the use of a diode AND matrix and associated driving transistors. Fig. 3 shows one of the ten read-out units for the complete decade. Inputs to the two base resistors are taken from one of the quinary elements and one of the transistors of the binary stage.

It has been arranged that with one of the inputs to the base resistors at lower level (approximately 6·0V), the read-out transistor is kept reverse biased. With both the inputs to these resistors at the lower level, i.e. in the '1' state which defines a particular bit position of the counter as shown by Table 1, the read-out transistor will be driven to light up a lamp in its collector circuit.

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A Very High Input Impedance D.C. Chopper Amplifier Using a Field Effect Transistor

By R. Verrill*, B.Sc.

A new field effect transistor type EXP380 which has been developed is characterized by very low gate leakage current of the order of 10^{-10}A at 25°C and very low gate capacitance of about 2·5pF. The device has proved very successful in an experimental series/shunt chopper d.c. amplifier where low gate leakage and capacitance are essential. The amplifier described has an input resistance of over 100MΩ and the full scale sensitivity is ±1mV giving ±100μA in the moving-coil instrument at the output. Input offset drift with temperature is about 1μV/°C and 15pA/°C without the necessity for matching of transistors.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

THE main advantage in using a chopper amplifier for low level d.c. amplification is the very low drift which can be obtained. There are basically two kinds of chopper amplifier in common use for the amplification of very low d.c. or slowly varying voltages, namely the shunt chopper and the series/shunt chopper. The shunt chopper is used when a relatively low input impedance amplifier, up to about 20kΩ is required, while an input impedance of hundreds of meghoms can be obtained using the series/shunt chopper followed by a high input impedance a.c. amplifier. In both arrangements the voltage to be amplified is modulated by the chopper so that it can be amplified by an a.c. amplifier to give an a.c. output proportional to the input voltage. This a.c. output is then synchronously demodulated to give a d.c. output. The arrangement of the series shunt chopper is shown in Fig. 1. The two switches S_{1a} and S_{1b} successively connect the a.c. amplifier input to the input terminal and earth via the capacitor C_1. Thus there is no loss in gain through the chopper and the input impedance is equal to $4R_{in}$ where R_{in} is the input resistance of the a.c. amplifier. If there is any overlap in the on periods of the two switches, and this is difficult to avoid in solid state switches, then the maximum obtainable input impedance is modified by this effective impedance produced by the overlapping action.

At low drain source voltages the field effect transistor operates like a variable resistor controlled by the gate voltage. In the case of the EXP380 the value of this resistor varies between a few thousand ohms at zero gate voltage to an extremely high value when the gate voltage is larger than the 'pinch off' voltage. In the low resistance, on, condition there is no pedestal voltage such as occurs with bipolar transistors and the device behaves almost as a perfect switch in series with a resistance. There is, however, a small offset current caused by gate leakage current flowing when the f.e.t. channel is in the high resistance state. This current causes a small error voltage when flowing through the resistance of the circuit connected to the drain or source of the f.e.t. In the case of the series/shunt chopper this resistance is that of the external circuit where the voltage is to be measured. Since only half the gate current, about 5×10^{-11}A, would flow in the external circuit the error voltage in a circuit of resistance as high as 1MΩ would be only about 50μV at 25°C. A more noticeable effect, however, is capacitive coupling of the gate square wave by the gate drain capacitance. The capacitive current is demodulated by the two switches to produce a net d.c. in the external circuit. Assuming a capacitance of 1·2pF and a square wave of 6V at 1kc/s this current would be $7{\cdot}2 \times 10^{-9}$A but the currents from the two f.e.t.'s are in opposite directions and they thus tend to cancel. Any difference in these two capacitive currents will cause an offset current with a slight temperature dependence as the capacitance varies slightly with temperature. A method

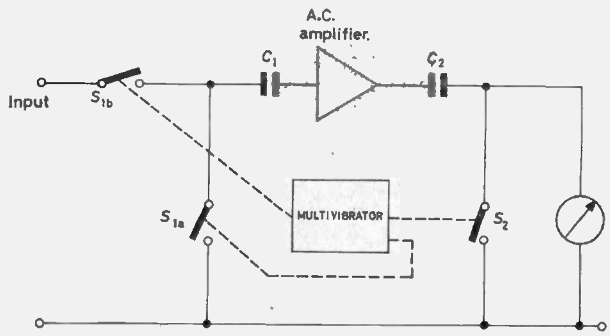


Fig. 1. Arrangement of series-shunt chopper amplifier

for neutralizing the offset due to leakage and capacitance will be described when discussing the actual circuit of the amplifier.

The complete circuit of the chopper amplifier is shown in Fig. 2. A cross coupled multivibrator drives both gates of the series and shunt f.e.t.'s and also the demodulating transistor VT_7 at the output. The diodes MR_4 and MR_{11} ensure that the gate voltage is exactly at earth potential when the chopping transistor is on and thus the minimum on resistance is obtained without taking gate current. The output from the chopper is followed by a high gain amplifier consisting of transistors VT_3, VT_4, VT_5 and VT_6. The Zener diodes MR_1, MR_2 and MR_3 used instead of decoupling capacitors produce greater stability particularly at low frequencies. The very high input impedance is achieved by using a field effect transistor with feedback on the input stage of the a.c. amplifier and also by overall d.c. negative feedback which is coupled into the source lead of the shunt chopping transistor. The potentiometer, R_{24}, alters the amount of d.c. feedback and thus serves as a gain control. The amount of feedback for 1mV full-scale deflexion is about 20dB.

The offset current due to capacitance and leakage is neutralized by C_3 a capacitance of 5pF connected between the chopper output and the waveform at one of the collectors of the multivibrator attenuated by the potentiometer R_{22} and R_{23}. The control should be adjusted for zero output when a large resistance, e.g. 1MΩ, is connected across the input after first adjusting the zero control R_{21} for zero output with the input shorted. This is probably the simplest method of reducing the offset to

* *Ferranti Ltd.*

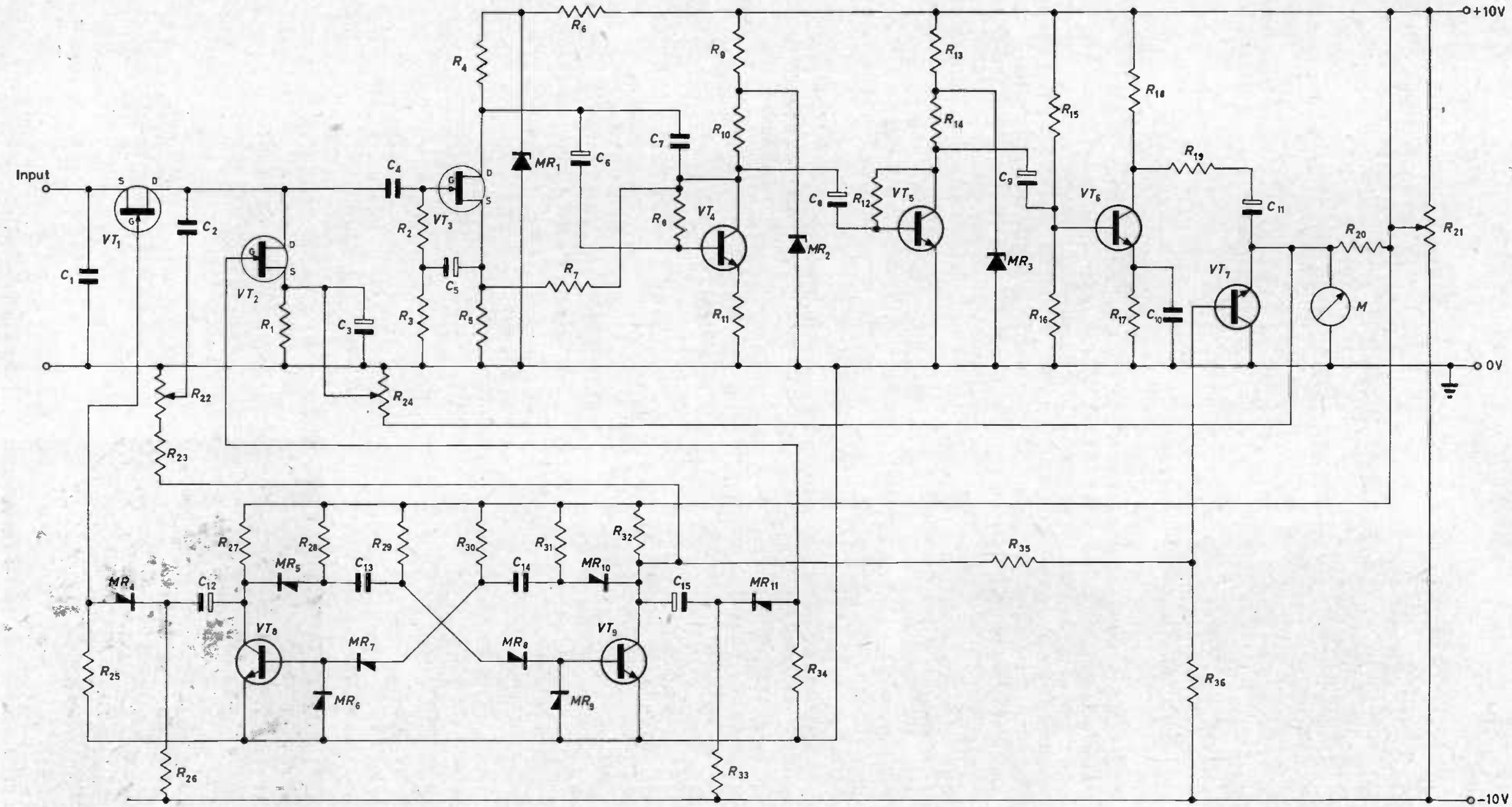


Fig. 2. Series-shunt chopper amplifier giving $\pm 100\mu A$ output for $\pm 1mV$ input at over $100m\Omega$ input impedance

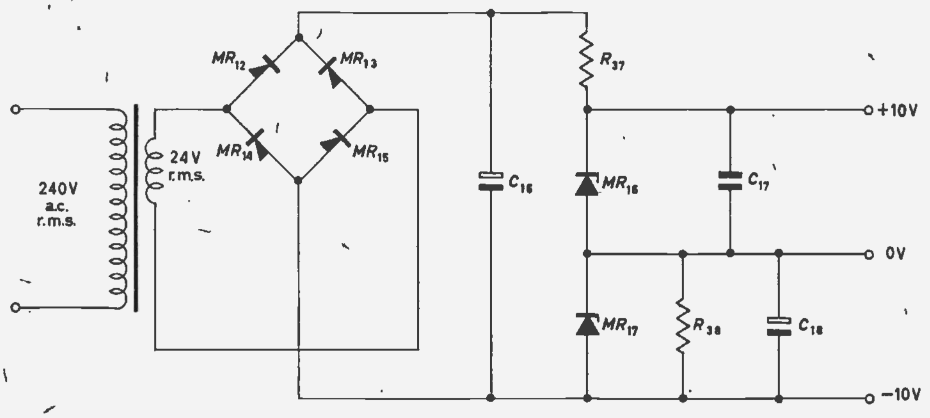


Fig. 3. Power supply for chopper amplifier

zero at one temperature and the drift with temperature will not be more than about 1μV/°C and 15pA/°C. An even smaller current drift could be obtained by using a proportion of the current from a reverse biased diode to compensate for the gate leakage current but this would involve a rather complicated setting up procedure with adjustments at two temperatures.

Metal earthed screening is necessary around the chopping stage and the input stage of the a.c. amplifier to prevent pick-up from the multivibrator and from the mains field of the power supply transformer. This can be done by mounting the chopper and a.c. amplifier stages inside an earthed metal box.

The circuit of a Zener diode regulated power supply suitable for the amplifier is shown in Fig. 3.

The main advantages of the amplifier are as follows. The very high input impedance means that the amplifier can be used for measurements in high resistance circuits up to about 1MΩ with negligible loading and negligible loss of sensitivity. Also measurements of very small currents as low as 10^{-9}A full-scale deflexion or less may be made. A universal shunt can be used at the input to provide any desired voltage or current range down to 1mV and 10^{-9}A full-scale deflexion. Another advantage is the simplicity of the circuit in that no transformer is needed for driving the chopper switches as is the case with most other forms of series/shunt chopper.

Typical Characteristics of the EXP380 Ferranti Field Effect Transistor

- I_{DO} Drain source current at 25°C with zero gate source voltage 0·25mA
- V_p Pinch off voltage −2·5V
- G_m Mutual conductance at 25°C, 0·15mA/V
- R_{sat} On resistance at 25°C, 8kΩ
- I_{GS} Gate to source and drain leakage current at 25°C, 10^{-10}A
- C_g Total gate capacitance at 25°C, 2·5pF

Fig. 4. Drain source characteristic

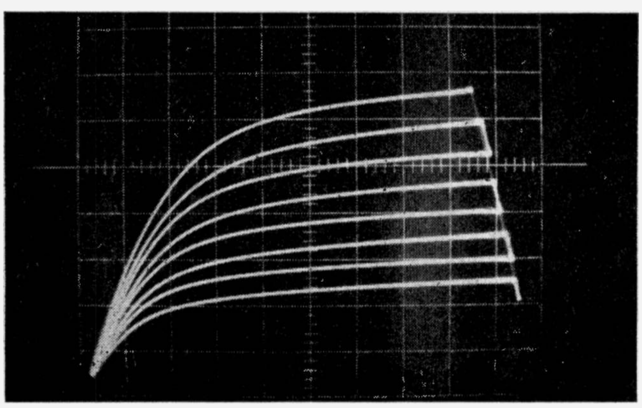


Fig. 4 shows a photograph of the drain source characteristics of an EXP380 for gate source voltages of 0V, −0·2, −0·4, −0·6, −0·8V, −1·0V, −1·2V and −1·4V. The scale of the vertical axis is 0·05mA per division and that of the horizontal axis 1V per division. Values of I_{DO}, V_p, G_m and R_{sat} similar to the above can be determined from these curves.

The part of the characteristic near the origin is that of interest in switching applications. The drain source resistance can be seen to be quite linear until about 1V is reached. When the gate drain voltage is greater than V_p about −2·5V, further increase in drain source voltage causes little increase in drain current. This part of the characteristic is suitable for linear amplification and because of the low gate capacitance and leakage the transistor can be used in very high impedance amplifiers up to fairly high frequencies.

Component Values

VT_1, VT_2	EXP380 Ferranti field effect transistors
VT_3	ZFT12 Ferranti field effect transistor
VT_4 to VT_9	ZT87 Ferranti silicon npn transistors
MR_1 to MR_3	KS39A Ferranti Zener diodes
MR_4	ZS130 Ferranti silicon diode
MR_5 to MR_{10}	ZS120 Ferranti silicon diodes
MR_{11}	ZS130 Ferranti silicon diode
MR_{12} to MR_{15}	ZS70 Ferranti silicon diodes
MR_{16}, MR_{17}	KS42A Ferranti Zener diodes

All resistors are ¼W rating unless otherwise stated.

R_1	100Ω	R_{14}	2·2kΩ	R_{27}	1·5kΩ
R_2	10MΩ	R_{15}	27kΩ	R_{28}	3·3kΩ
R_3	82kΩ	R_{16}	6·8kΩ	R_{29}	33kΩ
R_4	8·2kΩ	R_{17}	220Ω	R_{30}	33kΩ
R_5	390Ω	R_{18}	1kΩ	R_{31}	3·3kΩ
R_6	470Ω	R_{19}	1kΩ	R_{32}	1·5kΩ
R_7	3·3kΩ	R_{20}	33kΩ	R_{33}	150kΩ
R_8	150kΩ	R_{21}	10kΩ	R_{34}	47kΩ
R_9	390Ω	R_{22}	10kΩ	R_{35}	5·6kΩ
R_{10}	2·2kΩ	R_{23}	47kΩ	R_{36}	68kΩ
R_{11}	39Ω	R_{24}	10kΩ	R_{37}	270Ω 1W
R_{12}	150kΩ	R_{25}	47kΩ	R_{38}	390Ω ½W
R_{13}	390Ω	R_{26}	150kΩ		

C_1	0·5μF low leakage	C_{10}	250μF
C_2	5pF	C_{11}	12μF
C_3	500μF	C_{12}	0·5μF
C_4	6 800pF	$C_{13,14}$	0·02μF
$C_{5,6}$	12μF	C_{14}	0·02μF
C_6	12μF	C_{15}	0·5μF
C_7	150pF	$C_{16\text{-}18}$	500μF 50V
$C_{8,9}$	12μF	C_{17}	500μF 12V
C_9	12μF	C_{18}	500μF 12V

M −100μA, 0, 100μA centre-zero moving-coil instrument Resistance 450Ω

An Air Temperature Digitizer—Range 0°F-99°F

By W. V. Dromgoole*

This instrument was developed to provide a conversion from the air temperature to a digital form suitable for a tens and units lamp indication system.

The lamp display is part of a system which shows the time of day for 47sec, and the air temperature for 10sec, every minute and is visible over a large area of the towns where the equipment is installed.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 835)

A DEGUSSA air temperature thermometer type N2-nickel, which has a resistance of 100Ω at 32°F is used as the sensing element. This thermometer is part of a bridge network having equal top arms of 100Ω and the remaining lower arm is wound to have the same resistance as the thermometer at 0°F (i.e. 90·5Ω for the particular one used.)

The bridge is therefore balanced at 0°F and any change of temperature above 0°F unbalances the bridge in the same phase direction.

The bridge output, together with a re-balance voltage, is amplified, and controls via two thyratrons, the movement of two G.E.C. bothway uniselectors designated 'tens' and 'units' which function both as balancing potentiometers and lamp display directors.

A re-balancing voltage, which is opposite in phase to the bridge output voltage is applied to the main bridge from the position of the 'tens' and 'units' such that, at balance, no voltage appears at the input to the amplifier. When this occurs no further movement of the uniselectors takes place, and the lamp output wires signal the actual temperature in tens and units.

This type of bridge and balancing circuit was used because it allowed a relatively large change of resistance in the balancing circuit, while the change in the thermometer resistance is small (i.e. approximately 0·3Ω/°F). This allowed the use of uniselector arcs and wipers as the balancing potentiometers, while providing a convenient circuit for the lamp display.

Some slight non-linearity occurs since the portion of the bridge which includes the temperature sensing device does not function as a null bridge, however a resistance-temperature graph of the operation of the complete bridge and balancing circuit showed that over the range used the non-linearity is insignificant for the purpose for which the circuit was designed.

Fundamental Balancing Circuit (Fig. 1)

The bridge is balanced at the resistance value of the Degussa air temperature thermometer at 0°F (i.e. 90·5Ω) so that any increase in temperature above 0°F causes the bridge output voltage to increase in amplitude without phase reversal.

The bridge is fed from a screened winding of 2·5V on transformer T_4, and a separately screened winding of 6·3V feeds a tens and units potentiometer. This voltage which is 180° out of phase with the bridge output voltage, is selected as follows: Assume that at 0°F the tens and units uniselectors are at position 0·0. In this position no voltage appears at the grid of V_1 since the main bridge is also at balance.

Now assume that there is an increase in temperature of 5°. A voltage from the main bridge appears at the grid of V_1 which is amplified and causes the units uniselector to step from 0 towards 9, but at position 5 equal and oppositely phased voltages appear at the grid of V_1 and balance is achieved.

If the temperature now rises say, to 22°F the units selector will first move to position 9 and the tens uni-

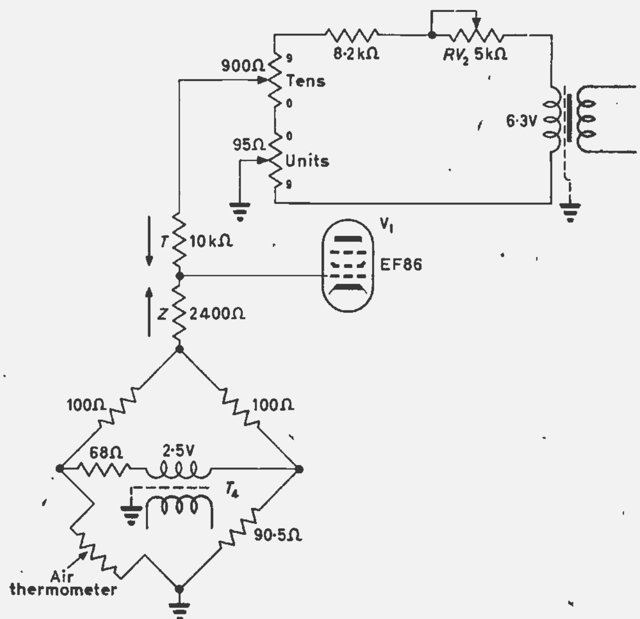


Fig. 1. Fundamental balancing circuit

selector now steps from 0 until the voltage at the grid of V_1 changes phase. The tens uniselector stops in this position (i.e. 2), but the units uniselector now steps backwards from 5 towards 0, and balance will now be found at position 2, when the units uniselector will stop. Thus the tens and units uniselectors are stationary at 2.2.

If the temperature is at some higher value and drops, the units selector first moves to 0, and the tens uniselector then moves towards 0 until a phase reversal occurs at the grid of V_1. The units uniselector then moves upward from 0 until a balance is obtained again.

In the circuit described the tens uniselector potentiometer consists of 9 sections each of 100Ω, and the units uniselector potentiometer of 19 sections of 5Ω. This allows resolution to 0·5°F, but since the indication required is to 1°F, any backlash in the system is adequately provided for.

Resistor T (10kΩ) prevents the grid of V_1 becoming grounded if the tens and units potentiometers are stand-

* *University of Canterbury, New Zealand.*

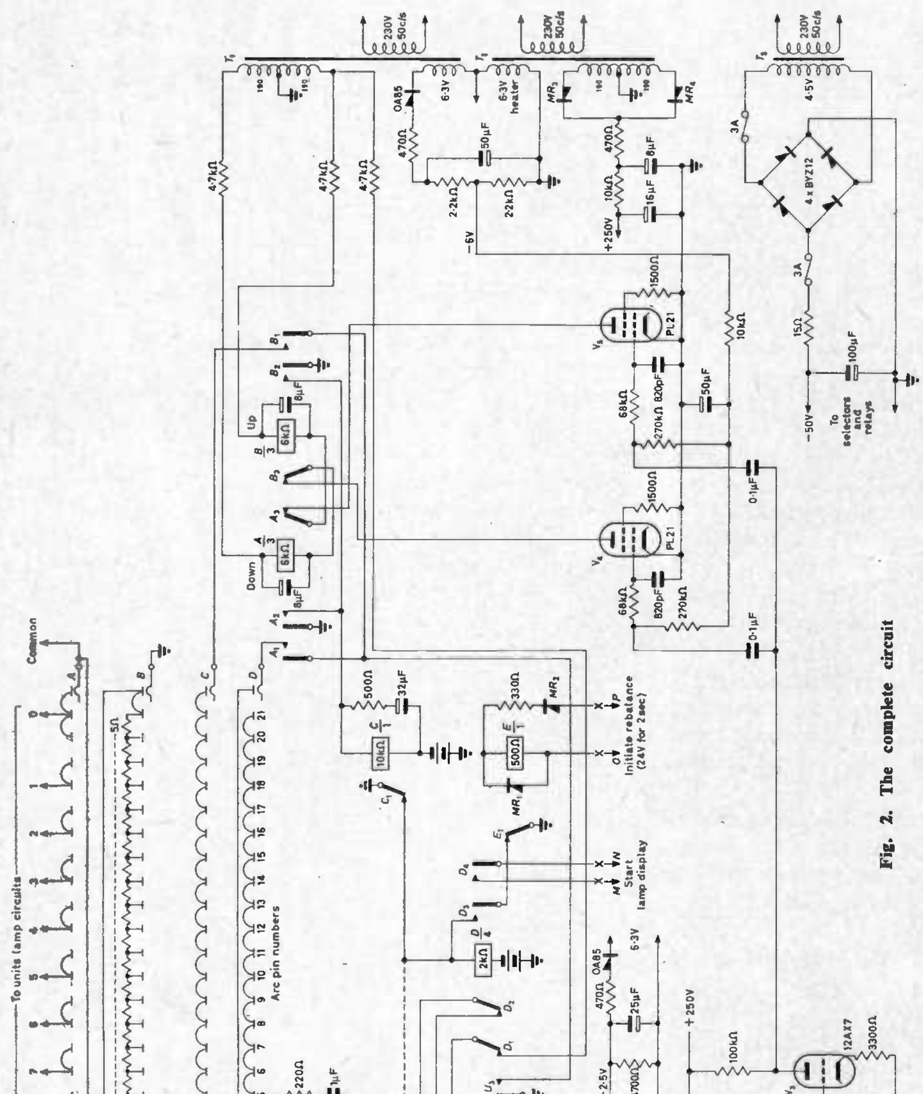


Fig. 2. The complete circuit

ing at position 0.0 which would prevent any output from the main bridge being amplified. Resistor Z (2·4kΩ) is chosen such that a change of 1°F causes a change in bridge output equal to a change of 10Ω in the units balancing potentiometer. The voltage existing across the tens and units potentiometer is preset by RV_2. This procedure is mentioned later.

Operation (Fig. 2)

The input signal to V_1 is amplified by V_1, V_2 and V_3, and fed to the grids of two thyratrons (V_4, V_5) the grids of which are held at −6V d.c. The anodes of V_1 and V_5 are fed with 190V a.c. (oppositely phased) via 4·7kΩ resistors and relay coils A and B. Interlocking between relays A and B is provided by connecting each of these relay coils to the anodes of V_4 and V_5 via the normally closed contact of the other relay.

Now assume that both the tens and units uniselectors are at position 0.0 (i.e. tens on arc pin 18 and units on arc pin 22). There will therefore be no voltage from the tens and units balancing potentiometers, and the only voltage at the grid of V_1 will be that due to the main bridge output. The phase of the voltage reaching V_4 and V_5 will therefore be such as to cause V_5 to conduct and operate relay B, which opens relay A circuit at B_3, and closes an impulsing ground from U_3 (this selector stepping circuit is described later) via relay B_1, wiper C and arc of units uniselector to 'units up' magnet. Wiper B changes the units potentiometer in such a direction that a voltage 180° out of phase with the bridge output voltage is now applied to the grid of V_1, and if the actual temperature lies between 0°F and 9°F the units uniselector will attain balance at some position in its travel, V_5 will become non-conducting and relay B will release.

Now say the temperature is 25°F. The units uniselector will step upwards from arc pin 22 (0°F) to pin 1 (9°F) and stop there, but the impulsing ground on wiper C is now extended to wiper C of the tens uniselector, arc pin 18 to the 'tens up' magnet. The tens uniselector will take two steps. The balancing voltage is now greater than the bridge output voltage and reversed in phase, therefore relay B releases and since thyratron V_4 now conducts, relay A operates and closes the impulsing ground via A_1 wiper D and arc pin 1 to the 'units down' magnet, and the units uniselector steps down to arc pin 12 or 13 (i.e. 5°F position), when balance is again achieved. The tens and units uniselectors are now standing at arc pins 16 and 12 or 13 respectively, and lamp display will be given for 25 via the A wipers.

Circuit Functions of Relays *C*, *D*, *E*

Consider that the equipment has just been switched on. Relay D operates immediately via normally closed contact C_1 to ground, and locks via normally open contact D_3 to ground on the normally closed contact E_1 to ground. Relay D opens its normally closed contacts D_1 and D_2, preventing the selector stepping circuit from functioning.

Now when the valves have warmed up, the amplified output from the bridge will cause either V_4 or V_5 to conduct, and therefore separate either relay A or B. A ground is now extended via the normally open contact A_2 or B_2 to relay C which operates. This does not affect relay D (which is already operated) in the meantime. A circuit is prepared to step either the 'units up' or the 'units down' magnet via the C and D wipers of the units selector via the normally open contacts B_1 and A_1 respectively to contact U_3.

When the auxiliary lamp display equipment signals a 24V a.c. pulse for 2sec on wires O and P, relay E operates and releases relay D at contact E_1. (Relay D cannot reoperate since relay C is energized). Contacts D_1 and D_2 now cause the selector stepping circuit to function and relay U impulses at about a half-second rate, and the impulsing ground at U_3 now steps the uniselectors via the previously prepared circuit, and when balance is achieved relays A and B will be both denergized removing ground at contacts R_2 and B_2, and relay C denergizes, and operates relay D, which opens the impulsing circuit at contacts D_1 and D_2, while relay D locks again to normally closed contact E_1.

Relay D closes a circuit via its normally open contact D_4 to inform the auxiliary lamp equipment that it may display. After a minute has elapsed, relay E is again energized for two seconds, but if the temperature has not changed relay C will still be denergized and therefore relay D will be operated, and no impulsing will be present. The cycle continues. Relay C is made slow release by shunting its coil with a 32μF capacitor. This is done to prevent any release of relay C while balancing is being made which may involve a change of phase when there will be a short period as say, relay A releases and before B can operate.

Selector Stepping Circuit

Pulses for stepping the uniselectors are provided by V_6 and relay U. The pulses are approximately half a second duration with the 'on' period (i.e. relay U operated) being short compared with the 'off' period. Consider the quiescent condition with relay D energized.

The grid of V_6 is at $-2{\cdot}5$V d.c. supplied by an OA85 and filter circuit. V_6 is therefore non-conducting. When relay D releases, a circuit is closed; 190V a.c. via a 4·7kΩ resistor, normally closed contact D_1 the coil of relay U to anode of V_6. A second circuit is also closed: a positive voltage is fed from the junction of the 3·3MΩ and 390kΩ resistors, via 680kΩ, and normally closed contact D_2, normally closed contact U_1 to a 2μF capacitor and the grid circuit of V_6. The 2μF capacitor is charged until the grid voltage of V_6 is sufficiently raised to cause conduction. Relay U operates and passes out a ground pulse at contact U_3. Contact U_1 opens the charging circuit to the capacitor and transfers the capacitor to contact U_2, and hence via a 10kΩ resistor to the negative 2·5V line, so that the grid of V_6 is rapidly reduced to $-2{\cdot}5$V again. V_6 becomes non-conductive, but relay U remains energized for a short time by virtue of the 8μF capacitor connected across its winding. The release of relay U restarts the cycle.

The positive charging voltage is fed via contact D_2 to ensure that the first impulsing cycle is the same as for subsequent cycles. (i.e. the 2μF capacitor is discharged at the grid voltage level of $-2{\cdot}5$V).

Temperatures Above 99°F or Below 0°F

It is sufficient for this circuit to limit the indication to 99°F or 0°. This is done by a connexion from tens C arc pin 7 and D arc pin 18 to relay D, so that if the tens uniselector moves to either of these positions relay D operates and opens the stepping circuit.

Uniselectors

The uniselectors used are G.E.C. 'bothway' types with bridging wipers and gold-plated wipers and arc pins to minimize changes in resistance. They were supplied for 50V operation, which voltage was convenient for relay operation as well.

Setting Up

(a) The Degussa air temperature thermometer resistance is near 100Ω at 32°F. However its *actual* resistance is measured at 32°F and at several values above this temperature. A temperature-resistance graph is plotted and extended backwards to 0°F. The 0°F value of this resistance is now placed in the opposite lower arm of the bridge.

(b) A resistance box in now connected in the bridge in place of the air thermometer and set at the resistance value of the air thermometer at say 75°F.

(c) Relay D is operated to prevent impulsing, and the tens and units uniselectors are set manually to arc posisions 10 (i.e. No. 7 lamp), and 12 or 13 (i.e. No. 5 lamp), respectively. Now adjust potentiometer RV_2 so that both relays A and B release, indicating that the balancing voltage is equal and opposite in phase to the bridge output voltage. This adjustment will now hold for other temperatures.

(d) The amplifier gain control RV_1 should be set so that no hunting occurs between relays A and B when the units uniselector has reached balance.

Mounting Air Thermometer

The thermometer is mounted on the south side (away from the sun in New Zealand) of the building and is housed in several truncated cone-shaped shields to avoid any direct radiation.

The instrument has been in use for over a year and has given no trouble. The display of time and temperature has evoked much public interest.

Acknowledgments

The close co-operation of the Director, Mr. T. H. Scott is much appreciated and this article is published with his permission.

A Transistor RC Oscillator Using Negative Impedances

By S. Pasupathy*

A generalized form of RC oscillator circuit using negative impedances, of which the Wien bridge type of oscillators are shown to be special cases, is described. A direct synthesis of this network with a negative impedance convertor results in a novel transistor oscillator. The oscillator circuit, its two primary modes of operation and some special features are discussed.

(Voir page 831 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 838)

THIS article is concerned with a new type of transistor oscillator circuit, developed from the concept of negative impedance. This circuit has certain similarities with some existing bridge types of *RC* oscillators, but has the advantages of using fewer components, having the common point of the tuning capacitor at ground potential and of achieving very low frequencies of oscillation.

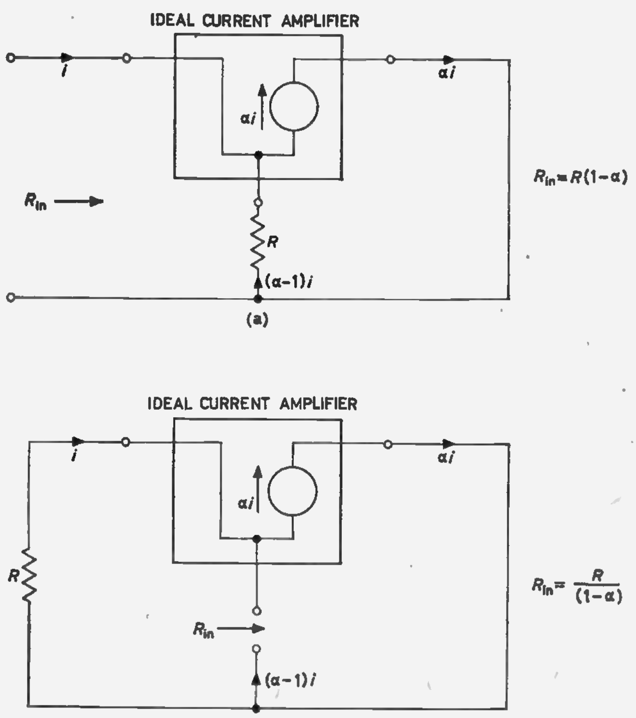


Fig. 1. Ideal current amplifier connected to produce
(a) Open circuit stable negative input resistance
(b) Short circuit stable negative input resistance
(current gain α is assumed to be greater than unity)

Generalized Form of RC Oscillator Circuit

It is general practice to analyse *RC* oscillators from the view-point of feedback. However, in many circuits, positive feedback produces the effects of negative impedance; for example, Fig. 1 shows how positive feedback using an ideal current amplifier produces open-circuit stable and short-circuit stable negative input resistances. Hence *RC* oscillators can also be analysed using the concept of negative impedance.

The current derived *RC* oscillator[1], shown in a simplified form in Fig. 2(a), can be taken as an example. It is known that the output and input currents, i_1 and i_2, are in phase at a frequency:

$$f = 1/2\pi RC \quad (1)$$

* *Indian Institute of Technology, Madras.*

and:

$$i_1/i_2 = 3 \quad (2)$$

at this frequency. Therefore, for a simplified analysis, the two-stage transistor amplifier with negative feedback can be replaced by an ideal current amplifier having a gain of 3 and with no phase shift between input and output; the oscillator circuit, thus modified, can be drawn as shown in Fig. 2(b). A comparison with Fig. 1 shows that each

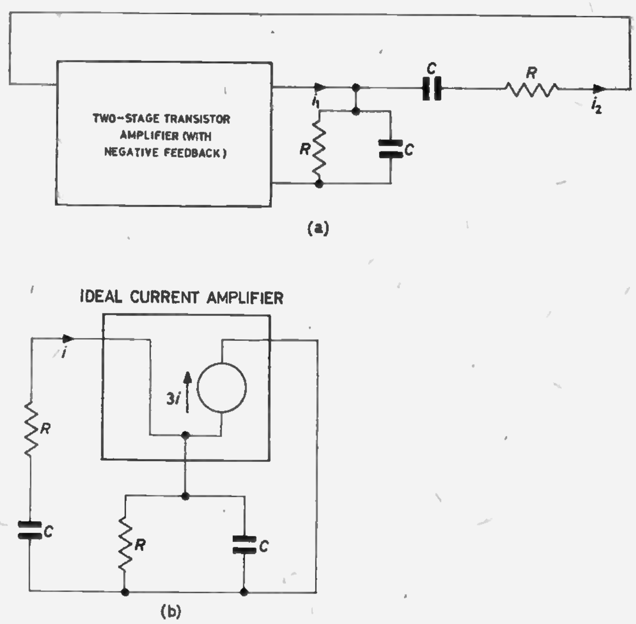


Fig. 2.
(a) Current derived RC oscillator
(b) Modified form of the current derived RC oscillator

RC pair in Fig. 2(b) is reflected across the other as a pair of negative impedances. Thus a generalized form of *RC* oscillator circuit, as shown in Fig. 3, can be evolved, with the negative sign being affixed to any impedance pair without loss of generality.

For sustained sinusoidal oscillations, the loop impedance of the network in Fig. 3 should have zeros on the imaginary axis of the complex frequency plane. The loop impedance:

$$Z(s) = \frac{1+s\,(R_1C_1+R_2C_2-R_2C_1)+s^2R_1R_2C_1C_2}{sC_1\,(1+sR_2C_2)} \quad (3)$$

From equation (3), the condition of oscillation is:

$$R_1/R_2 + C_2/C_1 = 1 \quad (4)$$

and frequency of oscillation:

$$f = \frac{1}{2\pi\sqrt{(R_1R_2C_1C_2)}} \quad (5)$$

The current derived *RC* oscillator which is the current-derived version of the Wien bridge oscillator, can now be seen to be a special case of this generalized circuit. From consideration of Figs. 1 and 2, it can be seen that the parallel RC combination is reflected across the series RC branch as $-2R$ in parallel with $-C/2$, thus satisfying the condition of oscillation given by equation (4). Similarly by considering the two-stage valve amplifier in the Wien bridge oscillator circuit as an ideal voltage amplifier having a gain of 3, the Wien bridge oscillator also can be reduced to the generalized form shown in Fig. 3.

Fig. 3 (right). Generalized form of RC oscillator circuit

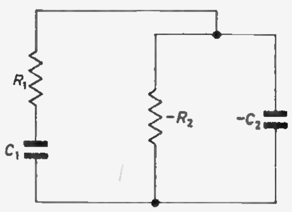
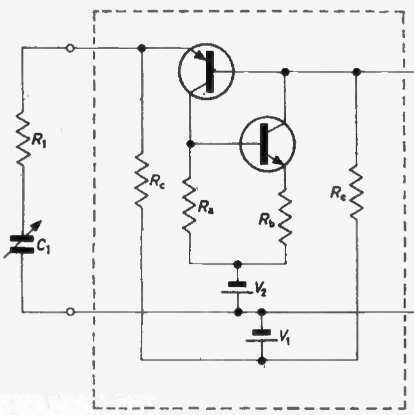


Fig. 4 (below). Transistor RC oscillator using negative impedance convertor
(The negative impedance convertor circuit is shown inside dotted lines)



RC Oscillator Using Negative Impedance Convertor

The generalized oscillator network of Fig. 3 can be directly synthesized with a well-known negative impedance convertor[2]. The resulting oscillator circuit, shown in Fig. 4, can be operated in two primary modes depending upon the ratio of (R_a/R_b) (Refer to Fig. 4).

MODE 1

$R_a = R_b$. This is the usual negative impedance convertor[2] circuit and assuming ideal convertor action, R_2 and C_2 will be reflected as $-R_2$ and $-C_2$ at the input of the convertor. In this mode of operation, the condition of oscillation given by equation (4) can be satisfied by keeping:

$$R_1/R_2 = C_2/C_1 = 1/2 \quad \ldots\ldots\ldots\ldots (6)$$

The frequency of oscillation becomes:

$$f = \frac{1}{2\pi R_1 C_2} = \frac{1}{2\pi R_2 C_2} \quad \ldots\ldots\ldots\ldots (7)$$

The frequency of oscillation can be varied by keeping R_1 and R_2 constant and by using a three-gang tuning such that two sections are in the series arm and the third section in the parallel arm. It is to be noted here that the input port of the convertor is open-circuit stable and the output port short-circuit stable, the RC branches cannot be interchanged.

MODE 2

$R_a = 2R_b$. In this case, assuming ideal convertor action, it can be shown that R_2 and C_2 are reflected as $-2R_2$ and $-C_2/2$ at the input. The circuit is similar to the circuit in Fig. 2(b). The condition of oscillation is now satisfied by keeping:

$$R_1/R_2 = C_2/C_1 = 1 \quad \ldots\ldots\ldots\ldots (8)$$

Results and Conclusions

In the experimental oscillator circuit, the entire audio frequency range of 20c/s to 20kc/s could be covered by keeping the resistances constant and varying the capacitances. The output taken at the collector of the pnp transistor was stable and maintained purity of waveform through the frequency range. By using capacitances of large values, it was found possible to achieve as low a frequency as 0·1c/s. Improvement in the oscillator performance and extension of the frequency range can be expected by using Darlington's compound connexions and by using transistors with high α-cut-off frequencies.

This oscillator scheme has all the advantages of RC oscillators as well as some additional merits. Its chief advantage is that the common point of the tuning can be grounded, unlike in the bridge type of oscillator where it has to be kept above ground potential. Moreover, due to the direct coupling between transistors and absence of a separate negative feedback network, this circuit uses fewer components. The circuit also achieves very low frequencies of oscillation due to direct coupling. Hence, this circuit can very well form the basis for a cheap and compact laboratory oscillator.

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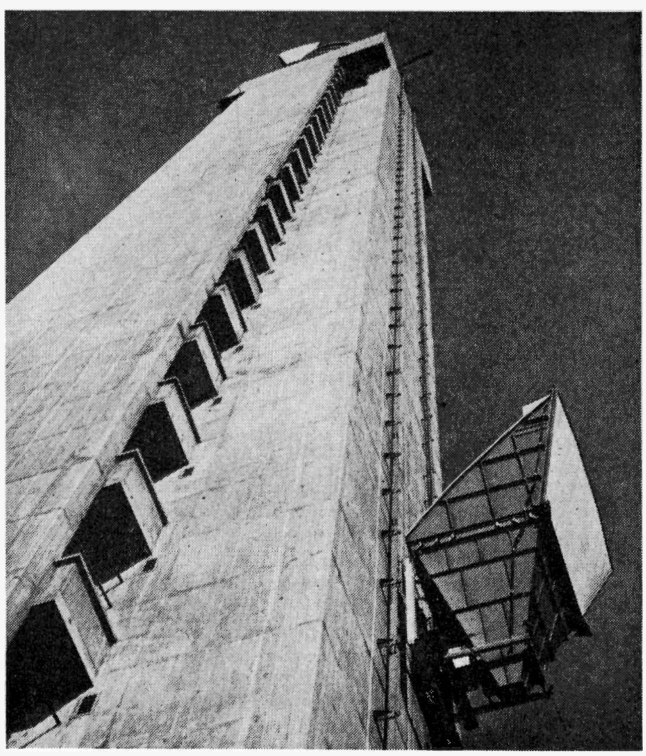
The New Post Office Tower in Birmingham

The photograph shows one of the aerials, running on steel guide rails, being pulled towards the top of the new 500ft Post Office Tower in Birmingham. The horn paraboloid aerials are 27ft long, 14ft wide and weigh about 1½ tons.



SHORT NEWS ITEMS

The Institute of Physics and The Physical Society is arranging a two-day conference on 'The transport properties of superconductors' on 30 to 31 March 1967 at The University of Kent at Canterbury.

The conference will cover both experimental and theoretical aspects of the transport of electric charge, heat and sound through superconductors. It is expected that the emphasis will be on type-II superconductors, although it is hoped that there will be some contributions on non-linear and quantum interference phenomena in type-I.

The organizing committee welcomes offers of contributions. Contributors should submit, not later than 17 February 1967, three copies of synopsis of about 200 words for consideration by the papers committee. Synopses should be sent to the Conference Secretary, Dr. G. Rickayzen, The University Physics Laboratories, Canterbury, Kent.

Advanced registration for attendance at the conference will be necessary and further details and application forms are available from the Meetings Officer, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

'A General Guide to the Safe Use of Lasers' has been prepared and published by the Electronic Engineering Association. As its title suggests, the Guide is general and informative and does not purport to be comprehensive or definitive.

In the absence of any statutory regulations governing the safe operation of lasers, this guide is primarily intended for use by commercial organizations and educational bodies engaged in their use.

Copies of the booklet are available from the Electronic Engineering Association, Berkeley Square House, Berkeley Square, London, W.1 (2s. per copy).

The National Research Development Corporation is giving support to work at the University of Sussex designed to find a more compact alternative to the cathode-ray tube.

The new device, known as a 'magnetic visual display panel', will be developed by Dr. A. W. Simpson of the Applied Sciences Laboratory, whose invention it is.

The magnetic display panel is made of a matrix of transparent magnetic elements each having two distinct magnetic states, one of which is opaque to polarized light, a beam of which is shone on to the matrix.

A picture or display is built up, the form of which is determined by the pattern in which the individual cells of the matrix are magnetized. The system is analogous to that of the advertising display which travels along a line of light bulbs.

The current paths provided through the matrix for switching the elements can be arranged in a pattern similar to that used in computer 'memories'.

The device depends on the Faraday effect—the effect of rotation of the plane of polarization of incident radiation in a direction depending on the direction of magnetization of a material having two distinct states of magnetic remanence.

The path for the light through the elements of the matrix is arranged so that there is a rotation of 45 degrees and switching of the magnetic state from one direction to the opposite rotates the plane of polarization by 90 degrees.

The current paths provided through the matrix for switching the elements can be arranged in a pattern similar to that already known for magnetic core stores in computer memories. That is to say the elements are arranged in a matrix of rows and columns, and a current conductor is provided for each individual column of elements. Then if the currents applied to the row and column conductors are arranged to be insufficient to reverse the magnetization of any element alone, but sufficient to switch the element in which current flows coincidently, a given element may be selected.

Possible materials from which the elements can be formed comprise yttrium iron garnet, gadolinium iron garnet or mixtures of these.

The Post Office has provided links for another new television transmitting station which is now in service by the BBC at Pontop Pike, County Durham. This is the latest in a series of extensions of vision circuits provided by the Post Office for BBC-2. Engineered to full 625-line standards, this expanding network is designed to extend BBC-2 programmes eventually to the whole of Britain. The circuits are inherently capable of the transmission of colour signals.

The completion of the new links follows the establishment over the past two to three years of a network of main vision circuits which, by the end of this year had already been extended as far north as Manchester (tying in with the BBC transmitters at Winter Hill, Lancashire, and Emley Moor, Yorkshire), as far west as Cardiff (serving the transmitter at Wenvoe) and as far south as Southampton (linking with the transmitter at Rowbridge, Isle of Wight).

The next transmitting stations to be established for BBC-2, for which Post Office vision circuits have been ordered, are to be located at Divis, Belfast (Winter, 1966); Durris, Kincardineshire (early 1967) and Tacolneston, Norfolk (early 1967). Provision by the Post Office of the various main, local and transmitter circuits needed to serve these new stations are already well advanced.

Post Office links for a number of such stations will have been completed by the end of 1967 or early 1968, bringing BBC-2 to some 70 per cent of the population.

Her Majesty The Queen has commanded that The Television Society shall now be known as "The Royal Television Society".

This honour has been bestowed at the beginning of the Society's 40th anniversary year and it highlights the contributions which members of the Society have made to the development of British television.

The Royal Air Force Engineer Branch is the new title of the Royal Air Force Technical Branch introduced as the result of a study of R.A.F. engineering duties which has led the Air Force Board to conclude that both at Command headquarters and at subordinate formations, technical and engineering staffs should come together to form a single organization, the duties of which can most effectively be divided between Mechanical Engineering and Electrical Engineering. This division of duties has been recognized in the training at the R.A.F. Technical College since 1955.

The technical and communications staffs at all formation levels will be combined to form a single organization under the control of the appropriate senior technical officer. Within each formation there will be a division between mechanical engineering and electrical engineering duties, on lines similar to the organization of technical staffs of the Ministry of Defence.

Mechanical Engineering will cover airframes, engines, weapons, air-launched missiles, mechanical ground servicing equipment, mechanical transport and marine craft. Electrical Engineering will cover communications, ground and airborne electrical and electronic equipment, instruments and surface-launched missiles.

The British Standards Institution has published the first part of a new British Standard for 'Retainers for electronic tubes and valves,' BS 4053 covering retainers for tubes and valves of the types included in BS 448, Dimensions of electronic tubes and valves.

BS 4053, Part 1: 'General requirements and methods of test,' defines terms, gives a classification into categories according to BS 2011 and gives tests and measuring methods. Part 2, which will be published shortly in loose-leaf form, will contain data sheets for individual types of retainers.

Copies of BS4053 may be obtained from the BSI Sales Branch, 101-113 Pentonville Road, N.1. Price 6/- each (postage 9d extra will be charged to non-subscribers).

The 7th International Conference on Microwave and Optical Generation and Amplification will be held in autumn 1968 in Hamburg (Federal Republic of Germany). The Conference will be organized by the Nachrichtentechnische Gesellschaft im Verband Deutscher Elektrotechniker, of Stressmann Allee 21 Frankfurt/Main S.10 Federal Republic of Germany.

A Conference on Semiconductor Device Research will be held in Bad Nauheim (Federal Republic of Germany) on April 19 to 22, 1967. This conference is sponsored by the Region 8 of the Institute of Electrical and Electronics Engineers (IEEE), the Deutsche Physikalische Gesellschaft (DPG), the Verband Deutscher Elektrotechniker (VDE) including the Nachrichtentechnische Gesellschaft im VDE (NTG).

Prospective authors for short papers are invited to submit eight copies of an abstract (10 lines) to:

Prof. Dr. W. J. Kleen, 8 Munchen 8 (F.R. Germany), Balanstr. 73.

The conference will follow the meeting of the Deutsche Physikalische Gesellschaft on Semiconductors, Metals and Magnetics, scheduled 17-19 April 1967 at the same place.

Prospective participants are requested to write to:

Dr.-Ing. H. H. Burghoff, German Section IEEE,
6 Frankfurt/Main 70 (F.R. Germany),
Stresemann Allee 21,
VDE-Haus

or to

Dr. phil. K.-H. Riewe, DPG,
645 Hanau (F.R. Germany),
Heraeusstr. 12-14.

The British Scientific Instrument Research Association (SIRA) is to hold a Conference on 'New developments in optics and their applications in industry' at the Grand Hotel, Eastbourne, on 11 to 12 April, 1967.

The object of the Conference is to present advances in and new applications of optics, and to bring out for discussion problems that have arisen (or are likely to arise) in the application of optics to such diverse activities as microelectronics, process control, and data processing, among others.

Among the subjects to be covered are:

Transmission of information by light
Holography and its possibilities
Data processing
Optics in the electronics industry (e.g. microelectronics)
Optical requirements in process control
Applications of optics in space.

Applications to attend the Conference will be accepted from non-members of SIRA, but priority will be given to SIRA members in allotting places. Requests to be sent registration forms when they become available may be addressed to the Publicity and Literature Services Department, SIRA, South Hill, Chislehurst, Kent.

The Post Office has provided three high-definition closed circuit television links to connect London Airport Control Tower with the new Southern Air Traffic Control Centre at West Drayton. As vital data about flight patterns, etc., is scanned by television cameras for transmission to the Control Centre, the service demands that transmissions should be of the highest possible definition.

These links operate on 875 lines and have a frequency response which is flat to 11Mc/s.

This is the first time the Post Office has been asked to provide this very wide band width for close circuit television.

British Overseas Airways Corporation has placed a further order for airborne navigation and communication equipment with The Marconi Co. Ltd, for for their fleet of VC-10 and Super VC-10 aircraft. This new order brings the total value of BOAC orders for Marconi avionic equipment for these two aircraft types, to nearly £1½M.

Equipment in the new order includes v.h.f. communication and navigation systems.

The entire BOAC jet fleet, including all the Boeing 707 aircraft, is now in the process of being fitted with the Marconi Doppler Navigation system, type AD560. The VC-10 and Super VC-10 aircraft are also equipped with a comprehensive range of Marconi avionic equipment which, in addition to communication and navigation systems, includes the low frequency automatic teleprinter receiver, type AD308 and a selective calling device, SELCAL type 2880.

I.C.T. has awarded Mullard Ltd a contract worth £400 000 to supply matrix core stacks for the data stores used in its 1900 series of computers.

Designed for the main working stores of the 1901, 1902 and 1909 machines, the stacks have capacities of 4 096, 8 192 or 16 384 25-bit words. More than 73 million ferrite cores will be used in fulfilling the contract. Each core has an outside diameter of 0·050in and is threaded by four wires used to read information into and out of the computer memory.

Plessey Radar—part of the Plessey Electronics Group—has been awarded a design study contract for the first international satellite control centre.

The installation will be at Noordwijk in Holland, at the European Space Technology Centre of the European Space Research Organization (ESRO), and will co-ordinate ground stations round the world used for controlling and commanding scientific research satellites launched by ESRO.

The contract is the first of its kind to be awarded to a British firm and will involve an intensive study to determine which of the alternative systems will best meet ESRO requirements.

Denmark's first u.h.f. television transmitter, to be used for colour and u.h.f. propagation tests, is to be supplied by The Marconi Co. Ltd. The transmitter will be installed at Gladsaxe, a suburb of Copenhagen, where black and white, technical test transmissions are due to begin early next year, a preliminary to the start of colour television broadcasts.

The order includes the supply of an aerial as well as the transmitter which is a completely new model. The power output is 10kW, generated by a new, high stability, vapour-cooled, 4 cavity klystron tube supplied by the English Electric Valve Company. The aerial will be specially built to suit an existing mast at the Copenhagen site.

NEW BOOKS

Synthesis of Filters

By J. L. Herrero and G. Willoner. 192 pp. Med. 8vo. Prentice-Hall International. 1966. Price 84s.

IT IS difficult to decide what type of reader this book is aimed at. The later chapters contain material which would be useful to (and perhaps in parts fully intelligible only to) a filter designer. Yet, the first few chapters discuss the fundamental notions with almost a naive approach to avoid, it would appear, a rigorous statement of the basic properties of passive networks.

The authors consider the design of ladder filters, on the basis of their insertion-loss characteristics. Synthesis procedures are discussed, basically for filters containing only reactive components which are loss-free. Near the end of the book *RLC* filters are briefly discussed and methods of compensating for the lossy components are indicated.

Filters are considered in four groups (the grouping being based on the type of the response function) as Butterworth, Chebyshev, Cauer Parameter and General Parameter filters. Although one might feel that the distinctions between the last three groups of filters are rather artificial, the properties of the various types are clearly stated and discussed.

An iterative method for the calculation of element values without solving for the roots of the polynomials in filter functions is presented in the Chapter 8, for which method computational advantages are claimed as compared with the methods in general use.

Another interesting feature of the book is the exposition, in Chapter 10, of a template method for determining the attenuation poles for given insertion-loss requirements.

A number of numerical examples are worked out and references given for further reading.

The overall impression the reviewer had of this book was that the presentation did not do justice to the material presented.

H. O. BERKTAY

Analysis and Design of Transistor Circuits

By L. G. Cowles. 309 pp. Med. 8vo. D. Van Nostrand. 1966. Price 78s.

ON THE loose cover we read: 'A refreshingly new and direct treatment of transistor circuit principles is presented in this book, which is primarily concerned with practical circuit design.' In the preface the author states that 'this is the first book to recognize the impact of the planar transistor on circuit design.'

These are perhaps rather sweeping statements and might cause the potential reader to believe that here was a completely new approach. All books have some emphasis and in many there is something novel. In the present one the assumption is made from the beginning that the reader is familiar with semiconductor device physics, α, β, the hybrid parameters and so on: it can therefore be regarded as complementary to the texts dealing with semiconductor device physics and equivalent circuits. The 'novel' feature of the book is the neglect of the internal collector-base resistance for both d.c. and a.c. applications. Very naturally this reduces the complexity of the analysis of linear circuits. It gives *de facto* recognition to what the practising circuit engineer has long done, even before the advent of the planar transistor.

The main part of the book is concerned with d.c. and low frequency circuit applications, in which the reactive elements of equivalent circuits can be ignored. There is, however, a short chapter on 'medium frequency power amplifiers', but this is rather limited in scope. The book deals adequately with biasing, low-frequency and d.c. amplifiers (including a good section on coupled pairs), power amplification and low-frequency power switching. The last three of the 21 chapters deal with tetrode and field effect transistors and are more limited than the more detailed chapters on bipolar devices.

The readership target is 'applications engineers, circuit designers or college students in upper levels of graduate courses'. It is not an undergraduate course book but will be of use to the intended readers.

F. J. HYDE

Modulation, Resolution and Signal Processing in Radar, Sonar and Related Systems

By R. Benjamin. 184 pp. Demy 8vo. Pergamon Press. 1966. Price 55s.

A VERY laudable feature of this book is the deliberate avoidance of tedious mathematical formulations (which are often used by authors to disguise their lack of real understanding of the subject) even if, on occasions, rigour has to be sacrificed in the process. The accent has been placed instead upon striving to produce a clear appreciation of the physical principles and relationships involved. This does not mean, however, that the book is easy to read and comprehend. A newcomer to the field of signal processing would find the wealth of new definitions, concepts and principles rather overwhelming and some rearrangement and simplification of the material would be necessary to render it suitable for use at undergraduate level. It should, however, be of considerable value to theoreticians and engineers, with some previous experience of the subject; this is indeed the market at which it is aimed.

The first part of the book is devoted to definitions of the discriminating and resolving powers of detection systems in range, azimuth, elevation and Döppler together with their similarities and relationships. There follows a detailed examination of detection systems which are matched to given signal and noise distributions in the frequency and time domains and these are very thoroughly developed in terms of a wide variety of modulation systems. The concept and significance of the ambiguity function is examined also and at the end of the book individual chapters are devoted to logical and non-linear processing and to practical considerations. The various systems discussed are clearly illustrated by a large number of block schematics but notwithstanding the title these are almost invariably chosen from the field of radar.

In two ways this book is unconventional; firstly it contains an abstract and a postscript, both of which the reader is advised to look at so that he can more readily attune himself to the authors individualistic approach to the subject and secondly, it contains only one reference. Although it is pointed out in the postscript that usually specific points and ideas cannot readily be credited to particular individuals the reviewer feels that some attempt should have been made to include at least a few key references especially in view of the deliberate sacrifice of rigour in certain places.

B. K. GAZEY

Introduction to Transistor Electronics

By R. L. Walker. 341 pp. Med. 8vo. Black and Son Ltd. 1966. Price 50s. (paperback 27s. 6d.)

THIS book follows the conventional pattern for textbooks on transistors, progressing from semiconductor physics through circuit models to a variety of circuit designs. The exception to this pattern is that one chapter on 'Further Developments in Semiconductor Devices', including the field-effect transistor and the tunnel diode, is placed in the slightly odd position between 'Band-pass Amplifiers' and 'Switching Circuits'. There are only a few lines on integrated circuits, phototransistors and voltage stabilizers and no suggestions for practical work.

The author has written 'Introduction to Transistor Electronics' in a style which is easy to read, with a number of light-hearted touches. At one stage the action of electrons and holes is explained by analogy to hawks and rabbits and in a later chapter Kelley's Second Law is quoted as "If anything can possibly go wrong it will!" Throughout most of the book the term hertz is used instead of cycles/sec. The author followed the commendable practice of explaining circuit action on the basis of both d.c. transistor characteristics and circuit models, wherever this is possible. The relative advantages and disadvantages of various methods for circuit design are explained in a practical manner.

A good feature of this book is the interesting and searching problems which follow each chapter. Any student who faithfully completed these problems would acquire a useful background in transistor electronics. As with most textbooks of American origin, no solutions are given to the problems. This is a frustrating omission.

The hard-cover version of this book is attractively printed on good quality paper and deserves a place in any technical library. The paperback version could be a worthwhile investment for an undergraduate who is interested in electronics.

J. R. ABRAHAMS

Fundamentals of Display Systems

By H. H. Poole. 403 pp. Med. 8vo. Macmillan Co. Ltd. 1966. Price £7 15s.

IN the preface the author states that "the book was undertaken as a general introduction to displays and display systems" and that "each technique is presented in sufficient detail to acquaint the reader with the display device and its applications". Mr. Poole's object has not in fact been to produce a work devoted primarily to basic scientific principles, as perhaps the title of the book might at first suggest, but rather to provide a comprehensive survey of displays and display systems with just sufficient attention given to basic principles to enable the most significant points to be noted. The scope of the book should give it a broad appeal among those concerned in any way with the display field and not limit it to those engineers engaged in specialist design.

The text is divided into five major parts and within this framework a large range of display techniques is embraced. These include electronic, photographic, electromechanical, electro-optical, photochromic and oil-film methods. Part I contains a complete section devoted to colour cathode-ray tubes and terminates with a section on the constraints imposed by various forms of specification, including those related to environment and human engineering. Part II contains a section on display peripheral devices, such as character and vector generators, as well as dealing with the presentation of information on large screens. Part III is concerned with system considerations and contains complete sections dealing with television, radar, and computer-generated displays. Part IV is a particularly useful addition to the book in that it deals with human engineering considerations, optics, and luminescence with special reference to displays. The book is concluded by Part V which looks at likely future display techniques.

Referring again to the preface, the author hopes that his book "will serve as a bridge to a more detailed study in any given area". Mr. Poole's book should achieve this purpose very successfully. It is well organized, well presented, and is unquestionably a very useful addition to the information published in this field.

L. W. WHITAKER

Electronic Designer's Handbook— A Practical Guide to Transistor Circuit Design

By T. K. Hemingway. 294 pp. Business Publications Ltd. 1966. Price 63s.

THERE are three clearly defined sections to this book. Part 1, which is half the book, deals with basic circuits. Chapter 1 is entitled 'Semiconductor Diode Properties' and is largely devoted to the Zener diode. The second chapter is concerned with d.c. characteristics of the transistor, but there is not one illustration of a typical set of characteristics. In fact, the only voltage/current graph in the whole book is a generalized curve for a germanium diode. The chapters on transistor circuits emphasize the effect of temperature variations, which is particularly significant for germanium transistors.

Part 2 gives details of five special classes of transistor circuit, namely complementary circuits, voltage-controlled oscillator, ultra-high gain stages, transistor pump and transistor cascode.

Part 3 consists of two chapters devoted to the techniques of boot-strapping and prototype testing. Some more mathematical work is dealt with in six appendices.

Inevitably, in a book of less than 300 pages, the subject matter is less comprehensive than the title would suggest. However, there are some surprising omissions. Less than half a page is devoted to logic circuits and these few lines deal only with the diode AND gate. The words 'hybrid' and 'tuned' do not appear in the index. Unfortunately the T-equivalent circuit is used throughout and there is no mention of band-pass amplifiers. In a book of this type there should always be a list which gives the meaning of the many symbols and abbreviations used in the text.

Apparently this book was written for the recently-qualified electronic engineer or physicist, but it cannot be recommended to the new graduate. The older engineer, who has been brought up on thermionic valves and now wishes to design low-voltage stabilized power supplies, might find the book useful. It could have been written seven years ago and would have been more valuable at that time, before the widespread acceptance of integrated circuits changed the emphasis away from circuit detail to system design.

J. R. ABRAHAMS

Antenna Analysis

By E. A. Wolff. 514 pp. Med. 8vo. John Wiley and Sons Ltd. 1966. Price 188s.

DR. WOLFF based this book on a postgraduate course for aerial design engineers which he gave in 1962-3. Although mathematical in treatment the book has an engineering outlook, being so written that results may be extracted easily. Wherever possible, solutions are derived from fundamental electromagnetic theory, tedious working being omitted and approximate or graphical values being taken when appropriate. Some experimental results are also included.

The subject matter is wide, with a definite emphasis on those fields in which active work is being carried out at present. Thus, after an introductory chapter and a brief note on point sources, Chapter 3 deals with wire aerials, dipoles, folded dipoles, monopoles, cylindrical, spherical and spheroidal aerials, loops, ferrite loops and bicones. There follows a chapter on apertures, including circular and cylindrical as well as rectangular areas.

The long chapter on slot antennas also includes slotted cylinders and waveguides, notches, open waveguides and a substantial section on waveguide horns. Chapter 6 covers planar, circular and unequally spaced arrays as well as linear arrays. The next chapter deals with reflector aerials, plane, corner, paraboloidal, Cassegrainian, cylindrical, doubly curved, spherical, toroidal, passive reflectors and reflector arrays. Chapter 8, on travelling wave aerials, is equally comprehensive, and the book concludes with chapters on broadband and lens aerials.

The author draws his material from many sources, around three hundred references being quoted. This alone makes the book a valuable source of information. The book is well presented, with about two hundred and forty diagrams and graphs. There are a few minor errors in this first issue.

This is a book which deserves a place on the shelf next to the traditional aerial handbooks. Compared with them, it covers a wider field, provides more formulae for computers to evaluate, and is of course fifteen years younger.

M. F. RADFORD

Electronics

By Roy H. Mattson. 620 pp. Med. 8vo. J. Wiley & Sons. 1966. Price 98s.

THIS is an undergraduate textbook following the now customary American pattern of intensive study. It is designed as a one-year course taken at the rate of three or four lectures per week and would need considerable laboratory support to achieve full effect. In this way, the author is able to assume a wide knowledge especially of circuit theory and Laplace transforms from the beginning of the book. Without commenting on the merits of this system of teaching it is clearly different from the British one, and this difference makes it hard to use American textbooks to full advantage in British courses. The longer and more diverse the subject matter the greater is the difficulty.

The aim of the book is to develop an understanding of electronic devices in practical circuits, paying attention to the internal working of devices, the functions they are required to perform, and the circuits in which they therefore function. No attempt is made to separate the work of the device manufacturer from that of the circuit designer, and the author is clearly looking forward into the era of integrated circuit when the two tasks will be inseparable.

The material falls into three fairly clearly defined sections, of which the first and last are the important ones. Where appropriate the centre section could be omitted without loss of continuity. The first section of four chapters deals with the theory of solids, the operating principles of diodes and multi-terminal devices, and the effective circuits of real devices. On the whole, the treatment here is clear and straightforward with the rather novel approach of using photodiodes as an introduction to transistors very effective. Only the first chapter is open to serious criticism. The solid state theory needed for a full understanding of semiconductors as materials is complex and subtle, but as several recent books have shown such a treatment is not necessary as a preamble to a book on devices and circuits. Here the first chapter falls heavily between two stools, making an unfortunate beginning to a book which gets better as it goes on.

The second section forms a link between devices and circuits, and deals with transducers, transformers, amplitude frequency and pulse modulation systems, and the functions of diodes, including rectification and frequency conversion. This material would often be included elsewhere in an undergraduate course. Where this is the case it could be omitted from the scheme, but in any practically oriented course there is clearly a case for using it in a linking role, as here.

The longest part of the book needs least said about it, for it is the best and most straightforward. It is concerned with circuit design and construction, starting with the circuit models of active devices developed earlier. First it deals with the biasing of devices, and then after a resumé of linear network theory goes on to single stage amplifiers, cascaded amplifiers and switching circuits. There are omissions from the treatment, notably oscillators, time-bases and all high frequency and microwave applications, but the book clearly does not aim to be comprehensive. At least there is a hope that those reading it could design simple circuits for themselves.

The emphasis on the functional nature of the subject is admirable, and is helped by the choice of subject matter, as well as by the many worked examples in the text. Useful examples are also given at the ends of the chapters, some with answers. The book can therefore be recommended to anyone interested in its subject matter, with one further proviso. The author's style is very repetitive and is of the kind that will drive you mad before the end of the book if you happen not to like it. Read a few paragraphs before you buy.

D. C. NORTHROP

Higher Mathematics Examples for Electrical Engineering Students

By D. G. S. Bedding and D. W. Porter. 300 pp. Demy 8vo. Macmillan & Co. Ltd. 1966. Price 25s.

A large number of mathematical examples are included in this book and are of use for students taking the Higher National Certificate and Part II Mathematics and Part III examinations of the Institution of Electrical Engineers.

Answers to all the examples are given, and the index is sufficiently detailed to enable the reader to find examples associated with any particular subject.

Questions and Answers—

George Newnes Ltd. 1966. Foolscap 8vo. Price 8s. 6d.

On Audio

By Clement Brown. 104 pp.

On Electronics

By Clement Brown. 112 pp.

On Transistors

By Clement Brown. 96 pp.

On Radio & Television

By H. W. Hellyer. 128 pp.

These four books, as their titles suggest, consist of a series of questions and answers designed to give the interested layman an insight into these subjects.

Transistor Bias Tables

By E. Wolfendale. 71 pp. Crown 4to. Iliffe Books Ltd. 1966. Price 21s.

This book contains a series of computed tables to assist in the design and construction of a transistor amplifier. These tables can be used directly to provide the values of the three resistors required for the conventional bias current, or alternatively, as a starting point for a more detailed bias current analysis.

A short introduction is included giving the aims of the tables and how they should be used.

Advances in Cryogenic Engineering

Edited by K. D. Timmerhaus. 712 pp. Royal 8vo. Plenum Press. 1966. Price $19.50

Volume 11 of this series contains the 76 papers presented at the Eleventh National Cryogenic Engineering Conference held at the Rice University, Texas, in August 1965. A number of these papers deal with cryogenics as applied to space systems.

Nuclear Electronics Conference Proceedings, Bombay, 22-26 November 1965

International Atomic Energy Agency. 662 pp. Med. 8vo. Her Majesty's Stationery Office. 1966. Price 81s.

This book contains the 48 papers presented at the Conference on Nuclear Electronics held at Bombay in November 1965 and organized by the International Atomic Energy Agency of Vienna in co-operation with the Indian Atomic Energy Commission.

The Management of Innovatiou

By Tom Burns and G. M. Stalker. 269 pp. Med. 8vo. Tavistock Publications Ltd. 1966. Price 21s.

First published in 1961, this book now appears as a paperback in the Social Science Paperback Series, and deals with the subject matter in three parts, namely External Circumstances, Organization and Change, Direction and Shaping of Management Conduct.

Radio Valve Data

229 pp. Demy 4to. Iliffe Books Ltd. 1966. Price 9s. 6d.

The characteristics of 7,000 valves, transistors, semiconductor diodes and rectifiers, and cathode-ray tubes are now given in the eighth edition of this book.

Television Receiver Theory: Part 1

By G. H. Hutson. 238 pp. Crown 4to. Edward Arnold (Publishers) Ltd. 1966. Price 35s.

This textbook is the first of two volumes which together will provide a systematic course in the principles of television receivers for students and technicians preparing for the Intermediate and Final Examinations in Radio and Television conducted by the City and Guilds of London Institute and the Radio Trades Examination Board (Course 48)

Designing Transistor I.F. Amplifiers

By W. Th. Hetterscheid. 314 pp. Med. 8vo. Cleaver-Hume Press Ltd. 1966. Price 80s.

The subject matter of this book is the design and construction of transistorized i.f. amplifiers for radio, television and radar receivers. A survey of the design theory is given from which practical procedure may be developed using special design charts.

Six fully worked out examples are included.

Handbook of Relay Switching Technique

By J. Appels and B. Geels. 321 pp. Med. 8vo. Cleaver-Hume Press Ltd. 1966. Price 72s.

This book deals with the theory of switching techniques and includes chapters on the design of circuits for counting, decoding, checking, locking and similar topics.

Measuring and Testing with Square Wave Signals

By W. Schultz. 196 pp. Med. 8vo. Cleaver-Hume Press Ltd. 1966. Price 57s. 6d.

After introductory chapters dealing with the square wave signal, this book describes the various methods of testing in the audio frequency range including stereo amplifiers and record players.

The testing of components and sub-assemblies is included.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Simple Gates Without Pedestal

DEAR SIR,—I would like to make a few comments concerning the letter 'An R.F. Gate with no Pedestal Voltage' by D. Anderson in your August issue.

(1) The output appearing when a transmission gate is opened and closed for the condition 'no input signal' is shown, for the general case, in Fig. 1. The 'pedestal', Δ, arises because of the

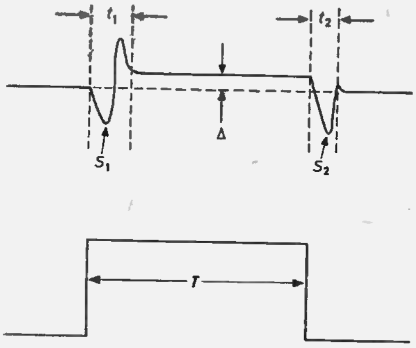


Fig. 1. General case

different d.c. conditions existing at the output for the two states 'gate open' and 'gate closed' when the gate circuit is actuated by a control signal (usually a pulse) of duration T. The amplitude, polarity, and duration (t_1, t_2) of the switching 'spikes' S_1, S_2 will depend on the circuit configuration employed. They arise because of non-coincidence and/or finite, and in general unequal, transition times of the two opposite polarity gating signals in balanced circuits, and/or because of reactive circuit components. (Carrier storage effects might also contribute to their amplitude, duration, etc.)

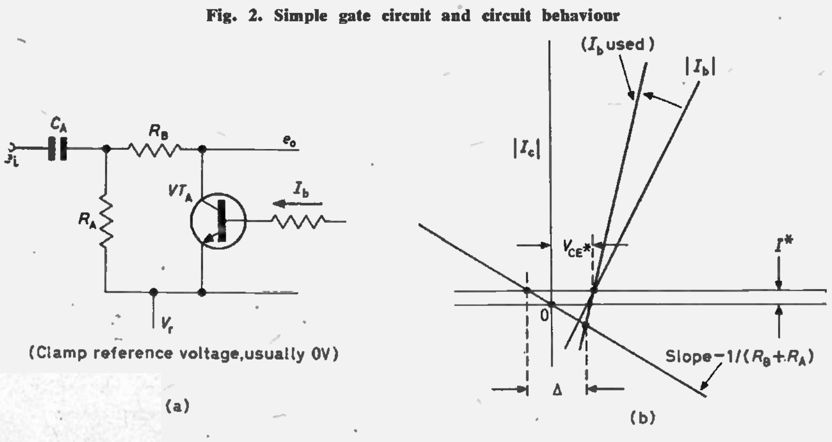
The presence of S_1 and S_2 does not depend on the presence of Δ though the magnitude of Δ may vary their amplitude and duration. An incoming signal (a.f., r.f., or complex waveform) is superposed on the waveform of Fig. 1. 'Pedestal' is undesirable in most transmission gates but so are spikes. Which of the two effects is most troublesome from the point of view of its minimization will depend on the gate duration and not necessarily the frequency of the input signal. Pedestal is more easily removed, or cancelled out, than spikes. This applies whether T be in nanoseconds or microseconds, though the manner in which this is accomplished might require somewhat sophisticated circuit arrangements (such arrangements will form the basis of a future communication).

The simplest possible gates avoid or reduce pedestal at the expense of the spikes S_1, S_2, or switch-on and switch-off times, but are attractive because of their economy and the fact that they require no adjustment. They may still be able to handle r.f. sinusoidal signals but not short gate durations. This is because they may employ a transistor operating in the saturating mode. Consider the arrangement of Fig. 2(a): this circuit (British Patent 861,263) was developed by the author in 1958 while working on problems relating to the production of inter-trace marker rings for radar c.r.t. displays.

The original circuit used OC44's because they appeared to be the best then commercially available for the purpose, and the input signal was 10kc/s.

Fig. 2(b) summarizes the circuit behaviour assuming the applicability of the ideal Ebers-Moll transistor model.

Fig. 2. Simple gate circuit and circuit behaviour



It may be shown that:

$I^* = I_{oo}(1-\alpha_R)/(1-\alpha_R\alpha_F)$

$V_{CE}^* = (KT/q)\ln(1/\alpha_R)$

r_d = Collector emitter incremental resistance with $I_c \simeq 0$

$$= (KT/q)\,.\,(1/|I_b|)\left[(1/\beta_F)+\frac{1}{(\beta_R+1)}\right]$$

where the symbols have their conventional significance. If the transistor is

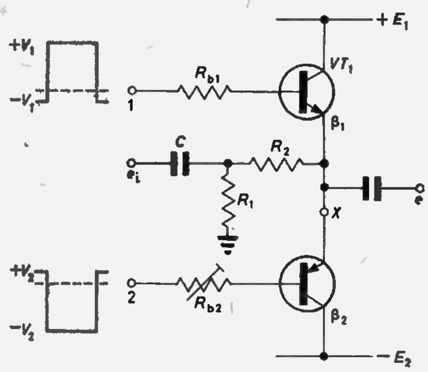


Fig. 3. Anderson's circuit

operated in the inverse sense then I_{oo} becomes I_{eo}; α_F and α_R are interchanged as are β_F and β_R. The pedestal voltage may be evaluated by simple geometry: it is very small (e.g. 1mV) and not significantly temperature dependent. The preferred base driving circuits for reducing S_1, S_2 and ensuring optimum speed performance of a transistor clamp as employed here are discussed elsewhere[1,2].

The advent of planar technology has made available such devices as the dual-emitter integrated chopper transistor and this has meant a smaller Δ (e.g. the Transitron ST5610 has a maximum offset voltage of 50μV at $I_{E_1E_2} = 0$) and improved switching performance.

Presumably Mr. Anderson has ruled out saturated switches because of carrier storage effects. In doing so he has sacrificed stability of pedestal cancellation as I think the following paragraphs will show. In view of the availability of high speed saturating transistor switches it is difficult to say how much he gains.

(2) Mr. Anderson's circuit is redrawn for convenience in Fig. 3.

It will be seen that the arrangement differs from my original circuit in two respects, the manner in which the output is clamped, and the necessity for two gate pulses.

According to the values he quotes it would appear that unsaturated emitter-followers clamp the output level when the gate is shut. This is presumably to enable the circuit to switch 'on' and 'off' faster.

When VT_1 and VT_2 are cut off, X will be very near earth potential. When VT_1, VT_2 are switched on then for zero shift in d.c. output level the emitter currents of VT_1 and VT_2 must be equal. Fig. 4 shows how this is achieved in the circuit of Fig. 3 by adjustment of R_{b_2}.

Neglecting leakage currents a simple calculation shows that for the emitter currents of VT_1 and VT_2 to be equal, under d.c. conditions, it is necessary that,

$$\frac{(\beta_1+1)}{(\beta_2+1)} = \frac{(V_2-V_{BE_2})}{(V_1-V_{BE_1})}\cdot\frac{R_{b_1}}{R_{b_2}} \quad ...... (1)$$

Adjustment of R_{b_2} might well ensure this at a given temperature but β_1, β_2, V_{BE_1}, V_{BE_2} are temperature dependent and while the variations will be the same direction for both devices exact cancellation cannot be expected of unselected discrete devices. Any difference between the emitter currents of VT_1 and VT_2 must flow in R_2 so a pedestal will appear.

Of course, the setting of R_{b_2} will affect the transient behaviour of the circuit. Referring again to Fig. 3, the turn-on delay time of VT_1 is given by t_{d_1} where,

$$t_{d_1} \simeq C_1 R_{b_1} \ln \frac{V_1+V_1}{V_1+V_{TH_1}} \quad .... (2a)$$

in which,

C_1 = sum of the 'effective' collector and emitter junction transition capacitances of VT_1

and V_{TH_1} = base-emitter voltage of VT_1 at the threshold of conduction.

Likewise the turn-on delay for VT_2, t_{d_2} is given by

$$t_{d_2} \simeq C_2 R_{b_2} \ln \frac{V_2+V_2}{V_2+V_{TH_2}} \quad .... (2b)$$

in which like symbols apply.

If $V_1 = V_2$, $v_1 = v_2$, $v_{TH_1} = v_{TH_2}$ then $t_{d_1} = t_{d_2}$ only if $C_1R_{b_1} = C_2R_{b_2}$. If $C_1 \simeq C_2$ then $t_{d_1} \neq t_{d_2}$.

If the turn on delays are not equal one transistor will take control and the output voltage will change until the other transistor starts to conduct. If $C_1 = C_2$, t_{d_1} can be made equal to t_{d_2} by adjusting v_1 or v_2.

Assuming $t_{d_1} = t_{d_2}$, a switch-on transient will exist if the emitter currents of VT_1, VT_2 do not increase to their final value at an identical rate.

Let γ_{b_1} be the base charge control time-constant of VT_1, i.e. the base charge per unit base current (in the d.c. state) and suppose a step base current I_{b_1} is applied when the gating pulse is applied at terminal 1.

Then the emitter current of VT_1, I_{E_1}, is given by

$$I_{E_1}(t) = I_{b_1} + \beta_1 I_{b_1}\,[1-\exp(-t/\tau_{b_1})] \quad .... (3a)$$

similarly for VT_2,

$$I_{E_2}(t) = I_{b_2} + \beta_2 I_{b_2}\,[1-\exp(-t/\tau_{b_2})] \quad .... (3b)$$

Equations (2) and (3) may be derived from basic charge control theory[3]. For minimum switch on spikes.

$$I_E(t) = I_{E_2}(t) \quad .... (4)$$

or substituting from equations (3a), (3b),

$$I_{b_1} + \beta_1 I_{b_1}\,[1-\exp(-t/\tau_{b_1})] = I_{b_2} + \beta_2 + I_{b_2}\,[1-\exp(-t/\tau_{b_2})] \quad .... (5)$$

For $t \to \infty$, equation (1) results.

Since normally β_1, β_2 are both $\gg 1$ then it follows from equation (5) that if the circuit is adjusted for zero pedestal minimum switch-on spike occurs provided $\tau_{b_1} = \tau_{b_2}$. This indicates the necessity of having VT_1 and VT_2 as complementary types. Small differences between τ_{b_1} and τ_{b_2} might be reduced by adjustment of E_1 or E_2 (this will in general alter the β of the relevant transistor also). Gross differences would require some other technique e.g. a series R and C in parallel with R_{b_2} would alter the switching performance of VT_2 without modifying the d.c. balance condition.

The switching times of VT_1 and VT_2 may be reduced and thus spike cancellation improved if VT_1 and VT_2 are 'charge driven.' This means R_{b_1}, R_{b_2} are shunted by capacitances C_{b_1}, C_{b_2} the choice of which is discussed by Sparkes[3].

(3) It is well known that at low frequencies the impedance looking in at the emitter of an emitter-follower is

$$R_{IE} = R_s + \beta r_e/(\beta+1)$$

where R_s = Base source resistance (including $r_{bb'}$)

$$r_e = KT/qI_e$$

Thus R_{IE} is only small if R_S is small.

Using Mr. Anderson's values $R_s \simeq R_{b_1} = 22\text{k}\Omega$, and assuming $\beta = 20$ for the P 346A (a data sheet minimum).

Thus $R_t \simeq 0{\cdot}5\text{k}\Omega$.

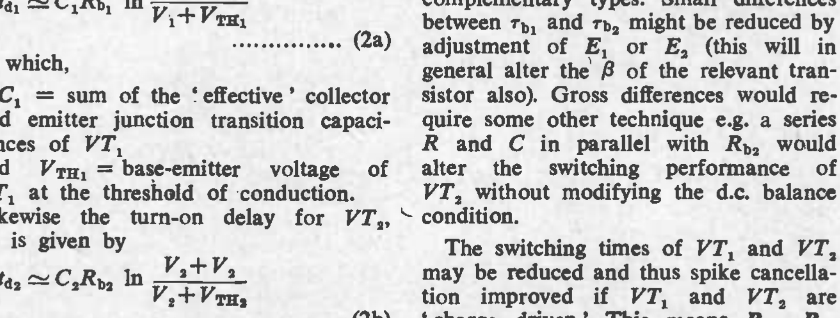
Taking the optimistic condition $R_b = \infty$ (where R_b is defined by Mr. Anderson as the impedance seen from R_2 looking into the back biased base emitter circuits of VT_1, VT_2 in parallel), then the low frequency 'on' to 'off' attenuation ratio is 30 compared with the figure of 10^3 or 10^4 quoted. A simple expression for the 'on' to 'off' attenuation ratio, involving resistors alone, will not suffice at higher frequencies (and the gate is supposed to deal with 'r.f.' signals) because reactive circuit elements have to be considered.

My own feeling is that the most economic simple no-pedestal gate will, when the devices become somewhat less expensive, be a series-shunt arrangement employing f.e.t.'s such as described, for example, by Hunter[4]. These arrangements would appear to be simple, have no pedestal, are fast, have an inherent tendency for spike cancellation and require no setting up.

Yours faithfully,

B. L. HART,
Department of Electrical Engineering.
West Ham College of Technology,
London, E.15

Fig. 4. Effect of adjustment of R_{b_2}
(a) $|V_1| = |V_2|$ (b) $|V_1| \neq |V_2|$



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3. SPARKES, J. J. Junction Transistors. (Pergamon 1966).
4. HUNTER, G. N. A M.O.S.T. Gate for Pre-Sense Amplifier Strobing. *Electronic Engng. 38*, 432 (1966).

Stabilizing Output of Wien Bridge

DEAR SIR,—One of the most convenient methods of stabilizing the output amplitude of a Wien bridge oscillator is to make one of the resistive arms level conscious by using a thermistor which controls the amount of negative feedback around the amplifier.

There are two reasons for the temperature of the thermistor in the circuit to rise—one is due to the dissipation by the oscillator output and the other because of the ambient temperature. During normal operation of the circuit, the operating temperature of the thermistor due to the dissipation is made large so that variations in ambient temperature have little effect.

In general to obtain high operating temperatures with little dissipation thermistors with high power sensitivities (e.g. STC type R series) are used.

Recently the variations in the output of the oscillator with temperature were studied in connexion with a project and during the investigations it was realized that use of a second thermistor in the other resistive arm can be made to give better circuit performance with temperature. The purpose of this brief note is to report our finding with the hope that it might be of some interest to your readers.

Fig. 1 shows the usual schematic of a bridge oscillator. To start assume that R_2 is a temperature independent

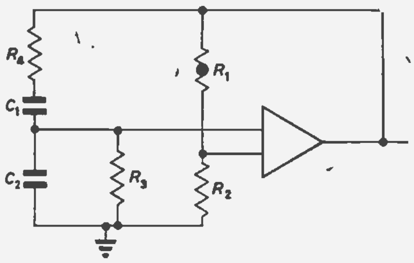


Fig. 1. Arrangement of bridge oscillator

resistor. As the ambient temperature rises the value of the thermistor R_1 decreases, increasing the amount of negative feedback and as a result the amplitude of the output drops.

Now, if the value of R_2 could be decreased so that the amount of negative feedback is restored towards the previous value the output tends to rise again to the original value. This principle can be mechanized by using another thermistor with R_2 to make it temperature sensitive. The second thermistor is used for sensing the ambient temperature only and hence a thermistor with low power sensitivity is sufficient for this purpose.

The interest of Mr. P. H. W. Jacoby in the work is thankfully acknowledged.

Yours faithfully,

R. Krishna,
Industrial and Research Electronics Pty,
Sydney, Australia.

A Simple Frequency Meter

Dear Sir,—In connexion with some work on gaussian noise, a requirement arose for a simple frequency meter for measuring the zero Crossing Rate. In view of the fact that the circuit may be used for sinusoidal and other waveforms, readers may be interested in the details.

S. O. Rice[1] has shown that for Gaussian noise the zero-crossing rate $N_{(0)}$ can be derived theoretically from the characteristics of the band limiting filter. For a filter whose gain-frequency characteristic may be represented by $G(f)$,

$$N_{(0)} = 2\left[\frac{\int_0^\infty f^2 g(f)df}{\int_0^\infty g(f)df}\right]^{\frac{1}{2}} \quad \text{......} (1)$$

An ideal band-pass filter of limits f_1 and f_2 would yield

$$N_{(0)} = \sqrt{\tfrac{2}{3}}\left[\frac{f_2^3 - f_1^3}{f_2 - f_1}\right]^{\frac{1}{2}}$$

which simplifies to $N_{(0)} = 1{\cdot}155\, f_2$ for an ideal low-pass filter.

It can be seen that the denominator within the brackets of equation (1) represents the area of the power gain-frequency response curve of a filter, a value which may be simulated by the output of a thermocouple activated by band limited noise.

If another thermocouple is fed from the same noise source through a capacitor then since the capacitive reactance varies inversely as the frequency, the voltage appearing across the thermocouple will vary directly with frequency. The output over a narrow frequency band will correspond to an input pro-

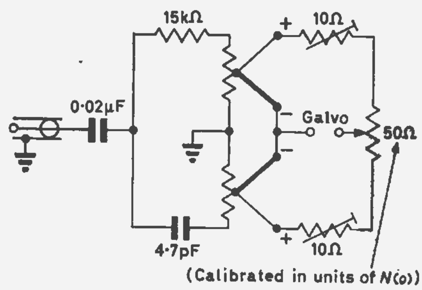


Fig. 1. Simple frequency meter

portional to (voltage)^2 and hence also to (frequency)^2. The total output of the capacitively fed thermocouple will therefore be proportional to the numerator within the brackets of equation (1).

The outputs of the two thermocouples may be fed to each end of a potentiometer whose wiper is returned to ground via a galvanometer. The wiper knob runs over a calibrated scale. The device operates correctly for sine waves as well as with wide band noise and hence the scale may be calibrated with sine signals from a standard signal generator. Once calibrated, of course, the instrument may be used as a direct reading frequency meter. The accuracy, of course, is not very high and depends upon the galvanometer sensitivity and the type of potentiometer used. In the original application using noise signals, this was not a serious problem.

Owing to the limited sensitivity of the thermocouples, the presence of stray capacitance and the finite value of thermocouple resistance, the device has a somewhat restricted range, but it is envisaged that further development would produce a more generally applicable instrument. Using the circuit values shown, the circuit was used at R.R.E. to measure zero crossing rates from 3 to 20 per μsec, but the two 10Ω rheostats allow adjustment of the range of interest.

Yours faithfully,

Brian A. Wyndham,
Royal Radar Establishment,
Great Malvern,
Worcestershire.

REFERENCE

1. Rice, S. O. Mathematical Analysis of Random Noise. *Bell Syst. Tech. J.* *23*, 282 (1944) and *24*, 46 (1945).

Phase Inversion in an Overloaded Amplifier

Dear Sir,—The phenomenon described by P. C. Pugsly in 'Phase Inversion in an Overloaded Amplifier' (October issue), has been noticed in stabilized power supplies when endeavouring to adjust them beyond their design limit. The phase inversion may be used with advantage to generate unidirectional pulses from transitions of a waveform as shown in Fig. 1.

The base is driven by a source with a low impedance compared to R and the transistor response is fast compared to the transition time. The width of the base of the triangular output is

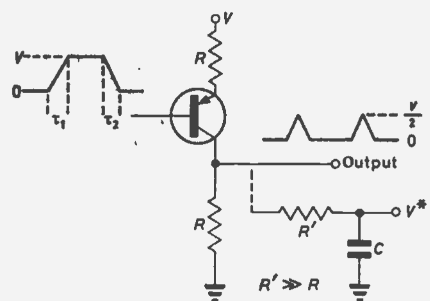


Fig. 1. Generation of unidirectional pulses

equal to the transition time of the input. Integration of the triangular pulses will give a measure of the average transition time, i.e. in Fig. 1, $V^* \alpha (\gamma_1 + \gamma_2)$.

Yours faithfully,

D. J. Grover,
Marconi Instruments Ltd,
St. Albans.

NEW EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

(Voir page 825 pour la traduction en français; Deutsche Übersetzung Seite 832)

SCHOOLS OSCILLOSCOPE

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustrated below)

Advance Electronics Ltd announce a new oscilloscope for the education and industrial markets. The oscilloscope—designated OS12—is designed to Nuffield specification 158 and offers special features including an autosync time-base circuit, which locks automatically, a direct coupled amplifier and, with the time-base off, an X input on the rear of the instrument which permits an X-Y dipslay.

The OS12 offers a sensitivity of 100mV per graticule division and time-base speed variable from 100msec to 100μsec per division. It has a 2¾in tube.

A very simple construction technique has been used so that the oscilloscope is built of only two pre-shaped pieces

of aluminium. Valves and components are mounted on a single printed circuit board. Advance expect this oscilloscope to achieve extensive use in industry as a built-in monitor for systems and production test equipment.

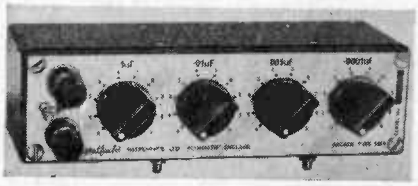
EE 01 751 *for further details*

CAPACITOR DECADE BOX

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devonshire

(Illustrated above right)

A new compact capacitor decade box (type 688/A) has been developed by Hatfield Instruments Ltd for use by design engineers in circuit tolerancing and in similar applications. The decade provides a rapid means of capacitor selec-



tion over the range 100pF to 1μF. The unit occupies minimum bench space, even when used in multiples, and accuracy is better than 5 per cent at any setting. Four decades give steps of 100pF, 1 000pF, 0·01μF and 0·1μF. Minimum capacitance with all switches set to zero is 30pF and voltage rating is 250V d.c. (0·1μ decade, 100V d.c.). Size: 5½in × 1¾in × 2¼in.

EE 01 752 *for further details*

EDGE CONNECTORS

Ferranti Ltd, King's Cross Road, Dundee

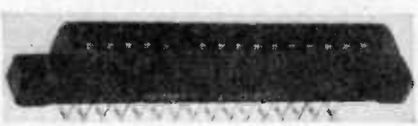
(Illustrated below)

Ferranti Ltd has introduced a new range of edge connectors for printed circuit boards which provide high reliability coupled with a deceptively simple design. Known as the EWD range of connector's they are available with either single-sided or double-sided contacts.

Models with single-sided contacts employ glass-filled nylon bearing inserts on the non-contacting sides in place of dummy contacts. These unique inserts provide excellent long-wearing low-friction surfaces for the printed circuit board, and contribute to the highly competitive prices of the connectors.

The connector mouldings are high quality blue diallyl phthalate with excellent electrical and mechanical properties, completely safe due to its self-extinguishing properties, and able to withstand soldering temperatures with ease.

A feature of the contact design is the rolling leaf spring which is inherently stress limiting so that overstressing cannot occur. In common with all Ferranti connectors, the contact spring has a low rate characteristic, which means that deflexions occurring in the operating range result is only small variations in

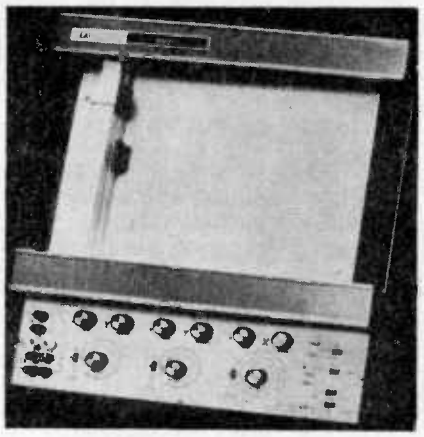


contact force, while insertion and withdrawal forces are notably low. Normal board thickness tolerances are readily accommodated, and controlled contact forces are obtained sufficient to disrupt tarnish films and ensure low and stable contact resistances no matter how low the working voltage may be.

Connectors from the new EWD range are supplied with solder spills in 8, 16, 24, 32 or 40 pole positions for single-sided and 8/8, 16/16, 24/24, 32/32, or 40/40 pole positions for double-sided. They are also designed to accept printed circuit boards with a nominal thickness of 0·062in. Pole pitch is 0·150in. Board penetration is 0·370in and insertion of the board into the connector is facilitated by the generous lead-in-angle.

Contacts are beryllium copper, hard gold plated to a thickness of ·0001in minimum giving a non-porous and durable surface.

EE 01 753 *for further details*



THREE INPUT PLOTTER

Electronics Associates Ltd, Burgess Hill, Sussex

(Illustrated above)

The series 1131 Variplotter is designed for applications requiring simultaneous plotting of two variables against a third.

Three independent servo-drive systems, basic sensitivity of 1mV/in, 18 calibrated ranges, and steplessly variable scale-factor potentiometers, provide a high degree of flexibility.

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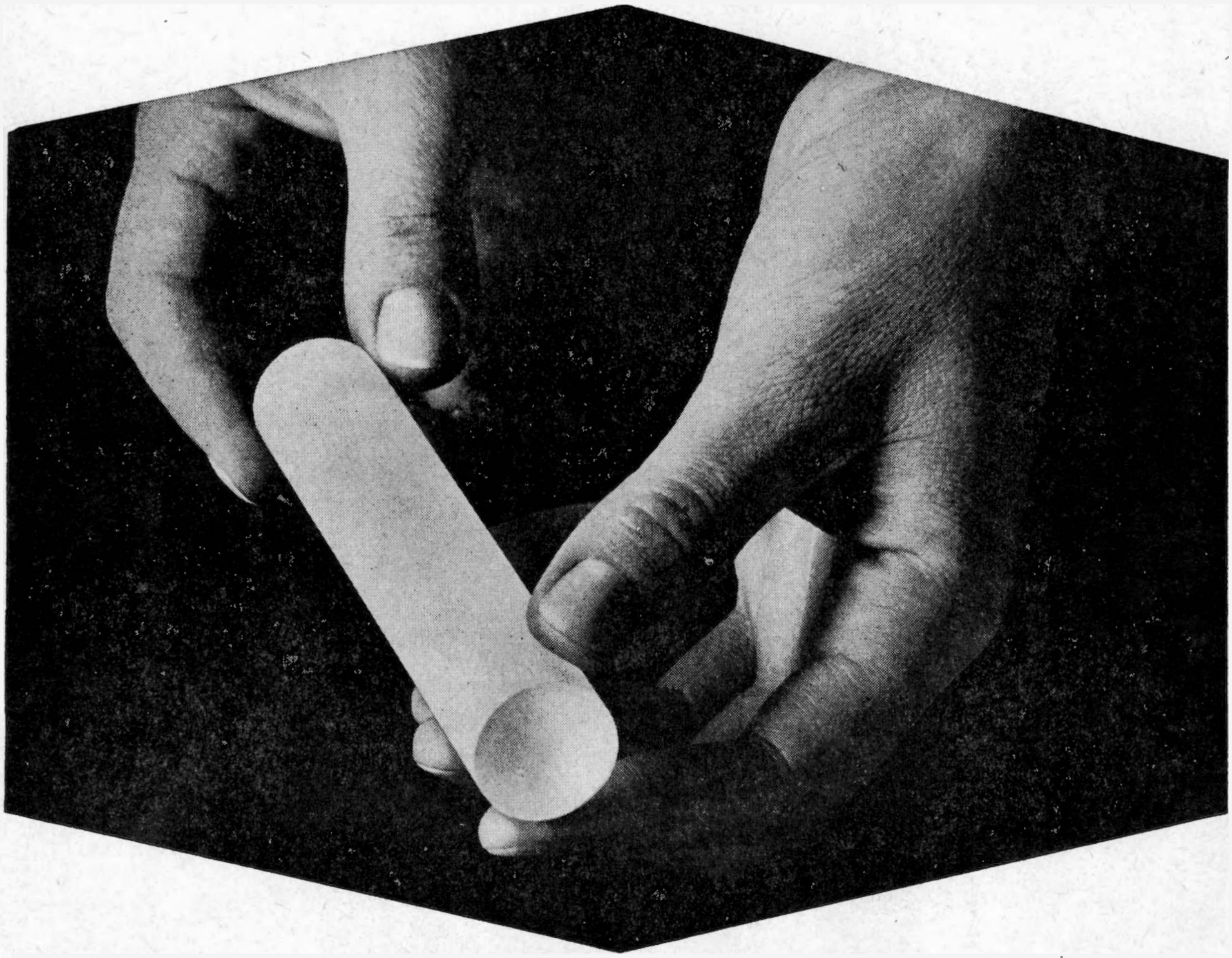

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EE 01 092 *for further details*

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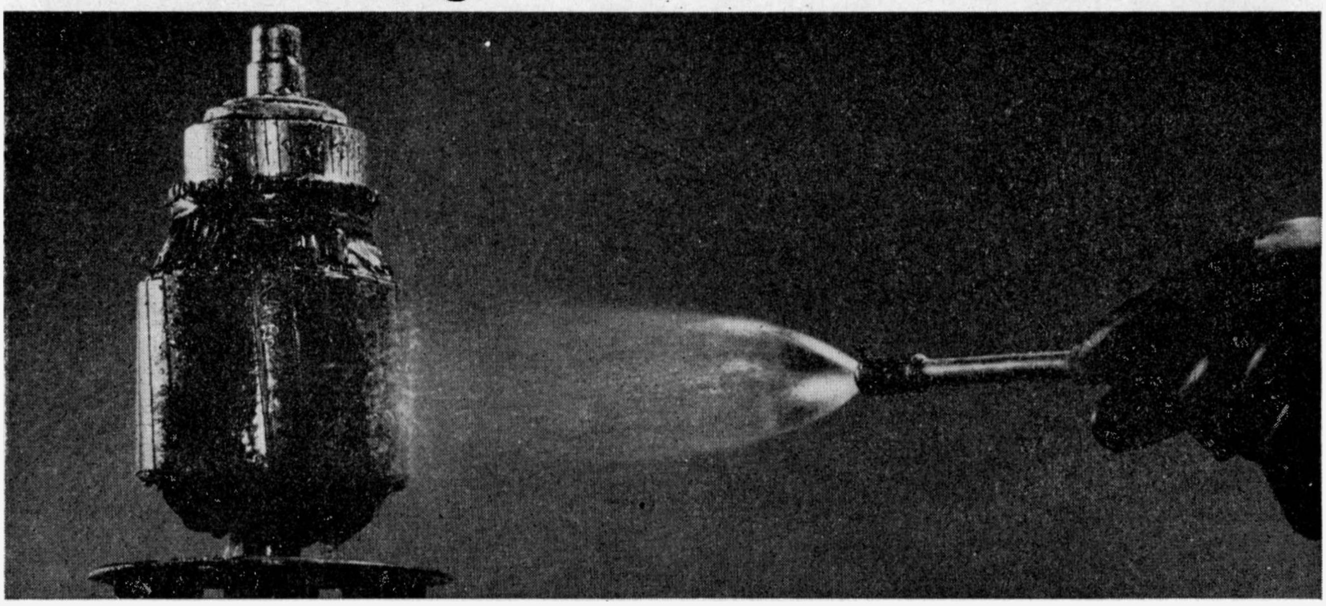

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diode reference system, the 1131 features leak-proof plug-in ink cartridges, back-lighted control switches, plug-in filters for line frequency and noise suppression, and blower type paper hold-down.

Specification is as follows: static accuracy, ±0·1 per cent; dynamic accuracy ±0·2 per cent at 7in/sec; repeatability, ±0·05 per cent. Slewing speed is 17in/sec for the arm and 20in/sec for the pens. Input resistance is 2·5MΩ/V on ranges up to 20mV/in, and 1MΩ constant on higher ranges. Available with either imperial or metric calibration, the unit can be either rack or deck mounted, and overall dimensions are 19in by 21in high by 6·5in deep.

EE 01 754 *for further details*

STATIC TIME DELAY RELAY

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustrated below)

Solid State Controls Ltd is now manufacturing a plug-in transistorized time delay relay of high reliability and economic cost which has been specially designed to meet the needs of modern industry.

The TDR/407 is a robust device, completely encapsulated, with no moving parts to wear or fail. It is unaffected by vibration, shock and transient voltages, and is capable, at a minimum, of 100 million continuous operations. This compact unit is completely free from drift, even after many thousands of hours in service, and requires no maintenance whatsoever.

TDR/407 is available in three voltage ranges, 12-35, 35-70, 70-150V, and functions accurately even when subject to large variations in operating voltage and ambient temperature. The main timing circuit consists of a completely stable, low impedance delay circuit with a silicon control rectifier output switch. This semiconductor contact will switch inductive or resistive loads of up to 1A at 150V and is capable of tolerating large in-rush currents normally associated with contactor or motor switching circuits.

Time delay periods in four ranges from 0·2sec to 11min are available, and adjustments to delay times may be made by the control mounted on the case, or may be pre-set. The timing accuracy when subject to extremes of ambient temperature and operating voltage is 5 per cent typical, and repetitive accuracy 1 per cent typical.

The high reliability and solid state nature of the unit make it particularly well suited to aircraft applications as well as meeting industrial requirements.

EE 01 755 *for further details*

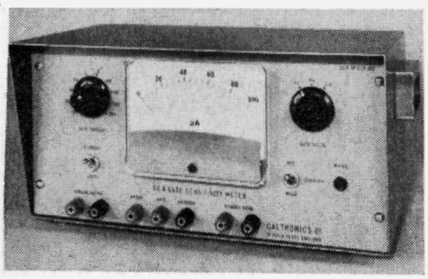
S.C.R. GATE SENSITIVITY METER

Caltronics Ltd, Huntingate, Hitchin, Hertfordshire

(Illustrated below)

This s.c.r. gate sensitivity meter provides accurate and rapid measurement of the gate current-to-fire for a wide variety of silicon controlled rectifiers. A Zener diode provides a stabilized anode-to-cathode voltage of 6V to the s.c.r. under test.

The gating characteristics of the s.c.r. under test are measured by applying half-wave rectified 50 to 60c/s pulses between the gate lead and cathode of the device. A trigger network in the anode supply lead senses the turn on point of



the s.c.r. under test and energizes an electronic switch that removes the gating signal. A peak reading voltmeter circuit is utilized to give a direct indication of the current or voltage level at which the s.c.r. fires. The gate current-to-fire reading is obtained by driving the gate circuit through a set of precision resistors which form an adjustable 10 step current source. The gate voltage-to-fire reading is obtained by driving the gate circuit from an adjustable three step voltage source.

Calibration potentiometers allow the sensitivity meter to be calibrated on both current and voltage.

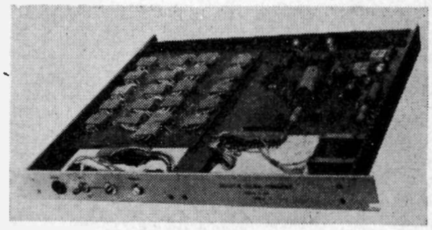
Terminals are provided for parallel remote operations of the instrument. Remote indication of the meter reading may be obtained from a pair of panel terminals which provide a +1V signal for full scale indication on the front panel meter.

EE 01 756 *for further details*

ANALOGUE-TO-DIGITAL CONVERTOR

Ether Engineering Ltd, Park Avenue, Bushey, Hertfordshire

(Illustrated above right)



The model MA 1102 convertor changes analogue input voltages of between 0 and −10V to an equivalent 10-bit pure binary coded output.

The instrument utilizes the successive approximation method to provide accurate a.d.c. facilities at a low cost. A substantial part of the circuit employs thin-film dual NOR modules. These are connected, according to circuit requirements as gates, bistable and monostable elements, or invertors.

Compact mechanical design facilitates embodiment of the convertor into existing systems, the unit only occupying 1¾in of panel height.

The full-scale analogue input is 10V, corresponding to 2^{10} digits. The resolution is ±1 digit and the accuracy 0·1 per cent. The conversion time is 200μsec.

EE 01 757 *for further details*

NOISE REDUCTION SYSTEM

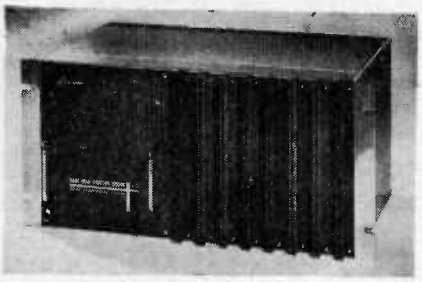
Dolby Laboratories, 590 Wandsworth Road, London, S.W.8

(Illustrated on page 820)

The Dolby Laboratories S/N Stretcher provides a method of combatting noise in high quality audio transmission and recording systems. The S/N Stretcher can be used in any situation in which the signal is available for processing at both the input and output of the audio chain. While the signal itself emerges in unaltered form, the noise reduction action attenuates the usual types of noises encountered in mastering and dubbing, motion picture production, video tape recording, landlines, etc.

Because of the well known masking effect in human hearing, quiet sounds are masked partially or even completely by loud sounds, especially if the frequency differences are not too great. Thus subjective signal-to-noise ratios in magnetic tape recording are better than would be indicated by analysis of the tape playback waveform; tape noise depends on and increases with the instantaneous amplitude of the signal, an effect known as modulation noise. The fact that the hearing mechanism effectively suppresses this additional noise under normal signal conditions suggests that a suitably designed electronic system should be able to employ the masking phenomenon to obtain an even further reduction of noise.

Utilizing the masking principle, together with signal compression and expansion, the S/N Stretcher achieves noise reduction (a) by boosting low level signal components during recording whenever possible (compression), followed by complementary attentuation during playback (expansion), and (b) by the masking effect whenever the signal level is



so high that compression and expansion are not possible.

Since masking is less effective with noise frequencies somewhat removed from the signal frequency, it is necessary to deal with the various portions of the spectrum independently. The noise reduction system then yields a lower—and apparently constant—noise level, the classical hush-hush or swish of normal compression and expansion being absent.

The S/N Stretcher splits the audio spectrum into four bands and compresses and expands each of these in an essentially independent manner. Separate bands are provided for the hum and rumble frequency range, for the mid-audio range, for medium high frequencies, and for high frequencies. A high level signal in one band hence cannot prevent noise reduction in another band in which the signal level may be low.

From another point of view, the system effectively produces a recording equalization characteristic which continuously conforms itself to the incoming signal in such a way as to optimize the signal-to-noise ratio during playback.

EE 01 758 *for further details*

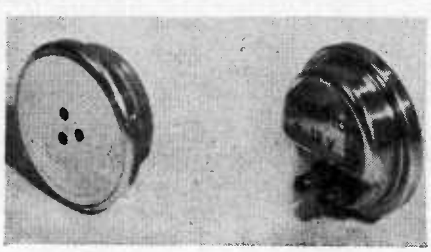
SONIC SIGNAL SOURCE

A. P. Besson & Partner Ltd, St. Josephs Close, Hove, Sussex

(Illustrated below)

The 'Bleeptone' has been developed to provide a compact and highly reliable sonic signal source. It comprises a self-contained oscillator circuit and rocking armature earphone designed to operate from a low voltage battery. In dimensions and appearance it is identical to the well known rocking armature receiver as used in modern telephone instruments. The note emitted can normally be heard at distances up to 500ft and the frequency is such that background noise has no appreciable masking effect.

The drive circuit is similar to that used for self-maintained tuning forks. Two separate coils are arranged in the magnetic circuit, one to drive the rocking armature and the other to pick up the induced signal and provide feedback to the base of a transistor which together with associated components forms the complete amplifier circuit.



The fundamental frequency of operation is determined basically by the mechanical characteristics of the rocking armature.

This device is intended for application in a wide variety of industries. It is particularly suitable in the role of warning note especially where a higher degree of reliability is required than can be obtained with a conventional buzzer.

EE 01 759 *for further details*

RADIO TELEPHONE AMPLIFIER

The Plessey Co. Ltd, Ilford, Essex

(Illustrated below)

A new transistorized amplifier for use in radio telephone equipment has been developed at the Bridgnorth factory of the Plessey Electronics Group.

Designed originally for the Company's type 700 and 700R radio telephone equipments, these amplifiers with a gain of 14dB can increase power output to over 8W and are equally suitable for incorporating into existing radio telephone links. They measure only 3in high, 6¼in wide and 3in deep.

In addition to the basic version, there is also a cased unit available which is

hermetically sealed and fully tropicalized for outdoor use in any part of the world.

Suitable for 12V d.c. operation, this equipment is also available for operation over frequencies of 68 to 174Mc/s.

The power consumption of approximately 1·6A makes the new amplifier suitable for adding to mobile systems or as power sources to drive varactor multipliers up to the gigacycle per second regions.

EE 01 760 *for further details*

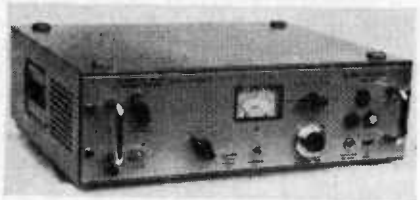
POWER SUPPLIES

Distributed by: Scientific Measurements & Equipment Ltd, 78 Main Street, Queniborough, Leicester

(Illustrated above right)

Manufactured by Knott Elektronik, Munich, these supplies are designed for use in conjunction with electronic equipments which require extremely constant and accurate d.c. voltages such as G.M. tubes, proportional counters, ionization chambers, image intensifiers, photomultipliers, scintillation counters, cathode-ray tubes, etc.

The high output current and very low internal resistance allows simultaneous



supply to several units without interaction. The salient features of these supplies are their high stabilization ratio, long term stability, adjustable fuse setting and adjustability to within one volt. They are supplied cased, or for rack mounting.

The range of units available covers requirements from 0 to 30V at 10A 0 to 15kV at 10mA. A stabilized current unit supplying 0 to 10A at 13V is also available.

EE 01 761 *for further details*

A.C.-D.C. MILLIVOLTMETER

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(Illustrated below)

The TM1 is a fully transistorized millivoltmeter having 12 ranges for both a.c. and d.c., covering from 1mV to 300V f.s.d. in a 1-3-10 sequence; a decibel scale is also provided covering −10 to +2dB (0dB = 1mW in 600Ω).

On d.c. the accuracy is 3 per cent of f.s.d. ±100μV and on a.c. it is 4 per cent of f.s.d. ±100μV from 20c/s to 100kc/s. The d.c. input resistance is 1MΩ/V up to 10V and constant at 10MΩ on higher ranges: the a.c. input impedance is 100kΩ from 1mV to 30mV, 1MΩ from 100mV to 300mV and 10MΩ from 1V to 300V; the maximum shunt capacitance is 40pF.

For d.c. the millivoltmeter has a resistance input attenuator, followed by a d.c. amplifier with meter and feedback control. In the a.c. section a frequency-compensated attenuator feeds an a.c.-to-d.c. amplifier, which contains a feedback loop and meter. In addition, on the lowest four ranges, an impedance-converting amplifier maintains the input, impedance at 100kΩ.

Two PP11 batteries are used by the equipment, and there is a built-in battery voltage check. Weight, less batteries, is 6 lb 5 oz, and dimensions 165mm high by 218mm wide by 220mm deep.

EE 01 762 *for further details*

THERMAL RECRYSTALLISED ALUMINA TUBES

Thermal Recrystallised Alumina is the most generally suitable refractory material at temperatures above the useful range of Vitreosil pure fused silica.

Fully recrystallised under controlled conditions, this pure impervious material is available in a wide range of sizes and shapes for furnace and laboratory applications at high temperatures.

Other refractory materials available include Thermal Aluminous Porcelain 525, and for more specialised applications, Thermal Magnesia, Thermal Thoria and Thermal Zirconia.

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EE 01 094 *for further details*

*** Large 17" display**

*** 1 Millivolt/cm sensitivity**

*** 3 input channels**

The Knott Display Oscilloscope "Sweep Skanner" has been specifically developed for working with sweep generators and provides the development engineer and production inspector with high resolution when used with sweep display systems.

The "Sweep Skanner" has a 17" large screen with two identical vertical input channels, calibrated sensitivities from 1 mV/cm to 1 volt/cm at an accuracy of $\pm 3\%$. A third input channel is provided which has a variable sensitivity from 25 mV/cm.

Two DC levels are incorporated to enable selection of DC references at any point on the oscilloscope—3db and 1db points of a filter response.

The price of the Knott "Sweep Skanner" is £414 0s 0d (excluding duty), with ex-stock delivery.

Please fill in the attached reply coupon for additional information or a demonstration.

THE Livingston GROUP

☐ I would like a demonstration of the Knott 'Sweep Skanner'.

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LIVINGSTON LABORATORIES LIMITED
Livingston House, Greycaine Road,
North Watford, Herts. Tel: Watford 44344

V.S.W.R. INDICATOR

Marconi Instruments Ltd, Sanders Division, Gunnels Wood Road, Stevenage, Hertfordshire

(Illustrated below)

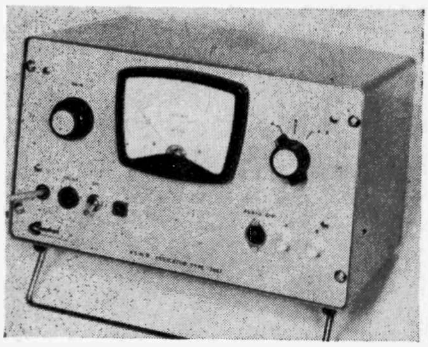
A new v.s.w.r. indicator type 6596 which is a low-noise, low cost selective amplifier has been introduced by the Sanders Division of Marconi Instruments Ltd. Its sensitivity and gain range make it suitable for the majority of microwave measurements where crystal detectors are employed and the r.f. signal is modulated at 1kc/s.

Coarse and fine gain controls are provided with sufficient variation to accommodate detected signals up to 1mV r.m.s.

The meter scale is calibrated in v.s.w.r., based on the assumption that the detector is working in the square-law region of its characteristic—i.e. output voltage proportional to microwave power.

Maximum sensitivity is better than 1μV f.s.d. and noise is less than 0·1μV, and an input selector switch enables either of two input signals to be selected individually. Also the difference of the two may be selected for extra sensitive microwave bridge techniques.

This unit is ideally suitable for operation in conjunction with the Sanders type 599 educational test bench.



EE 01 763 *for further details*

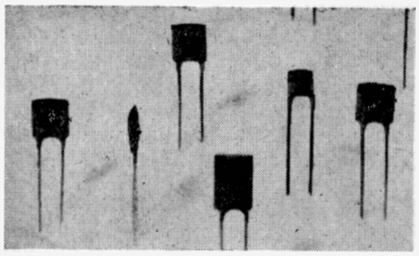
MINIATURE CERAMIC CAPACITORS

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

(Illustrated above right)

A new series of high-quality miniature ceramic capacitors announced by Mullard are rectangular in shape and are only 1·9mm thick to allow high packing densities on printed circuit boards using a 0·1in grid. The capacitors are suitable for use in both domestic and industrial equipment. They consist of a thin plate of metallized ceramic material which is insulated with a protective lacquer to assure excellent performance under the most humid conditions.

Each capacitor in the new series (C333) measures 5 × 8·5 × 1·9mm (excluding leads of 13·5mm). The capacitance range is from 3·9pF to 150pF with a tolerance of 0·5pF or ±2 per cent whichever is greater. Working voltage is 40V over a temperature range of −25°C to +85°C and the insulation resistance measured at 10V is greater



than 1000MΩ. The temperature coefficient varies between zero and negative 750 parts per million depending on the capacitance value.

The close tolerance and high stability of the new capacitors make them particularly suitable for use in television i.f. transformers, tuned circuits and other applications calling for low-loss, and high performance.

EE 01 764 *for further details*

PRECISION BRIDGE

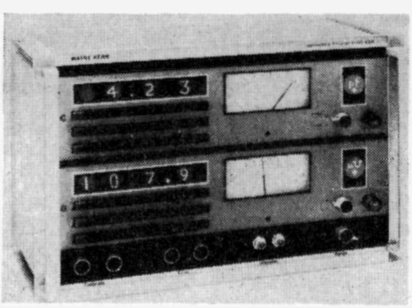
The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated below)

Latest addition to the Wayne Kerr range of Autobalance bridges, model B331, combines the high accuracy demanded by standard laboratories with the speed and simplicity of operation given by electronic nulling. Two meters provide simultaneous readings of the in-phase and quadrature terms of any component or complex impedance. Pushbuttons, associated with illuminated in-line displays, permit both readings to be backed-off by three decades, giving up to six-figure resolution on all ranges.

Special circuits are built-up to compensate automatically for the impedance of the measurement leads which are terminated in an advanced type of Kelvin clip. Outputs are provided for operating digital voltmeters, printers, recorders, pass/reject mechanisms or control circuits. Sockets for external standards permit comparative measurements and, used with the vernier controls, a discrimination of 10 parts per million can be realized.

The internal source and detector operate at 1·00kc/s but operation can be from 50c/s to 20kc/s using external equipment. Overall measurement capability at 1kc/s is ·0001pF to 0·25pF, 1pΩ⁻¹ to 1kΩ⁻¹ 1mΩ to 1TΩ and 100nH to 250 MH. The full accuracy of 0·01 per cent applies from 1pF to 10μF and 10nΩ⁻¹ to 100mΩ⁻¹. The B331 is 19in wide × 12in high × 9in deep and weighs approx. 50lb.



EE 01 765 *for further details*

REED RELAY

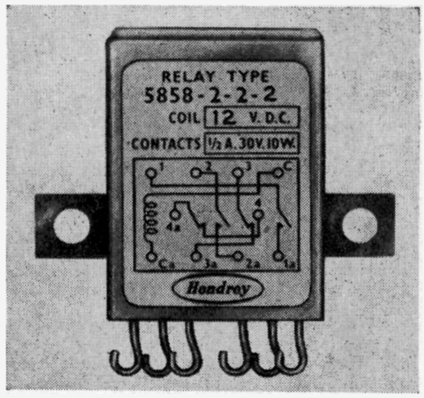
Hendrey Relays & Electrical Equipment Ltd. 390-394 Bath Road, Slough, Buckinghamshire

(Illustrated below)

Designed to meet the requirements of B.S. 2G.100, this reed relay type 5858 is suitable for ground or airborne use, with limiting values of 5c/s to 2kc/s at 10*g* and acceleration 50*g* applied in any direction. The energizing coil and reed units are encapsulated in a steel case which provides both mechanical and magnetic protection. The encapsulating material is an epoxy resin having excellent electrical characteristics and good temperature performance. The resin also forms the base of the relay carrying the connexion pins, thus forming a very robust unit unaffected by humidity and suitable for use in the most adverse environmental conditions in ambient temperatures from −60°C to +120°C.

Carefully selected reeds and pre-use forming guarantees a minimum of 2 × 10^6 operations on full load and response time of the contacts is better than 2msec, with negligible bounce both making and breaking.

Maximum rating is 0·5A or 30V d.c. or 10W resistive loads. Inductive loads



should be suitably quenched. Three standard contact arrangements are provided: (a) two normally open; (b) four normally open; (c) two normally open and two normally closed. Coils can be wound for any d.c. voltage up to 100V d.c.

The relay is available with solder hook terminations or with flying leads, bracket or strap mounting; or with pins for plug-in-bases.

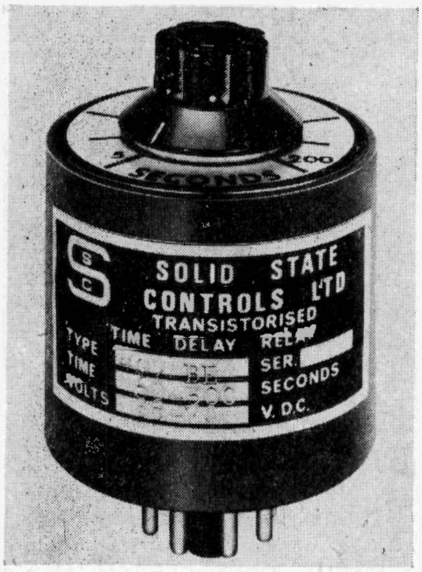
EE 01 766 *for further details*

DELAY SWITCH

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustrated on page 822)

The 'Radian' is a low cost, high performance replacement for thermal delay switches, simple capacitor/relay timers and synchronous motor units where high repetitive accuracy, long-term stability and a fast recovery time is essential. The Radian relay delay requires no cool down period unlike thermal units, but resets instantaneously and gives highly consistent timing operations.



The operating voltages available are 24V d.c. and 24V a.c. while the output will operate a wide range of relays and contactors.

The Radian is a plug-in unit which provides two types of operation; delay on energize and delay on de-energize with two timing ranges, 1 to 60sec and 1 to 5min.

The unit is unaffected by fleeting transients on the supply line, by fluctuations in operating voltage of ±15 per cent, and variations in ambient temperature of −10°C to +55°C.

EE 01 767 *for further details*

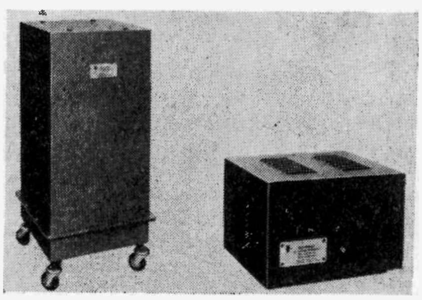
CONSTANT VOLTAGE SUPPLIES

Servomex Controls Ltd, Crowborough, Sussex

(*Illustrated below*)

Servomex Controls Ltd has developed versions of the voltage stabilizers types AC2 and AC7 for use in applications where the rack-mounting system is not required. These new types, models AC2 Mk 111A and AC7 (industrial), are more convenient to install and to service, having been designed for installation by factory electrical staff as distinct from electronic personnel. All the parts which require periodical attention are easily accessible by removing the cover without its being necessary to move the instrument to obtain access to the back as is usually the case with the rack-mounting models.

The individual electrical components are identical with those of the rack-mounting models as are the quality and workmanship employed in the construction of these new types. Model AC2 Mk 111A is smaller and handier than the rack-mounting type, while model AC7 (industrial) stands on castors on the floor and is hoseproof and dustproof. It is therefore ideal for installation in a cellar or in a much swept passage.



Both instruments operate on an input supply in the range 200 to 250V and provide a highly stable output voltage not varying by more than ±0·1 per cent of the nominal voltage of the supply. They are unaffected by rectifier loads or by changes in power factor from resistive to pure reactive, including capacitive loads.

The method of operation is by means of a variable transformer which is used to supply a buck or boost voltage as may be necessary to correct the input to the required value. This variable transformer is driven by a two-phase servo motor with enclosed gearing and torque limiting device. Variations of the output voltage are detected by a bolometer bridge, the output of which is amplified in a two-valve servo amplifier and applied to the motor.

The instruments are of exceptionally strong construction and can stand impacts of up to 40*g* in any direction. No relays, thyratrons or electrolytic capacitors are used.

The new types are cheaper not only on account of the simpler cases used but also because the voltmeter, ammeter, on/off switch and main fuse are omitted. These last two items are meant to be mounted on the wall with an external voltmeter as an optional item.

EE 01 768 *for further details*

DIGITAL STRAIN INDICATOR

S.E. Laboratories (Engineering) Ltd. Astronaut House, Feltham, Middlesex

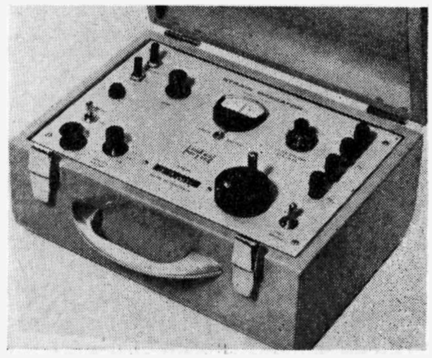
(*Illustrated above right*)

The digital strain indicator type SE.601 is a portable instrument for the measurement of dynamic and static strain incorporating many new features. The digital read-out ensures maximum operator convenience and, with the addition of an 'add-on' switch, covers a total range of ±50000 microstrain with an accuracy of ±0·1 per cent of reading or ±5μE—whichever is greater (G.F. = 2·0).

Internally fitted rechargeable cells enable the unit to be used in the field for long periods. A further feature is the 'gauge factor' control which can also be used to obtain direct indication from devices such as load cells, pressure and displacement transducers, etc.

Square wave excitation enables the unit to be used with strain gauge circuits having up to 0·01μF capacitance without any appreciable loss of accuracy, while sinusoidal excitation may be provided if required.

The unit is suitable for use with gauges having a resistance of 50Ω to 2kΩ. Outputs of 1mA and 2V can be obtained



from separate galvanometer and oscilloscope jack sockets to drive conventional read-out devices or recording instruments.

EE 01 769 *for further details*

SELF-HOLDING SOLENOIDS

H. E. & B. S. Benson Ltd, Exning Road, Newmarket, Suffolk

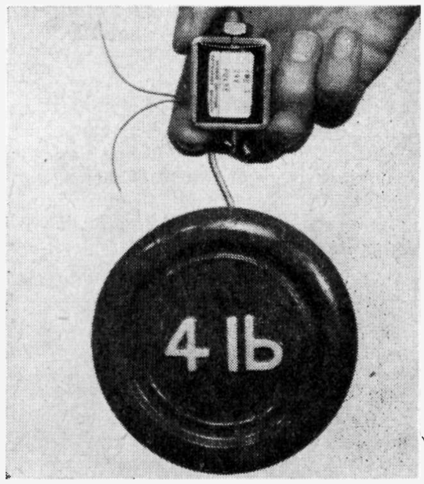
(*Illustrated below*)

H. E. & B. S. Benson Ltd announce an entirely new self-holding solenoid. The plunger is of permanent magnet material and, when closed from a brief pulse, will remain closed for an unlimited period, sustaining a load without drawing current. A reverse pulse releases the plunger.

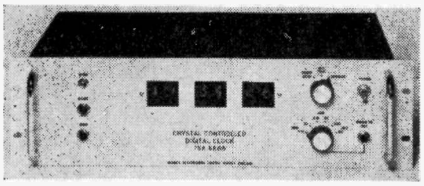
Any form of rectified a.c. or d.c. (including a dry battery) will operate such a solenoid. There is an a.c. circuit which needs only a diode and a small resistor, which can be soldered directly to the operating switch.

Advantages include zero temperature rise for maximum reliability (also useful in a high ambient); a wide range of force characteristics, all with continuous hold from momentary switching and maintained hold in spite of supply failure.

The PMB self-holding solenoids are externally identical with the existing B.2, 3, 4 and 5 types. PMB's can replace latching solenoids and, when fitted with a return spring, can give the definite two-position action of a more costly dual-coil double-acting unit.



EE 01 770 *for further details*



DIGITAL CLOCK

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

(Illustrated above)

Venner Electronics Ltd has introduced an all-silicon digital clock, type TSA 6686, using glass fibre printed circuit boards and giving a display in hours, minutes and seconds derived from six gas-filled numerical indicator tubes. Two versions are available, giving time on a 12 or 24 hour basis as required. Electrical outputs are available for driving a printer or a punched tape unit.

Start/stop facilities are provided either by negative-going pulses of 4V amplitude or by the operation of push-buttons mounted on the front panel. The display can be manually set for the independent showing of hours, minutes and seconds, and has a single reset to zero.

Designed to operate over the ambient temperature range of $-5°C$ to $+60°C$, the clock has an accuracy of two parts in 10^6.

Power supply is 120 to 125V or 200 to 250V, 50c/s a.c. the size of the TSA 6686 is 5¼in high, 9½in deep, 19in rack mounting.

EE 01 771 *for further details*

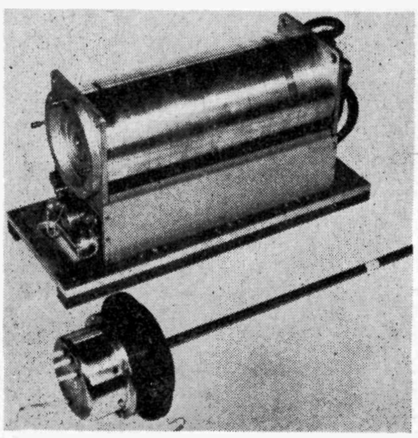
TRAVELLING WAVE TUBE

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Illustrated below)

The M-O Valve Co. Ltd has introduced a new high power pulsed travelling wave amplifier, type TWX16, suitable for radar applications. It gives over 40dB gain at 5kW peak output power (30W mean power and 5 to 20kW saturated peak power) over a bandwidth of 500Mc/s in the range 8 to 9·3Gc/s.

The tube is of metal-ceramic construction and uses a ring and bar slow wave structure giving freedom from unwanted oscillations. Focusing is achieved by operation in a solenoid mount assembly, type SMX16, which incorporates the r.f. couplings and provides conduction cooling of the tube.



The TWX16 is designed to operate with the helix maintained at d.c. potential to simplify pulse power supply requirements.

The M-O V. TWX8 1-watt travelling wave tube is suitable for use as a driver.

EE 01 772 *for further details*

KLYSTRON POWER SUPPLY

Distributed by: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(Illustrated below)

The Oltronix (Sweden) LS 525R is a very stable d.c. power supply designed for operating reflex klystrons. Beam, reflector, grid and heater voltages are separately regulated. The reflector voltage may be unmodulated for c.w. operation of the klystron or it may be modulated either with an internally generated square wave, pulse sawtooth or sine wave, or with an external signal. The grid voltage may be internally modulated with square wave or pulse. Square wave and pulse modulation voltages are clamped to the c.w. level of the grid or reflector.

The beam voltage is variable from

$-200V$ to $-3600V$ at 0 to 125mA. A 10 per cent line voltage variation causes a change of less than 300mV in the output while a change from no load to full load causes a variation of less than 200mV.

The output is continuously variable in four ranges: 200 to 1 200, 1 000 to 2 000, 1 800 to 2 800 and 2 600 to 3 600V and is completely protected against overloads and shorts by an electronic fuse adjustable from 10 to 125mA.

The equipments are also suitable for photomultiplier applications, it being possible to operate over 100 photomultipliers from the same supply.

In the illustration the LS 525R is shown at the top, with an additional unit for driving klystrons at the bottom.

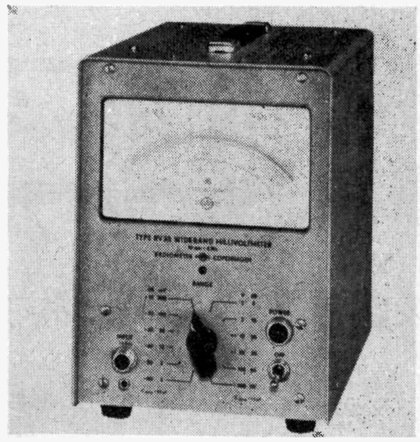
EE 01 773 *for further details*

WIDEBAND MILLIVOLTMETER

Distributed by: Livingston Laboratories Ltd, Livingston House, Greycaines Road, North Watford, Hertfordshire

(Illustrated above right)

Radiometer of Copenhagen has now introduced a successor to its valve-voltmeter RV 33. This is the new wideband



millivoltmeter, type RV 35, which gives overload protection up to 500V a.c. or d.c., and measures voltages from 10µV to 300V. Accuracy is 2 per cent within the greater part of the frequency range, which is 10c/s to 6Mc/s.

An additional feature of the RV35 is that it can be used as a high-gain amplifier with a full scale output voltage of approximately 80mV across 75Ω.

EE 01 774 *for further details*

CERAMIC C.R.T.

Ferranti Ltd, Gleo Mill, Oldham, Lancashire

(Illustrated below)

The Electronics Department of Ferranti Ltd has introduced what is believed to be the first ceramic cathode-ray tube with complete electrostatic focus. Designed to withstand severe environmental vibration, the new tube combines very high resolution with a useful screen diameter of 0·5in (12·7mm).

The new tube has an overall length of 4·5in (114·3mm) and a body diameter of 0·6in (15·2mm). The face plate has a useful screen diameter 0·5in (12·7mm) and is made from a high quality glass developed specifically for use with the ceramic envelope to which it is sealed by specially evolved techniques. The typical first anode voltage is 300V and the final anode voltage is 8kV. The line brightness, measured with the first anode at 450V and the final anode at 7kV, is of the order of 6000ft lamberts. Complete with scan coils and Mumetal screen, the tube weighs less than 4oz (110g).

The components of the gun and the focusing system are machined to very close tolerances, making it possible to achieve very good definition and a resolution better than 500 lines This enables the same amount of information to be presented on the screen as would normally be displayed by a cathode-ray tube having a much larger screen diameter.

EE 01 775 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.
Date: December 1.
Discussion. Numerically Controlled Machine Tools. (Joint meeting with the Automatic Control Group of the I.MechE. at 1 Birdcage Walk, London, S.W.1, at 6 p.m.)
Date: 8 December.
Colloquium: Miniature Computers.
(Joint meeting with I.E.R.E. Computer Group at the London School of Hygiene and Tropical Medicine, Keppel Street, W.C.1, at 6 p.m.)
Date: 13 December.
Lecture: Computer Control of Heavy and Light Plate Mills.
By: A. T. K. Betts and R. Shepherd.
(Joint meeting with the Automatic Control Group of the I.Mech.E.)
Date: 14 December.
Discussion: Light-Emitting Diodes.
Opened by: A. E. Brewster and C. D. Dobson.
Date: 15 December.
Lecture: Thermoelectric Materials.
By: G. E. Hare and H. J. Goldsmid.

East Midland Electronics and Control Section
Held at: T.I. Building, University of Nottingham.
Date: 6 December.
Time: 6.30 p.m.
Lecture: The Future of World Communication.
By: C. Cherry.

East Anglian Sub-Centre
Held at: Norwich, The Assembly House.
Date: 13 December.
Time: 7.30 p.m.
Lecture: The Choice between Analogue and Digital Computing. Techniques in Electrical Engineering Analysis and Design.
By: R. A. Laws.

Cambridge Electronics and Control Section
Held at: No. 4 Lecture Theatre, University Engineering Department, Trumpington Street, Cambridge.
Date: 8 December.
Time: 8 p.m.
Lecture: Modern Telephone Developments.
By: L. R. F. Harris.
(Joint meeting with I.E.R.E.)
Held at: Electric House, Ipswich.
Date: 6 December.
Lecture: Transistors—The First Encounter.
By: V. H. Attree.

North Midland Centre
Held at: Leeds University, Elec. Eng. Lecture Theatre, Room 152.
Date: 6 December.
Time: 6.30 p.m.
Lecture: Static Electronic Protection.
By: J. B. Patrickson.
Held at: Faraday Lecture at St. George's Hall, Bradford.
Date: 14 December.
Time: 7 p.m.
Lecture: Practical Uses of Nuclear Energy.
By: R. V. Moore.

Sheffield Sub-Centre
Held at: Faraday Lecture at the City Hall, Sheffield.
Date: 7 December.
Time: 7.30 p.m.
Lecture: Practical Uses of Nuclear Energy.
By: E. V. Moore.
Held at: The Sheffield Industries Exhibition Centre.
Date: 14 December.
Time: 6.30 p.m.
Lecture: Some Electrical Transducers.
By: J. Dean.

North Western Centre
Held at: The University of Manchester Institute of Science and Technology. Renold Building.
Date: 13 December.
Time: 7.15 p.m.
Lecture: Research in Education.
By: F. W. Warburton.
(22nd Annual Lecture in co-operation with the Manchester University.)

North Western Education and Training Circle
Held at: The Harris College, Preston.
Date: 12 December.
Time: 6.15 p.m.
Lecture: Energetics.
By: K. E. V. Willis.

South-East Scotland Sub-Centre
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Date: 7 December.
Time: 6 p.m.
Lecture: Field Effect Transistors.
By: J. M. Morrison.
(Electronics and Control Section Joint Meeting with I.E.R.E.)
Held at: Heriot-Watt University, Edinburgh.
Date: 14 December.
Time: 7 p.m.
Discussion: Transistor Circuit Design.
Opened by: J. Morrison.
(Education and Training Circle meeting.)

South-West Scotland Sub-Centre
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Date: 7 December.
Time: 6 p.m.
Lecture: Radiophonic Workshop.
By: C. Brooker.
Held at: Room 24, University of Strathclyde, Glasgow.
Date: 8 December.
Time: 6 p.m.
Lecture: Field Effect Transistors.
By: J. M. Morrison.
(Electronics and Control Section Joint Meeting with I.E.R.E.).

North Staffordshire Sub-Centre
Held at: The Crewe Arms Hotel, Crewe.
Date: 1 December.
Time: 7 p.m.
Lecture: Modern Applications of Closed Circuit Television.
By: V. J. Cooper.
Held at: English Electric Leo Marconi Computers Ltd., Kidsgrove.
Date: 19 December.
Time: 7 p.m.
Lecture: Pattern Processing in the Human Visual System and Some Engineering Applications.
By: C. R. Evans.

Southern Centre
Held at: Farnborough Technical College, Boundary Road, Farnborough, Hants.
Date: 6 December.
Time: 6.30 p.m.
Lecture: The History of Computers.
By: D. J. Truslove.

Southern Centre Electronics and Control Section
Held at: The Lanchester Theatre, Southampton University.
Date: 13 December.
Time: 6.30 p.m.
Lecture: Character Recognition.
By: H. A. Bell.

Western Centre
Held at: The Queen's Building, University of Bristol.
Date: 12 December.
Time: 6 p.m.
Lecture: Lasers and their Uses.
By: J. C. North.
(Joint meeting with the Electronics, Supply and Utilisation Sections).

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Electro-Acoustics Group
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Date: 6 December.
Time: 6 p.m.
Lecture: An Introduction to Acoustic Measurement.
By: F. H. Brittain.

Radar Group
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 7 December.
Time: 6 p.m.
Lecture: The Design of Radio Navigation Stations for Unattended Operation.
By: D. H. Boycott, D. L. Hillman and J. H. G. Huggins.

Joint I.E.R.E./I.E.E. Computer Group
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Date: 8 December.
Time: 6 p.m.
Colloquium: Miniature Computers.

Annual General Meeting of the Institution
Date: 15 December.
7 p.m.: The Presidential Address of Professor Emrys Williams.

Southern Section
Held at: Basingstoke Technical College.
Date: 1 December.
Time: 7.30 p.m.
Lecture: Television Camera Tubes and Image Intensifiers.
By: R. S. Filby.

West Midland Section
Held at: Birmingham University, Department of Electronic and Electrical Engineering.
Date: 8 December.
Time: 7.15 p.m.
Lecture: Fuel Cells.
By: T. M. Fry.

East Midland Section
Held at: Leicester University.
Date: 8 December.
Time: 6.30 p.m.
Lecture: Field Effect Transistors.
By: C. S. Den Brinker.

South Western Section
Held at: University of Bristol, Small Lecture Theatre. University Walk, Clifton, Bristol 8.
Date: 7 December.
Time: 7 p.m.
Lecture: Mediator, (???)
By: D. R. Evans.
(Joint meeting with the Western Centre of the I.E.E. (Electronics Section).

South Midland Section
Held at: BBC Club, High Street, Evesham.
Date: 6 December.
Time: 7 p.m.
Lecture: Radiophonic Workshop.
By: F. C. Brooker.

East Anglian Section
Held at: University Engineering Department, Trumpington Street, Cambridge.
Date: 8 December.
Time: 8 p.m.
Lecture: Modern Telephone Developments with Reference to Electronic and Switching Techniques.
By: L. R. F. Harris.
(Joint meeting with the Cambridge Electronics Section of the I.E.E.).

South Wales Section
Held at: University of Cardiff, Department of Physics.
Date: 7 December.
Time: 6.30 p.m.
Lecture: The Nature of Musical Sounds and the Significance of the Transient.
By: C. A. Taylor.
(Joint meeting with the I.E.E.).

Scottish Section
Held at: Edinburgh University, Department of Natural Philosophy, Drummond Street, Edinburgh.
Date: 7 December.
Time: 7 p.m.
Lecture: Field Effect Transistors.
By: J. M. Morrison.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2
Date: 8 December.
Time: 7 p.m.
Lecture: Field Effect Transistors.
By: J. M. Morrison.

Merseyside Section
Held at: Liverpool College of Technology, Byrom Street.
Date: 14 December.
Time: 7 p.m.
Lecture: Exploring the Upper Ionosphere with Very High Power Radar.
By: C. D. Watkins.

North Western Section
Held at: R/D7 Renold Building, Manchester College of Science and Technology, Altrincham Street.
Date: 8 December.
Time: 7 p.m.
Lecture: A Four Tube Colour Camera.
By: W. T. Underhill.

North Eastern Section
Held at: The Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Date: 14 December.
Time: 6 p.m.
Lecture: A.C. Motor Control.
By: K. H. Williamson.

Thames Valley Sectioo
Held at: J. J. Thomson Physical Laboratory, University of Reading.
Date: 8 December.
Time: 7.30 p.m.
Lecture: Field Effect Transistors.
By: R. G. Bailey.

THE TELEVISION SOCIETY

Held at: The Conference Suite, ITA, 70 Brompton Road, London, S.W.3.
Date: 9 December.
Time: 7 p.m.
Lecture: Pay-TV.
By: J. Russell.

EE 01 096 *for further details*

Covers the very wide Frequency Range 0·2c/s to 1·22Mc/s

TRANSISTOR DECADE OSCILLATOR TYPE TG66A

Frequency is selected by means of four in-line additive decade controls and a five position multiplier switch. The last of the additive controls is continuously variable so that any frequency may be selected with a discrimination better than ±0.03% or 1/10th of the specified frequency accuracy, whichever is the greater.
The output source voltage of the oscillator is monitored by a meter with an expanded scale, and a continuous control is fitted so that the output may be set with a discrimination better than ±0.05dB.

SPECIFICATION

ACCURACY

±0.02c/s below 6c/s.
±0.3% from 6c/s to 100kc/s,
±1% from 100kc/s to 300kc/s,
±3% above 300kc/s,
all measured at 25°C.

STABILITY

±6% mains voltage change produces less than ±0.005% frequency change up to 100kc/s. Change of frequency with temperature is less than ±0.025% per °C above 100c/s.

DISTORTION

Less than 0.15% from 15c/s to 15kc/s
Less than 0.5% at 1.5c/s and 150kc/s

OUTPUT

Continuously variable from −94dBm. to +10dBm. into 600 ohms. Source voltage variable from 30μV to 5V. Output impedance 600 ohms at all settings. Meter fitted with dBm and V scales.

TEMPERATURE RANGE

−10°C to +45°C.

POWER SUPPLY

"Mains"—100/125V, 200/250V; 50/60c/s; 6VA, or "Batteries"—4 Self-contained type PP9, life 300 hours.

SIZE

6½" high x 10½" wide x 7" deep.

WEIGHT

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NOUVELLES Réalisations

Traduction des pages 818 à 823

Une description basée sur des renseignements fournis par les fabricants de nouveaux composants, accessoires et instruments d'essai

OSCILLOSCOPE D'ENSEIGNEMENT

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Illustration à la page 818)

La société Advance Electronics Ltd vient d'annoncer la mise au point d'un nouvel oscilloscope à l'usage des établissements d'enseignement et des marchés industriels. L'oscilloscope OS12 a, en effet, été conçu suivant la spécification Nuffield 158 et il présente des caractéristiques spéciales comportant un circuit de base de temps à auto-synchronisation se verrouillant automatiquement, un amplificateur à couplage direct et une entrée X à l'arrière de l'instrument qui permet un affichage X-Y.

Le OS12 a une sensibilité de 100 mV par division micrométrique et la vitesse de la base de temps varie de 100 msec à 100μsec par division. Il comprend un tube de 6,98 cm.

Une technique du construction très simple a été utilisée: l'oscilloscope se compose seulement de deux pièces d'aluminium préfaçonnées. Les lampes et les composants sont montés sur une seule plaquette à circuit imprimé. La société Advance Electronics Ltd pense que cet oscilloscope sera utilisé de façon très étendue dans l'industrie sous forme de contrôleur incorporé pour le matériel de contrôle de systèmes et de production.

EE 01 751 *pour plus amples renseignements*

BOÎTE À DÉCADES DE CONDENSATEURS

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devonshire

(Illustration à la page 818)

Une nouvelle boîte à décades de condensateurs (type 688/A) a été mise au point par la société Hatfield Instruments Ltd à l'usage des réalisateurs de circuits pour les travaux de tolérance et autres applications similaires. La boîte assure une sélection rapide de condensateurs dans la gamme de 100 pF à 1 μF. Elle occupe un minimum d'espace sur un banc, même lorsqu'elle est utilisée en multiples, et sa précision est supérieure à 5% à n'importe quel réglage. Quatre décades donnent des plots de 100 pF, 1000 pF, 0,01 μF et 0,1 μF. La capacitance minima lorsque tous les commutateurs sont fixés à zéro est de 30 pF et la tension nominale est de 250 V c.c. (0,1 μF par décade, 100 V c.c.). La boîte mesure: 13,97 cm × 4,12 cm × 6,98 cm.

EE 01 752 *pour plus amples renseignements*

CONNECTEURS DE CÔTÉS

Ferranti Ltd, King's Cross Road, Dundee

(Illustration à la page 818)

La société Ferranti Ltd a mis au point une nouvelle gamme de connecteurs à montage par la tranche pour les plaquettes à circuit imprimé. Ces connecteurs assurent une grande fiabilité alliée à une simplicité de conception remarquable. Ces connecteurs, qui appartiennent à la gamme EWD, peuvent être fournis avec contacts à un seul côté ou avec contacts à deux côtés.

Les modèles à contacts sur un côté utilisent des pièces d'insertion au nylon et emplis de verre des côtés non-contactants à la place des faux contacts. Ces pièces d'insertion uniques fournissent d'excellentes surfaces à faible frottement et à usage prolongé pour les plaquettes à circuit imprimé. Elles contribuent à maintenir le prix hautement compétitif des connecteurs.

Les moulages des connecteurs sont en phthalate diallyl bleu de haute qualité. Ils ont d'excellentes propriétés électriques et mécaniques et ils sont parfaitement sûrs en raison de leurs qualités d'auto-extinction. Ils peuvent résister facilement aux températures de soudage.

Le contact se distingue par son ressort à lame roulante qui a pour effet de limiter les contraintes. A l'instar de tous les autres connecteurs Ferranti, le ressort de contact à une faible caractéristique de variation, ce qui signifie que des déviations se produisant dans la gamme de fonctionnement ne donnent lieu qu'à de faibles variations de force de contact, les forces d'insertion et de retrait étant notablement réduites. Les tolérances d'épaisseur de plaquette normales sont facilement admises. Les forces de contact contrôlées s'obtiennent dans une mesure suffisante pour pouvoir briser des films de ternissure et assurer des résistances de contact faibles et stables quelle que soit la faiblesse de la tension de régime.

Les connecteurs de la nouvelle gamme EWD sont fournis avec des chevilles de soudage dans des positions à 8, 16, 24, 32 ou 40 pôles pour les contacts à un côté et à 8/8, 16/16, 24/24, 32/32 ou 40/40 pôles pour les contacts à deux côtés. Ils ont, en outre, été étudiés pour recevoir des plaquettes à circuit imprimé d'une épaisseur normale de 1,5 mm. L'espacement entre pôles est de 3,8 mm. La pénétration des plaquettes est de 9,40 mm et l'introduction de la plaquette dans le connecteur est facilitée par l'angle généreux d'entrée.

Les contacts sont en cuivre au béryllium et ils sont dorés sur une épaisseur de 2,5 microns, donnant ainsi une surface durable et non-poreuse.

EE 01 753 *pour plus amples renseignements*

DISPOSITIF DE REPORT À TROIS ENTRÉES

Electronics Associates Ltd, Burgess Hill, Sussex

(Illustration à la page 818)

Le "Variplotter", serie 1131, est un appareil conçu pour les applications exigeant le report simultané de deux variables par rapport à une troisième.

Trois systèmes indépendants à entraînement asservi, une sensibilité de base de 1 mV/2,5 cm, 18 gammes étalonnées et des potentiomètres à variations sans plots assurent à l'appareil une très grande souplesse.

La base de temps incorporée ajoute à l'appareil une possibilité de report t-y qui permet des balayages sur la totalité de l'échelle dans six gammes étalonnées de 0,5 sec à 20 sec/2,5 cm. Des commandes séparées de base de temps et de levage de plume assurent des courses sèches avant l'enregistrement.

Le Variplotter 1131, qui est muni de circuits constitués de corps solides et d'un système de référence à diodes Zéner, comprend des cartouches à encre à fiches et ne coulant pas, des commutateurs de contrôle à lumière arrière, des filtres interchangeables pour la fréquence de ligne et la suppression de bruit et un serre-papier du type à ventilateur.

La spécification est comme suit: précision statique ±0,1%, précision dynamique ±0,2% à 42,5 cm/sec; répétabilité ±0,5%. La vitesse de pivotement est de 42,5 cm/sec pour le bras et de 50 cm/sec pour les plumes. La résistance d'entrée est de 2,5 MΩ/V dans des gammes allant jusqu'à 20 mV/2,5 cm, et elle est constante à 1 MΩ aux gammes supérieures. Livrable avec graduation métrique ou anglaise, l'appareil peut être monté sur bâti ou sur pont. Les dimensions hors-tout sont: 48,26 cm × 53,34 cm de hauteur × 16,51 cm de profondeur.

EE 01 754 *pour plus amples renseignements*

RELAIS À RETARD À TEMPS STATIQUE

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustration à la page 819)

La société Solid State Controls Ltd fabrique actuellement un relais à retard de temps transistorisé et à fiches d'une grande fiabilité et d'un prix économique, spécialement conçu pour répondre aux besoins de l'industrie moderne.

Le TDR/407 est un dispositif robuste et entièrement encapsulé; il n'a aucune partie mobile sujette à l'usure ou aux pannes. Il est insensible aux vibrations, aux chocs et aux tensions transitoires et il peut effectuer un minimum de 100 millions d'opérations continues. Cet élément compact est entièrement à l'épreuve de la dérive, même après plusieurs milliers d'heures de service, et il n'exige aucune espèce d'entretien.

Le TDR/407 est livrable en trois gammes de tension: 12-13, 35-70 et 70-150 V. Il fonctionne avec précision même lorsqu'il est soumis à d'importantes variations dans la tension d'utilisation et dans la température ambiante. Le circuit de minutage principal se compose d'un circuit de retard à faible impédance et complètement stable muni d'un commutateur de sortie à redresseur de commande au silicium. Le contact semiconducteur peut brancher des charges inductives ou résistives allant jusq'à 1 A à 150 V et il peut recevoir d'importants courants d'entrée qui sont normalement propres aux circuits de commutation de moteurs ou de contacteurs.

Les périodes de retard de temps en quatre gammes de 0,2 sec à 11 min peuvent être assurées et les réglages des temps de retard peuvent être effectués par la commande montée sur le coffret ou préréglés. La précision du minutage lorsque le relais est soumis à des extrêmes de température ambiante et de tension d'utilisation est typiquement de 5%, et la précision de répétion typique est de 1%.

La grande fiabilité et la constitution de corps solides du composant le rendent particulièrement approprié aux utilisations de l'aviation ainsi qu'aux usage industiels. Parmi les nombreuses applications qui ont déjà été considérées pour ce relais, il faut cites les systèmes de commande de circulation et les processus de fabrication, les circuits de contrôle de moteurs, les commandes de levage, les centrales de puissance, le matériel de contrôle téléphonique, la commutation de canaux de télémétrie au sol et dans l'air, la commande de générateurs de réserve et les machines de conditionnement automatique.

EE 01 755 *pour plus amples renseignements*

INSTRUMENT DE MESURE SENSIBLE AUX IMPULSIONS SÉLECTRICES À REDRESSEUR PILOTÉ AU SILICIUM

Caltronics Ltd, Huntingate, Hitchin, Hertfordshire

(Illustration à la page 819)

Cet instrument de mesure sensible aux impulsions sélectrices à redresseur piloté au silicium fournit une mesure rapide et précise du déclenchement périodique courant/déclenchement pour une variété de redresseurs pilotés au silicium. Une diode zéner fournit une tension stabilisée anode/cathode de 6 V au redresseur piloté au silicium soumis à l'essai.

Les caractéristiques de déclenchement périodique du redresseur piloté au silicium soumis à l'essai sont mesurées en appliquant des impulsions redressées d'une demi-onde de 50 à 60 Hz entre le conducteur de porte et la cathode du dispositif. Le réseau de déclenchement dans le câble d'alimentation de l'anode est sensible au pivotement du redresseur piloté au silicium soumis à l'essai et amorce un commutateur électronique qui supprime le signal de déclenchement périodique. Un circuit à voltmètre à lecture de pointe est utilisé pour fournir une indication directe du courant ou du niveau de tension auquel le redresseur piloté au silicium est déclenché. La lecture périodique du courant/déclenchement est obtenue en entraînant le circuit de déclenchement à travers une série de résistances de précision qui forment une source de courant réglable de 10 degrés. La lecture de porte tension-déclenchement est obtenue en entraînant le circuit à déclenchement périodique à partir d'une source de tension réglable à gradins.

Les potentiomètres d'étalonnage permettent d'étalonner l'instrument de mesure de sensibilité tant pour le courant que pour la tension.

Des bornes sont prévues pour l'utilisation à distance en parallèle de l'instrument. L'indication à distance de la lecture peut être obtenue au moyen d'une paire de bornes qui fournissent un signal de +1 V pour l'indication sur la totalité de l'échelle de l'instrument de mesure sur le panneau frontal.

EE 01 756 *pour plus amples renseignements*

CONVERTISSEUR ANALOGIQUE/NUMÉRIQUE

Ether Engineering Ltd, Park Avenue, Bushey, Hertfordshire

(Illustration à la page 819)

Le convertisseur modèle MA 1102 transforme des tensions d'entrée analogique de 0 à −10 V en une sortie codée binaire pure équivalente de 10 chiffres binaires.

L'instrument utilise la méthode d'approximation successive pour fournir des données précises d'ordinateur à coût réduit. Une partie importante du circuit utilise des modules doubles NI à pellicule mince. Ces modules sont branchés, suivant les exigences du circuit, comme circuits à déclenchement périodique, éléments bistables et monostables ou comme inverseurs.

La construction mécanique compacte du convertisseur permet de l'incorporer à des systèmes existants, car il n'occupe que 4,44 cm d'espace de panneau.

L'entrée analogique totale est de 10 V, correspondant à 2^{10} chiffres binaires. La résolution est de ±1 chiffre et la précision est de 0,1 %. La durée de conversion est de 200 µsec.

EE 01 757 *pour plus amples renseignements*

SYSTÈME RÉDUCTEUR DE BRUIT

Dolby Laboratories, 590 Wandsworth Road, London, S.W.8

(Illustration à la page 820)

Le Tendeur Signal/Bruit des Laboratoires Dolby permet de réduire le bruit dans les systèmes d'enregistrement et d'émission basse fréquence de qualité. Le Tendeur Signal/Bruit peut être utilisé dans n'importe quelle circonstance où le signal est obtenu pour le traitement à l'entrée et à la sortie de la chaîne acoustique. Tandis que le signal lui-même émerge sans modification de forme, l'action de réduction du bruit atténue les types habituels de bruits que l'on enregistre dans le doublage des films, l'enregistrement sur bande vidéo, etc.

En raison de l'effet bien connu de masquage dans l'ouie humaine, les sons faibles sont masqués en partie ou même complètement par des sons plus élevés, particulièrement lorsque la différence de fréquence n'est pas trop grande. Ainsi les rapports signal/bruit d'enregistrement sur bande magnétique sont supérieurs à ceux qui seraient indiqués par l'analyse de la forme d'onde de réenregistrement sur bande magnétique; le bruit de la bande magnétique dépend de l'amplitude instantanée du signal et augmente en fonction de cette amplitude, effet qui porte le nom de bruit de modulation. Le fait que le mécanisme d'audition supprime effectivement ce bruit supplémentaire dans les conditions de signal normales semblent indiquer qu'un système électronique de conception appropriée pourrait utiliser le phénomène de masquage pour obtenir une réduction supplémentaire du bruit.

Utilisant le principe de masquage ainsi que la compression et l'expansion du signal, le tendeur signal/bruit effectue une réduction de bruit d'une part en élevant le composant de signaux à faible niveau durant l'enregistrement à n'importe quel moment possible (compression), suivi d'une atténuation complémentaire durant le réenregistrement (expansion) et d'autre part en masquant l'effet lorsque le niveau de signal est tellement élevé que la compression et l'expansion ne sont plus possibles.

Etant donné que le masquage est moins effectif avec des fréquences de bruit quelque peu éloignées de la fréquence de signal il est nécessaire de traiter les différentes parties du spectre indépendamment. Le système de réduction de bruit fournit alors un niveau de bruit plus bas et apparemment constant, le "chuchotement" classique ou le bruissement de la compression normale et de l'expansion étant absents. Le tendeur signal/bruit divise le spectre acoustique en quatre bandes et comprime et détend chacune de ces bandes de manière essentiellement indépendante. Des bandes séparées sont prévues pour la gamme de fréquence du bourdonnement, pour la gamme acoustique médiane, pour les fréquences mi-hautes et pour les fréquences élevées. Un signal de niveau élevé dans une bande ne peut donc empêcher la réduction du bruit dans une autre bande où le niveau de signal est faible.

D'un autre point de vue, le système produit effectivement une caractéristique d'égalisation de l'enregistrement qui se conforme de manière continue au signal d'entrée de telle sorte qu'il puisse produire un rapport optimum signal/bruit durant le réenregistrement.

EE 01 758 *pour plus amples renseignements*

SOURCES DE SIGNAUX ACOUSTIQUES

A. P. Besson & Partner Ltd, St. Josephs Close, Hove, Sussex

(Illustration à la page 820)

Le "Bleeptone" a été mis au point pour fournir une source de signaux acoustiques de construction compacte et d'une grande fiabilité. Il comprend un circuit oscillateur autonome et un écouteur à induit oscillant conçu pour fonctionner à partir d'une batterie à faible tension. Par ses dimensions et son aspect, il est identique au récepteur à induit oscillant bien connu utilisé dans les instruments téléphoniques modernes. L'émetteur de notes peut normalement être entendu à des distances allant jusqu'à 305 m et la fréquence est telle que des bruits de fond n'ont aucun effet appréciable de masquage.

Le circuit d'entraînement est semblable à celui utilisé pour les diapasons à entretien automatique. Deux bobines séparées sont disposées dans le circuit magnétique, l'une pour entraîner l'induit oscillant et l'autre pour capter le signal induit et fournir une réaction à la base du transistor qui, avec les composants connexes, forme le circuit amplificateur complet.

La fréquence fondamentale de fonctionnement est déterminée en principe par les caractéristiques mécaniques de l'induit oscillant.

Le dispositif est prévu pour des utilisations dans une gamme étendue d'industries. Il est particulièrement indiqué pour donner un signal d'avertissement, notament dans les cas où le degré de fiabilité exigé doit être supérieur à celui pouvant être obtenu avec un vibreur ordinaire.

EE 01 759 *pour plus amples renseignements*

AMPLIFICATEUR DE RADIOTÉLÉPHONE

The Plessey Co. Ltd, Ilford, Essex

(Illustration à la page 820)

Un nouvel amplificateur transistorisé pour le matériel radiotéléphonique a été mis au point à l'usine de Bridgnorth du groupe d'électronique Plessey.

Conçus à l'origine pour les radiotéléphones, types 700 et 700R, de la société Plessey, ces amplificateurs dont le gain de 14 dB peut augmenter la sortie de puissance au-delà de 8 W peuvent être également incorporés à des faisceaux radiotéléphoniques existants. Il ne mesurent que 7,62 cm de hauteur, 15,87 cm de largeur et 7,62 cm de profondeur.

En plus du modèle de base, il existe également une version en coffret, hermétiquement scellée et entièrement tropicalisée pour usage extérieur dans n'importe quelle partie du monde.

Convenant pour l'utilisation sur courant continue de 12 V, ce matériel peut également être prévu pour le fonctionnement sur des fréquences de 68 à 74 MHz.

La consommation électrique d'environ 1,6 A permet d'ajouter le nouvel amplificateur à des systèmes mobiles ou de l'employer comme source de puissance pour entraîner des multiplicateurs à varactor jusqu'aux régions du gigacycle par seconde.

EE 01 760 *pour plus amples renseignements*

BLOCS D'ALIMENTATION

Distributeurs: Scientific Measurements & Equipment Ltd, 78 Main Street, Queniborough, Leicester

(Illustration à la page 820)

Ces blocs d'alimentation construits par la société Knott Electronic de Munich, ont été conçus pour l'utilisation en liaison avec le matériel électronique exigeant des tensions continues extrêmement constantes et précises telles que les tubes G.M., les compteurs proportionnels, les chambres d'ionisation, les intensificateurs d'image, les multiplicateurs photoélectriques, les compteurs à scintillations, les tubes cathodiques, etc.

Le courant de sortie élevé et la résistance interne très faible permettent l'alimentation simultanée de plusieurs éléments sans interaction. Ces blocs se caratérisent en particulier par leur rapport élevé de stabilisation, leur stabilité à long terme, le réglage de fusibles et l'ajustabilité à 1 volt près. Ils sont fournis en coffret ou pour montage sur bâti.

La gamme de blocs livrables couvre des utilisations de 0 à 30 V à 10 A, et de 0 à 15 KV à 10 mA. Un élément à courant stabilisé fournissant de 0 à 10 A à 13 V est également fourni.

EE 01 761 *pour plus amples renseignements*

MILLIVOLTMÈTRE C.A.-C.C.

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(Illustration à la page 820)

Le Millivoltmètre TMI est un appareil entièrement transistorisé comportant douze gammes pour tension alternative et pour tension continue, allant de 1 mV à 300 V, de déviation sur la totalité de l'échelle en une séquence de 1-3-10; une échelle de décibels est également prévue allant de −10 à +2 dB (0 dB = 1 mW dans 600Ω).

Sur tension continue la précision est de 3 % de la totalité de l'échelle ±100 µV et sur tension alternative elle est de 4 % d'une déviation sur la totalité de l'échelle ±100 µV de 20 Hz à 100 kHz. La résistance d'entrée de courant continu est de 1 MΩ/V jusqu'à 10 V et elle est constante à 10 MΩ sur les gammes supérieures. L'impédance d'entrée de courant alternatif est de 100 kΩ de 1 mV à 30 mV, 1 MΩ de 100 mV à 300 mV et de 10 M de 1 V à 300 V; la capacité de shunt maxima est de 40 pF.

Pour la tension continue, le millivoltmètre a un atténuateur d'entrée à résistance, suivi d'un amplificateur avec instrument de mesure et commande de réaction. Dans la section de tension alternative, un atténuateur à compensation de fréquence alimente un amplificateur de tension continue à tension alternative qui comporte un circuit à réaction et un instrument de mesure. De plus, dans les quatre gammes inférieures, un amplificateur convertisseur d'impédance maintient l'impédance d'entrée à 100 kΩ.

Deux batteries PP11 sont utilisées par l'appareil qui comprend, en outre, un dispositif de contrôle de la tension de batterie. Le poids de l'appareil sans les batteries est de 2,86 kg et il mesure 165 mm de hauteur × 218 mm de large × 220 mm de profondeur.

EE 01 762 *pour plus amples renseignements*

INDICATEUR DE RAPPORT D'AMPLITUDE DE TENSION D'ONDES STATIONNAIRES

Marconi Instruments Ltd, Sanders Division, Gunnels Wood Road, Stevenage, Hertfordshire

(Illustration à la page 821)

Un nouvel indicateur de rapport d'amplitude de tension, type 6596, à faible bruit et à amplificateur sélectif et à coût réduit a été réalisé par la division Sanders de la société Marconi Instruments Ltd. Sa sensibilité et sa gamme de gain le rendent approprié pour la plupart des mesures de micro-ondes utilisant des détecteurs piézoélectriques et à modulation du signal haute fréquence à 1 kHz.

Des commandes de réglage précis et approximatif sont fournies et elles comportent des variations suffisantes pour contrôler des signaux détectés allant jusqu'à 1 mV efficace.

L'échelle de l'instrument de mesure est étalonnée en rapport d'amplitude de tension d'ondes stationnaires, basée sur l'hypothèse que le détecteur fonctionne dans la région des ondes carrées de ces caractéristiques, c'est à dire une tension de sortie proportionnelle à la puissance en microondes.

La sensibilité maxima est supérieure à 1 μV de déviation sur la totalité de l'échelle et le bruit est inférieur à 01 μV; un sélecteur d'entrée permet la sélection individuelle de l'un ou de l'autre des deux signaux d'entrée. En outre, la différence des deux peut être sélectionnée pour les méthodes à pont de microondes extra-sensibles.

L'appareil est idéalement conçu pour l'utilisation en liaison avec le banc d'essai pédagogique Sanders type 599.

EE 01 763 *pour plus amples renseignements*

CONDENSATEUR EN CÉRAMIQUE MINIATURE

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

(Illustration à la page 821)

La société Mullard vient d'annoncer la réalisation d'une nouvelle série de condensateurs en céramique miniature de haute qualité, de forme rectangulaire et ne mesurant que 1,9 mm d'épaisseur pour permettre des densités d'entassement élevé sur plaquettes de circuit imprimé utilisant un réseau de 2,5 mm. Ces condensateurs peuvent être utilisés tant dans le matériel domestique que dans le matériel industriel. Ils se composent d'une plaque mince de matériau céramique métallisé et isolé à l'aide d'une couche protectrique de laque qui leur assure une performance excellente dans les conditions d'humidité les plus élevées.

Les condensateurs de la nouvelle série C333 mesurent 5 mm × 8,5 mm × 1,9 mm (à l'exclusion des câbles de 13,5 mm). La gamme de capacité s'étend de 3,9 pF à 150 pF avec une tolérance de 0,5 pF ou ±2 % selon celles de ces deux valeurs qui est la plus élevée. La tension de régime est de 40 V dans une gamme de température de −25° C à +85° C et la résistance à l'isolement mesurée à 10 V est supérieure à 1000 MΩ. Le coefficient de température varie entre 0 et 750 parties négatives par million, suivant la valeur de capacité.

La tolérance réduite et la stabilité élevée des nouveaux condensateurs les rend particulièrement indiqués pour l'emploi dans les transformateurs moyenne fréquence de télévision, les circuits accordés et les autres applications exigeant une haute performance et une perte réduite.

EE 01 764 *pour plus amples renseignements*

PONT DE PRÉCISION

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 821)

Le plus récent modèle de la gamme Wayne Kerr de pont à équilibrage automatique, à savoir le modèle B331, allie la haute précision exigée par les laboratoires à la vitesse et à la simplicité de fonctionnement assurées par l'annulation électronique. Deux instruments de mesure fournissent des indications simultanées des termes de quadrature et en phase de n'importe quel composant ou impédance complexe. Des boutons poussoirs d'affichage en ligne éclairé permettent de contrebalancer les lectures par trois décades donnant une résolution maxima de six chiffres sur toutes les gammes.

Des circuits spéciaux assurent la compensation automatique de l'impédance des conducteurs de mesure qui aboutissent à une pince Kelvin de type perfectionné. Les sorties permettent d'actionner des voltmètres numériques, des imprimeurs, des enregistreurs, des mécanismes passe/rejet ou des circuits de contrôle. Les prises pour étalon externe permettent les mesures comparatives et lorsqu'elles sont utilisées avec les commandes Vernier, elles peuvent réaliser une discrimination de 10 parties par million.

La source interne et le détecteur fonctionnent à 1,00 kHz mais ils peuvent fonctionner entre 50 Hz et 20 kHz en utilisant un équipement externe. La capacité de mesure totale à 1 kHz est de 0,001 pF à 0,25 F, 1 pΩ à 1 kΩ, 1 mΩ à 1 TΩ et de 100 nH à 250 MH. La précision totale de 0,01% s'applique de 1 pF à 10 μF et de 10 nΩ à 100 mΩ. Le pont B331 mesure 48,26 cm de large × 30,48 cm de haut × 22,86 cm de profondeur et son poids est d'environ 23 kg.

EE 01 765 *pour plus amples renseignements*

RELAIS À LAME VIBRANTE

Hendrey Relays & Electrical Equipment Ltd, 390-394 Bath Road, Slough, Buckinghamshire

(Illustration à la page 821)

Conçu conformément aux fonctions de la norme britannique B.S. 2G. 100, ce relais à lame vibrante type 5858 peut être utilisé au sol ou en l'air, avec des valeurs-limite de 5 Hz à 2 kHz à 10 g et une accélération de 50 g appliquée dans n'importe quelle direction. La bobine d'excitation et les éléments à lame vibrante sont encapsulés dans un coffret en acier qui assure une protection mécanique aussi bien que magnétique. Le matériau d'encapsulation est en résine d'époxyde ayant d'excellentes caractéristiques électriques et une bonne performance aux variations de température. La résine forme également la base du relais portant les broches de connexion, format ainsi un élément très robuste, insensible à l'humidité et pouvant être utilisé dans les conditions d'environnement les moins favorables et dans des températures ambiantes de −60° à +120°C.

Des lames vibrantes soigneusement choisies et le formage avant l'utilisation garantissent un minimum de 2 × 10^6 opérations sous pleine charge et le temps de réponse des contacts est supérieur à 2 msec, avec un rebondissement négligeable tant à la mise en circuit qu'à la coupure.

La valeur nominale maxima est de 0,5 A ou 30 Vc.c. ou 10 W de charge résistive. Les charges inductives doivent être éteintes de façon appropriée. Trois dispositifs de contact standard sont prévus: (a) deux normalements ouverts; (b) quatre normalement ouverts; (c) deux normalement ouverts et deux normalement fermés. Le bobinage des bobines peut être prévu pour n'importe quelles tensions continues jusqu'à 100 V.

Le relais est fourni avec des terminaisons soudées ou avec des conducteurs mobiles, montage sur support ou par bande, ou, enfin, avec des broches pour bases à fiches.

EE 01 766 *pour plus amples renseignements*

COMMUTATEUR À RETARD

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Illustration à la page 822)

Le Radian est un composant à bas prix et de haute qualité qui remplace avantageusement les commutateurs à retard thermique, les minuteries à relais/condensateur simples et les moteurs synchrones dans les conditions exigeant une précision de répétition élevée, une stabilité à long terme et un temps de rétablissement rapide. Le relais Radian n'exige aucune période de refroidissement contrairement aux éléments thermiques car il se rétablit instantanément et donne des opérations de minutage d'une grande uniformité. Les tensions de fonctionnement pouvant être obtenues sont de 24 Vc.c. et 24 Vc.a. tandis que la sortie met an action une gamme étendue de relais et de contacteurs.

Le radian est un élément à fiches qui assure deux types d'opérations: le retard à l'excitation et le retard au désamorcement avec deux gammes de minutage, une à 60 sec et une à 5 min.

L'élément est insensible aux phénomènes transistoires sur la ligne d'alimentation, aux fluctuations dans la tension de régime de ±15 % et aux variations de température ambiante de −10° C à +55° C.

EE 01 767 *pour plus amples renseignements*

BLOC D'ALIMENTATION EN TENSION CONSTANTE

Servomex Controls Ltd, Crowborough, Sussex

(Illustration à la page 822)

La société Servomex Controls Ltd a mis au point deux nouvelles versions de ses stabilisateurs de tension type AC2 et AC7, en vue d'applications n'exigeant pas le système de montage sur bâti. Ces nouveaux types, modèle AC2 Mk 111A et AC7 (Industriel), sont plus faciles à installer et entretenir ayant été conçus pour l'installation par le personnel électrique des fabriques par opposition au personnel électronique. Toutes les parties qui exigent un entretien périodique sont facilement accessibles en retirant le couvercle sans qu'il soit nécessaire de déplacer l'instrument pour avoir accès à l'arrière, comme c'est habituellement le cas pour les modèles à montage sur bâti.

Les composants électriques individuels sont identiques à ceux des modèles pour montage sur bâti. Cela s'applique également à la qualité et au fini de la construction de ces nouveaux types. Le modèle AC2 Mk 111A est plus petit et plus maniable que le type pour montage sur bâti, tandis que le modèle AC7 (Industriel) est monté sur roue et à l'épreuve des poussières et de l'eau. Il est donc idéal pour l'installation dans une cave ou dans un passage fréquemment balayé.

Les deux instruments fonctionnent sur alimentation d'entrée dans la gamme de 200 à 250 V et fournissent une tension de sortie à haute stabilité ne variant pas de plus de ±0,1% de la tension nominale de l'alimentation. Ils sont insensibles aux charges de redresseurs ou au changement dans le facteur de puissance, c'est à dire du facteur résistif au facteur purement réactif, y compris les charges capacitives.

La mise en oeuvre se fait au moyen d'un transformateur variable utilisé pour fournir une tension de biberonnage ou d'appoint, suivant les nécessités, pour corriger l'entrée à la valeur requise. Ce transformateur variable est entraîné par un moteur asservi biphasé avec engrenages enfermés et dispositif de limitation du couple. Les variations de la tension de sortie sont détectées par un pont bolométrique dont la sortie est amplifiée dans un amplificateur asservi à deux valves et appliqué au moteur.

Les instruments sont de construction exceptionnellement robuste et peuvent résister à des impacts d'un maximum de 40 g dans n'importe quelle direction. Ils ne comportent ni relais, ni thyratrons, ni condensateurs électrolytiques.

Les nouveaux types sont moins coûteux, non seulement en raison des coffrets plus simples utilisés mais également parce qu'ils ne comprennent pas de voltmètre, d'ampèremètre, de commutateur arrêt/marche et de fusible principal. Ces deux derniers composants sont censés être montés sur le mur avec un voltmètre externe comme composant facultatif.

EE 01 768 *pour plus amples renseignements*

INDICATEUR DE CONTRAINTE NUMÉRIQUE

S.E. Laboratories (Engineering) Ltd, Astronant House, Feltham, Middlesex

(Illustration à la page 822)

L'indicateur de contrainte numérique, type 601, est un instrument portatif pour mesurer la contrainte dynamique et statique comportant de nombreuses innovations. La lecture numérique facilite considérablement le travail de l'opérateur et, grâce à son commutateur "d'addition," elle couvre une gamme totale de ±50 000 microcontraintes avec une précision de ±0,1 pour cent de la lecture ou ±5 μE, suivant celles de ces deux valeurs qui est la plus élevée (G.F. = 2,0).

Les cellules rechargeables fixées à l'intérieur permettent d'utiliser l'appareil pendant de longues périodes. Enfin, la commande du "facteur de mesure" peut être utilisée pour obtenir une indication directe à partir de dispositifs tels que les cellules de charge, les transducteurs de pression et de déplacement, etc.

L'excitation par ondes carrées permet l'utilisation de l'appareil avec des circuits extensométriques ayant une capacitance maxima de 0,01 μf sans perte appréciable de précision, cependant que l'excitation sinusoïdale peut être assurée si nécessaire.

L'appareil peut être utilisé avec des extensomètres ayant une résistance de 50 Ω à 2 kΩ. Des sorties de 1 mA et 2 V peuvent être obtenues à partir de douilles séparées galvanométriques et oscilloscopiques pour entraîner des dispositifs à lecture conventionnelle ou des instruments d'enregistrement.

EE 01 769 *pour plus amples renseignements*

SOLÉNOÏDES À MAINTIEN AUTOMATIQUE

H. E. & B. S. Benson Ltd, Exning Road, Newmarket, Suffolk

(Illustration à la page 822)

La société H. E. et B. S. Benson Ltd vient d'annoncer la réalisation d'une solénoïde indépendante entièrement nouvelle. Le plongeur est en matériau magnétique permanent et lorsqu'il est fermé pendant une impulsion de courte durée il reste fermé pendant une période de temps illimitée et peut soutenir une charge sans tirer du courant. Une impulsion inverse déclenche le plongeur.

Toute forme de tension continue ou tension alternative redressée (y compris une batterie sèche) peut mettre en oeuvre cette solénoïde. Elle comporte un circuit de courant alternatif qui ne nécessite qu'une seule diode et une petite résistance pouvant être soudée directement au commutateur de fonctionnement.

Les avantages de cette solénoïde comprennent la montée de température nulle pour une fiabilité maxima (utile également dans une température ambiante élevée); une gamme étendue de caractéristiques de force toutes avec maintien continu depuis la commutation momentanée en dépit de toute panne de courant.

Les solénoïdes à maintien automatique PMB sont identiques extérieurement aux solénoïdes existantes B.2, 3, 4 et 5. Elles peuvent remplacer les solénoïdes de verrouillage et lorsqu'elles sont munies d'un ressort de retour elles peuvent fournir une action positive à deux positions d'un élément à double action et à deux bobines plus coûteux.

EE 01 770 *pour plus amples renseignements*

HORLOGE NUMÉRIQUE

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

(Illustration à la page 823)

La société Venner Electronics Ltd a mis au point une horloge numérique entièrement constituée de composants au silicium, type TSA 6686, utilisant des plaquettes de circuit imprimé en fibre de verre et assurant un affichage en heures, minutes et secondes provenant de six tubes indicateurs numériques à gaz. Deux versions sont prévues, à savoir une version donnant l'heure sur une base de 24 heures et une autre donnant l'heure sur une base de douze heures. Des sorties électriques permettent d'entraîner un imprimeur ou un élément à bande perforée.

L'arrêt ou la marche sont mis en oeuvre soit par des impulsions négatives de 4 V d'amplitude, soit par boutons-poussoirs montés sur le panneau frontal. L'affichage peut être réglé à la main pour l'indication indépendante des heures, des minutes et des secondes et comporte un réenclenchement unique à zéro.

Conçue pour fonctionner dans une gamme de températures ambiantes de −5° C à +60° C, la précision de l'horloge est de deux parties dans 10^6.

L'alimentation est de 120 à 125 V ou de 200 à 250 V, 50 Hz c.a. les dimensions de l'horloge TSA 6686 sont de 13,33 cm de hauteur, 24,13 de profondeur et 48,26 cm pour montage sur bâti.

EE 01 771 *pour plus amples renseignements*

TUBE À ONDES PROGRESSIVES

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Illustration à la page 823)

La société M-O Valve Co. Ltd a mis au point un nouvel amplificateur à ondes progressives par impulsions de grande puissance, le type TWX16, pour les applications de radar. Il donne un gain supérieur à 40 dB pour une puissance de sortie de pointe de 5 kW (30 W de puissance moyenne et 5 à 20 kW de puissance de pointe saturée) dans une largeur de bande de 500 MHz dans la gamme de 8 à 9,3 GHz.

Le tube est de construction en céramique/métal et utilise une structure à ondes lentes à anneaux et barres qui supprime les oscillations indésirables. La focalisation s'obtient en mettant en oeuvre dans un assemblage à montage de solénoïdes, le type SMX16, qui comprend les couplages HF et assure le refroidissement du tube.

Le TWX16 a été étudié pour fonctionner avec l'hélice maintenue à un potentiel de courant continu pour simplifier l'alimentation de la puissance des impulsions.

Le tube à ondes progressives de 1 watt, type TWX8 peut être utilisé comme élément d'entraînement.

EE 01 772 *pour plus amples renseignements*

BLOC D'ALIMENTATION À KLYSTRON

Distributeurs: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(*Illustration à la page 823*)

L'appareil suédois Oltronix, LS 525R, est un bloc d'alimentation en tension continue très stable, conçu pour la mise en action de klystrons "réflexe." Les tensions de faisceaux, de réflecteurs, de grilles et de chauffe sont contrôlées séparément. La tension de réflecteur peut ne pas être modulée pour le fonctionnement sur ondes progressives du klystron où elle peut être modulée, soit avec une onde carrée produite intérieurement, une dent de scie d'impulsion ou une onde sinusoïdale ou enfin avec un signal externe. La tension de grille peut être modulée intérieurement avec une onde ou une impulsion carrée. Les tensions d'onde carrée ou de modulation d'impulsion sont fixées au niveau des ondes progressives de la grille ou du réflecteur.

La tension de faisceau varie de −200 V à −3600 V de 0 à 125 mA. Une variation de tension de ligne de 10 % provoque un changement inférieur à 300 mV de la sortie tandis qu'un changement de charge nulle à charge complète provoque une variation inférieure à 200 mV.

La sortie est à variation continue en quatre gammes: 200 à 1200, 1000 à 2000, 1800 à 2800 et 2600 à 3600 et elle est entièrement protégée contre les surcharges et les court-circuits par un fusible électronique réglable de 10 à 125 mA.

Les nouveaux équipements peuvent également être utilisés pour les applications de multiplicateurs photoélectriques, car il est possible d'utiliser 100 multiplicateurs photoélectriques à partir de la même alimentation.

On voit sur le haut de notre gravure le LS 525R et sur le bas un élément supplémentaire pour entraîner les klystrons.

EE 01 773 *pour plus amples renseignements*

MILLIVOLTMÈTRE À LARGE BANDE

Distributeurs: Livingston Laboratories Ltd, Livingston House, Greycaines Road, North Watford, Hertfordshire

(*Illustration à la page 823*)

La société Radiometer de Copenhague vient de mettre au point un "successeur" à son voltmètre électronique RV 33. Il s'agit du nouveau millivoltmètre à large bande, type RV 35, qui assure la protection contre les surcharges jusqu'à 500 Vc.a. ou c.c., et mesure des tensions de 10 V à 300 V. La précision est de 2 % sur la majeure partie de la gamme de fréquence qui s'étend de 10 Hz à 6 MHz.

Le RV 35 peut en outre être utilisé comme amplificateur à gain élevé avec une tension de sortie sur la totalité de l'échelle d'environ 80 mV à travers 75 Ω.

EE 01 774 *pour plus amples renseignements*

TUBE CATHODIQUE EN CÉRAMIQUE

Ferranti Ltd, Glen Mill, Oldham, Lancashire

(*Illustration à la page 823*)

Le département d'électronique de la société Ferranti Ltd a réalisé ce qu'on peut considérer comme le premier tube cathodique en céramique avec focalisation électrostatique complète. Conçu pour résister aux vibrations d'environnement les plus fortes, le nouveau tube allie une très haute résolution à un diamètre d'écran utile de 12,7 mm.

Le nouveau tube a une longueur totale de 114,3 mm et un diamètre de corps de 15,2 mm. La plaque de la face a un diamètre d'écran utile de 12,7 mm et elle est faite en verre de haute qualité spécifiquement mis au point pour l'emploi avec l'enveloppe en céramique à laquelle elle est scellée au moyen de techniques spéciales. La première tension d'anode typique est de 300 V et la tension d'anode finale est de 8 kV. La brillance de ligne mesurée avec la première anode à 450 V et l'anode finale à 7 kV est de l'ordre de 600 pieds lambert. Le tube complet avec bobine de balayage et écran en mu-métal pèse moins de 110 gm.

Les composants du canon et le système de focalisation sont usinés suivant des tolérances très serrées ce qui permet d'obtenir une très bonne définition et une résolution supérieure à 500 lignes. Ces qualités permettent de présenter le même volume d'information sur l'écran que celui qui serait normalement affiché par un tube cathodique ayant un diamètre d'écran beaucoup plus grand.

EE 01 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Un système de contrôle pour une jauge à vide moléculaire par R. J. Christian

Résumé de l'article aux pages 772 à 777

Cet article traite d'un appareil de contrôle électronique utilisé en liaison avec une jauge à vide moléculaire à palette vibrante. L'auteur examine les conditions de contrôle par jauge, les dispositifs de perception et d'entraînement étant sous la forme d'un condensateur différentiel.

La méthode utilisée est celle qui consiste à mesurer les changements de capacitance durant les vibrations de la palette au moyen d'un système de modulation de fréquence qui fournit une tension en fonction de l'amplitude de vibration. La réaction de cette tension à travers des circuits appropriés fournit l'entraînement de la jauge. L'amplitude des oscillations est stabilisée au moyen d'un circuit de commande, la tension de contrôle étant une fonction logarithmique de la pression. Cette tension est utilisée pour fournir une échelle de pression logarithmique, laquelle est indiquée par un voltmètre électronique qui assure en outre d'autres fonctions. L'accroîssement rapide de l'amplitude de la palette est effectué au moyen d'un déclencheur de mise en marche. Les résultats expérimentaux indiqués ne s'appliquent qu'a l'air bien que la jauge ait été utilisée avec succès pour le hélium (masse 4) et l'hexafluoropropylène (masse 150) ainsi que pour des gaz ayant des masses moléculaires intermédiaires.

Mesure de la température de cryostat de 0,1°K à 20°K à l'aide d'un oscillateur à pont de Wien par P. R. Adby

Résumé de l'article aux pages 778 à 781

Les oscillateurs dont il est question dans cet article sont une modification du type à pont de Wien conventionnel dont la puissance dissipée dans les résistances du réseau à déphasage est très réduite. L'une des résistances dépend de la température et peut être placée dans un cryostat à des températures minima de 0,1°K sans affecter de manière sensible les conditions thermiques. La fréquence d'oscillation est fonction de la température et elle peut être mesurée au moyen d'un compteur.

Le contrôle de mode électronique d'amplificateurs opérationnels par A. D. Bond et P. L. Neely

Résumé de l'article aux pages 782 à 786

Les auteurs analysent la réalisation du matériel nécessaire à l'extension des facilités actuelles de calcul analogique de manière à inclure le mode répétitif/tiératif de fonctionnement.

La base du dessin est un pont à six diodes dans un montage de commutation à réaction.

La production d'ondes à modulation triangulaire de largeur d'impulsions par J. F. Young

Résumé de l'article aux pages 787 à 789

Pour les recherches pratiques concernant la performance des systèmes amplificateurs modulés en largeur d'impulsions, il est utile d'avoir une source d'impulsions dont la largeur varie linéairement ou triangulairement en fonction du temps. Une telle source peut être produite en confrontant deux ondes rectangulaires ayant des fréquences légèrement différentes et en utilisant soit un circuit logique ET soit un circuit logique EXCLUSIF-OU comme mélangeur.

Un stabilisateur de tension continue à transistor compensé par M. Pacak

Résumé de l'article aux pages 789 à 791

L'auteur décrit un circuit stabilisateur permettant d'assurer une stabilisation parfaite du courant continu ou de la tension continue réglable dans une gamme étendue, avec la stabilité à long terme de $\pm 1{,}5.10^{-4}$ dans 8 heures, la stabilité à court terme étant de $\pm 2{,}5.10^{-6}$ pendant plusieurs minutes. Le circuit comporte l'utilisation de la réaction positive et il est compensé contre les principaux effets perturbateurs.

Amplificateur à gain variable par S. Ghosh

Résumé de l'article aux pages 792 à 793

Il s'agit ici d'un amplificateur dont le gain peut être varié dans une gamme d'environ 60 dB en changeant une seule valeur de résistance, sans modifier appréciablement la caractéristique de fréquence, les impédances d'entrée et de sortie et le gain de circuit de l'amplificateur.

La réalisation d'une machine a additionner à chiffres binaires parallèles par J. B. Earnshaw et P. M. Fenwick

Résumé de l'article aux pages 794 à 796

Cet article décrit l'étude de base d'une machine à additionner à chiffres binaires parallèles et à propagation rapide. Le dessin est basé sur l'emploi d'un circuit logique du type á réaction et à diodes à transistors à deux niveaux dont le retard est de près de 1sec par niveau logique. Les mesures initiales indiquent que le temps de production de somme pour des mots à 16 chifires binaires est d'environ 50nsec.

Étude du cas le plus défavorable dans la réalisation d'un instrument d'entraînement d'impulsions par P. F. Jones

Résumé de l'article aux pages 797 à 799

L'auteur décrit l'étude du cas le plus défavorable pour des composants à excursion totale, étude où les valeurs restrictives de chaque composant sont calculées directement à partir des valeurs de limitation des paramètres de performance initiaux. Les composants de valeurs préférées qui sont ensuite choisies se trouvent à l'intérieur de ces limites calculées. La méthode indique un ensemble d'inégalites dans les paramètres de performance et résoud ces inégalités simultanément pour produire un second ensemble qui spécifie les valeurs de limitation pour chaque composant individuel.

Un compteur de décades binaires-quinaires utilisant la logique de résistance par R. Parshad et S. P. Suri

Résumé de l'article aux pages 800 à 801

Cet article traite d'un compteur binaire-quinaire utilisant la logique de résistance. Le circuit du compteur utilise moins de transistors que les circuits conventionnels. La lecture numérique est simple et économique dans l'emploi de composants. Le comptage bi-directionnel peut également être effectué sans difficulté avec ce circuit.

Un amplificateur à relais modulateur de courant continu à impédance d'entrée très élevée utilisant un nouveau transistor à effet de champs Ferranti par R. Verrill

Résumé de l'article aux pages 802 à 804

Le nouveau transistor type EXP380 a effet de champ qui vient d'être mis au point se caractérise par un très faible courant de fuite de l'orde de 10^{-10}A à 25°C et une très faible capacitance de porte d'environ 2,5pF. Ce dispositif a prouvé sa valeur dans un amplificateur expérimental de courant continu à relais modulateur série/shunt pour lequel une fuite et une capacitance de porte réduites sont des facteurs essentiels. L'amplificateur en question a une résistance d'entrée de plus de 100MΩ et la sensibilité sur la totalité de l'échelle est de ±1mW pour donner ±100µA dans l'instrument à cadre mobile à la sortie. La dérive de compensation à l'entrée, en fonction de la température, est d'environ 1µV/°C et 15pA/°C sans qu'il soit nécessaire d'adapter les transistors.

Un analyseur digital de la température de l'air de 0°F à 99°F par W. V. Dromgoole

Résumé de l'article aux pages 805 à 807

Cet instrument a été mis au point pour la conversion de la température de l'air sous une forme numérique convenant pour un système d'indications par lampe de dizaines et d'unités.

L'image par lampe fait partie d'un système qui indique le moment de lá journée pendant 47 secondes et la température de l'air pendant 10 secondes toutes les minutes et qui est visible sur une grande superficie des villes où le matériel est installé.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 818 bis 823

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Schuloszillograf

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(*Abbildung Seite 818*)

Ein neuer Oszillograf für Schulung und Industrie wurde von Advance Electronics Ltd herausgebracht. Der Oszillograf OS12 genügt dem Nuffield-Pflichtenblatt 158 und bietet Spezialeinrichtungen einschliesslich einer selbstsynchronisierenden Zeitablenkung, die automatisch mitnimmt, eines direktgekoppelten Verstärkers und—bei abgeschalteter Zeitablenkung—eines X-Eingangs auf der Rückseite des Gerätes, der Anwendung mit X-Y-Darstellung erlaubt.

Der OS12 hat einen Ablenkfaktor von 100 mV je Rasterteilung und Zeitablenkgeschwindigkeiten von 100 ms...100 μs pro Teilung. Das Gerät ist mit einer 7-cm-Röhre bestückt.

Der Oszillograf ist in sehr einfacher Konstruktion aus nur zwei vorgeformten Aluminiumstücken aufgebaut. Röhren und Bauelemente sind auf nur eine gedruckte Schaltung montiert. Advance erwartet, dass der Oszillograf in der Industrie weitgehend als eingebautes Überwachungssystem und in Prüfgeräten für die Fertigung Verwendung finden wird.

EE 01 751 *für weitere Einzelheiten*

Kondensatorendekade

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devonshire

(*Abbildung Seite 818*)

Eine neue kompakte Kondensatorendekade wurde von Hatfield Instruments Ltd für Entwurfsingenieure zur Anwendung bei der Toleranzbestimmung von Schaltungen und ähnliche Aufgaben entwickelt. Die Dekade gestattet sehr schnelle Kondensatorenauswahl im Bereich 100 pF bis zu 1 μF. Selbst wenn mehrere Geräte zusammen benutzt werden, nehmen sie sehr wenig Raum auf dem Tisch ein, und ihre Genauigkeit ist für jede Einstellung besser als 5 Prozent. Vier Dekaden geben Schritte von 100 pF, 1000 pF, 0,01 μF und 0,1 μF. Die Mindestkapazität ist mit allen Schaltern auf Null gestellt 30 pF und die Nennspannung 250 V— (0,1-μF-Dekade 100 V—). Abmessungen 140 × 41 × 70 mm.

EE 01 752 *für weitere Einzelheiten*

Federleisten

Ferranti Ltd, King's Cross Road, Dundee, Schottland

(*Abbildung Seite 818*)

Ferranti Ltd hat neue Federleisten für gedruckte Schaltungen herausgebracht, in denen hohe Zuverlässigkeit mit täuschend einfach wirkender Konstruktion verbunden ist.

Modelle mit einseitigen Kontakten haben auf der anderen Seite Lagereinsätze aus glasgefülltem Nylon anstelle von Blindkontakten. Diese einzigartigen Einsätze geben ausgezeichnete Langlebensdauer-Oberflächen mit geringer Reibung für die Druckschaltungsplatten und tragen zum konkurrenzfähigen Preis der Steckverbindung bei.

Die Presslinge der Federleisten sind aus blauem Diallylphthalat mit ausgezeichneten elektrischen und mechanischen Eigenschaften hergestellt und dank ihrer selbst erlöschenden Eigenschaften vollkommen betriebssicher und in der Lage, Löttemperaturen ohne weiteres auszuhalten.

Ein Merkmal der Kontaktkonstruktion ist die Walzblattfeder mit der ihr eigenen Beanspruchungsbegrenzung, die Überlastung unmöglich macht. Gemeinsam mit allen anderen Ferranti-Steckverbindungen hat die Kontaktfeder eine niedrige Ratecharakteristik; das bedeutet, dass die im Arbeitsbereich auftretenden Biegungen nur geringe Änderungen der Kontaktkraft hervorrufen, während die für Herausziehen und Einstecken erforderlichen Kräfte bemerkenswert niedrig sind. Die normalen Dicketoleranzen der Druckplatten werden durch Anpassung aufgefangen, die kontrollierten Kontaktkräfte reichen aus, um Anlauffilme zu unterbrechen, was niedrigen und konstanten Kontaktwiderstand gewährleistet, ohne Rücksicht darauf, wie niedrig die Betriebsspannung ist.

Federleisten der neuen Baureihe EWD werden mit Lötfahnen in 8-, 16-, 24-, 32- und 40poliger Ausführung einseitig und in 8/8-, 16/16-, 24/24-, 32/32 und 40/40poliger Ausführung zweiseitig geliefert. Sie werden für Druckschaltungsplatten von 1,5 mm Nenndicke und 3,81 mm Abstand zwischen Polen geliefert. Die Platte durchdringt 9,4 mm, und Einstecken der Platten in die Federleiste wird durch einen reichlichen Führungswinkel erleichtert.

Die Kontakte aus Berylliumkupfer sind auf mindestens 2,5 Mikron hartvergoldet, was eine nichtporöse und dauerhafte Oberfläche gewährleistet.

EE 01 753 *für weitere Einzelheiten*

Schreiber für drei Eingänge

Electronics Associates Ltd, Burgess Hill, Sussex

(*Abbildung Seite 818*)

Der Variplotter 1131 wurde für Anwendungen entwickelt, bei denen zwei Variable gleichzeitig über einer dritten aufgetragen werden sollen.

Drei unabhängig servogetriebene Systeme, eine Grundempfindlichkeit von 1 mV/Zoll (etwa 0,39 mV/cm), 18 geeichte Bereiche und stufenlos regelbare Massstabpotentiometer gewährleisten grosse Anpassungsfähigkeit.

Eine eingebaute Zeitablenkung gibt eine zusätzliche t-y-Kurvenschreibmöglichkeit, die volles Überstreichen von sechs Bereichen von 0,2 s/cm bis zu 8 s/cm erlaubt. Getrennte Regler für Zeitablenkung und Abheben des Schreibstiftes gestatten Leerversuche zur Massstabbestimmung vor dem Auftragen.

Der 1131 hat Festkörperschaltungen und ein Zenerdioden-Bezugssystem, lekkagefeste Einsteck-Tintenpatronen, von hinten beleuchtete Schalter, Einsteckfilter zur Unterdrückung von Netzfrequenz und Rauschen, sowie eine Gebläseeinrichtung zum Festhalten des Papiers.

Die technischen Daten sind wie folgt:

statische Genauigkeit ±0,1 Prozent, dynamische Genauigkeit ±0,2 Prozent bei 178 mm/s; Wiederholbarkeit ±0,05 Prozent. Schwenkgeschwindigkeit des Armes 432 mm/s und des Schreibstiftes 508 mm/s. Eingangswiderstand 2,5 MΩ/V in Bereichen bis zu 8 mV/cm und für höhere Bereiche durchweg 1 MΩ. Der in Zoll oder metrischen Einheiten geeichte Schreiber ist mit Aussenabmessungen von 19" × 533 mm hoch und 165 mm tief für Gestelle oder Tisch lieferbar.

EE 01 754 *für weitere Einzelheiten*

Statisches Verzögerungsrelais

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(Abbildung Seite 819)

Solid State Controls Ltd hat die Fertigung eines transistorierten Verzögerungssteckrelais aufgenommen, das bei hoher Zuverlässigkeit und niedrigem Preis besonders zur Befriedigung des Bedarfs der modernen Industrie entwickelt wurde.

Typ TDR/407 ist ein robuster Baustein, völlig eingekapselt und ohne bewegliche Teile, die sich abnutzen oder ausfallen können. Er wird weder durch Vibrationen oder Stösse, noch Einschwingspannungen beeinflusst und kann mindestens 100 Millionen kontinuierliche Schaltspiele aushalten. Dieser kompakte Baustein ist selbst nach vielen tausend Betriebsstunden driftfrei und erfordert keinerlei Wartung.

Modell TDR/407 ist in drei Spannungsbereichen lieferbar: 12...35 V, 35...70 V und 70...150 V und arbeitet selbst bei grossen Schwankungen der Betriebsspannung und Umgebungstemperatur genau. Die Hauptzeitgeberschaltung besteht aus einer völlig konstanten, niederohmigen Verzögerungsschaltung mit steuerbarem Siliziumgleichrichter-Ausgangsschalter. Dieses Halbleiterelement kann induktive oder reinohmsche Lasten bis zu 1 A bei 150 V schalten und die üblicherweise bei Schützen und Motoranlassern auftretenden hohen Stromstösse vertragen.

Laufzeiten sind in vier Bereichen von 0,2 Sekunden bis zu 11 Minuten lieferbar und können mittels eines Reglers am Gehäuse gewählt oder voreingestellt werden. Die typische Zeitgenauigkeit ist bei extremen Umgebungstemperaturen und Betriebsspannungen 5 Prozent, die Wiederholgenauigkeit 1 Prozent.

Wegen seiner hohen Zuverlässigkeit und Festkörpertechnik eignet sich der Baustein besonders für Anwendungen in Flugzeugen und der Industrie. Unter den vielen bereits in Betracht gezogenen Anwendungsmöglichkeiten sind zu nennen: Verfahrenstechnik und Verkehrskontrolle, Motorsteuerungen, Fahrstuhlsteuerungen, Kraftwerke, Fernsprechämter, Schalteinrichtungen für Bord- und Bodentelemetriekanäle, Steuerung von Notaggregaten und automatischen Verpackungsmaschinen.

EE 01 755 *für weitere Einzelheiten*

Thyristor-Steuerempfindlichkeitsmesser

Caltronics Ltd, Huntingate, Hitchin, Hertfordshire

(Abbildung Seite 819)

Mit dem Thyristor-Steuerempfindlichkeitsmesser kann der zum Zünden erforderliche Steuerstrom einer breiten Auswahl von Thyristoren genau und schnell gemessen werden. Die an den Prüfling gelegte Anoden-Kathoden-Konstantspannung von 6 V wird von einer Zenerdiode abgegeben.

Die Steuereigenschaften des Prüflings werden durch Anlegen eines halbwellengleichgerichteten 50- oder 60-Hz-Impulses zwischen Steuerelektrode und Kathode des Thyristors gemessen. Ein Triggernetzwerk in der Anodenspannungsleitung fühlt den Einschaltpunkt des zu prüfenden Thyristors und betätigt einen elektronischen Schalter, der das Steuersignal abschaltet. Eine Spitzenspannungsmesserschaltung zeigt direkt den Strom- oder Spannungspegel an, bei dem der Thyristor zündet. Für Anzeige des Steuerstrompegels für die Zündung wird die Steuerschaltung durch Präzisionswiderstände getrieben, die eine regelbare 10stufige Stromquelle bilden. Treiben der Steuerschaltung von einer 3stufigen Spannungsquelle sorgt für die zur Zündung erforderliche Steuerspannung.

Der Empfindlichkeitsmesser lässt sich mittels Eichpotentiometern sowohl auf Spannung wie Strom eichen.

Klemmen für Parallelfernbetrieb des Gerätes sind vorhanden. Fernanzeige kann über Frontplattenquellen angeschaltet werden, denen man für Vollausschlag des eingebauten Messinstrumentes ein Signal von +1 V entnehmen kann.

EE 01 756 *für weitere Einzelheiten*

Analog-Digital-Umsetzer

Ether Engineering Ltd, Park Avenue, Bushey, Hertfordshire

(Abbildung Seite 819)

Der Umsetzer MA 1102 wandelt analoge Eingangsspannungen zwischen 0 und −10 V in einen rein binärcodierten 10-Bit-Ausgang um.

Zur Erreichung genauer Umsetzmöglichkeiten mit geringem Aufwand arbeitet das Gerät nach dem stufenweisen Annäherungsverfahren. Ein grosser Teil der Schaltung besteht aus Doppel-NOR-Moduln in Dünnfilmtechnik, die je nach den Anforderungen der Schaltung als Tore, bi- oder monostabile Elemente oder Inverter zusammengeschaltet sind.

Kompakte Konstruktion gestattet Einbau des Umsetzers, der nur 44,5 mm Feldhöhe in Anspruch nimmt, in bestehende Systeme.

Der analoge Höchsteingangswert ist 10 V und entspricht 2^{10} Binärziffern. Das Auflösungsvermögen ist ±1 Ziffer, die Fehlergrenze 0,1 Prozent und die Umsetzzeit 200 µs.

EE 01 757 *für weitere Einzelheiten*

Rauschverminderungssystem

Dolby Laboratories, 590 Wandsworth Road, London, S.W.8

(Abbildung Seite 820)

Der "S/N Stretcher" (Rauschabstanddehner) der Dolby Laboratories ermöglicht Bekämpfung des Rauschens in hochwertigen Tonübertragungen und Tonaufnahmesystemen. Der "S/N Stretcher" kann überall dort eingesetzt werden, wo das Signal sowohl am Eingang wie am Ausgang der Tonfrequenzkette für Verarbeitung zur Verfügung steht. Während das Signal selbst in unveränderter Form abgegeben wird, schwächt die Rauschverminderungswirkung die üblichen Rauscharten, die bei der Erstellung des Originals und der Nachsynchronisierung in der Filmindustrie, bei Videobandaufnahmen und in Landleitungen usw. auftreten.

Durch den bekannten Verdeckungseffekt des menschlichen Ohrs wird leiser Ton durch lauten Ton teilweise oder auch vollständig verdeckt, besonders wenn der Frequenzunterschied nicht gross ist. Der subjektive Rauschabstand bei Magnetbandaufnahmen ist beshalb besser, als von der Analyse der Wiedergabewellenform des Bandes zu erwarten wäre; das Bandrauschen hängt von der Momentanamplitude des Signals ab und wächst mit ihm: ein mit Modulationsrauschen bezeichneter Effekt. Da das Ohr unter normalen Signalbedingungen das zusätzliche Rauschen wirksam unterdrückt, liegt der Gedanke nahe, dass ein geeignetes elektronisches System unter Ausnutzung des Verdeckungseffektes das Rauschen noch weiter reduzieren könnte.

Durch Anwendung des Verdeckungsprinzips in Verbindung mit Signalpressung- und -dehnung erreicht der "S/N Stretcher" Rauschverminderung (a) durch Verstärken der Signalkomponenten mit niedrigem Pegel, wenn immer das während der Aufnahme möglich ist (Pressung), und komplementäre Abschwächung während der Wiedergabe (Dehnung); (b) durch den Verdeckungseffekt, wenn der Signalpegel so hoch ist, dass Pressung und Dehnung unmöglich sind.

Da bei Rauschfrequenz mit etwas grösserem Abstand von der Signalfrequenz Verdeckung nicht so wirksam ist, müssen die verschiedenen Abschnitte des Spektrums unabhängig voneinander behandelt werden. Das Rauschverminderungssystem gibt einen niedrigeren—und scheinbar konstanten—Rauschpegel, da das klassische Husch-Husch oder Sausen der üblichen Pressung und Dehnung abwesend sind.

Der "S/N Stretcher" teilt das Tonfrequenzspektrum in vier Bänder und presst und dehnt jedes derselben im wesentlichen unabhängig voneinander. Es gibt getrennte Bänder für die Brumm- und Rumpelfrequenz, für mittlere, mittelhohe und hohe Frequenzen. Ein hochpegliges Signal in einem Band kann daher Rauschverminderung in einem anderen Band mit niedrigem Signalpegel nicht verhindern.

Von einem anderen Punkt gesehen gibt das System eine wirksame Aufnahmeentzerrung, die sich laufend dem Eingangssignal auf eine Weise anpasst, die den Rauschabstand während der Wiedergabe optimiert.

EE 01 758 *für weitere Einzelheiten*

Schallsignalquelle

A. P. Besson & Partner Ltd, St. Josephs Close, Hove, Sussex

(Abbildung Seite 820)

"Bleeptone" wurde als kompakte und hochzuverlässige Signalquelle entwickelt. Sie besteht aus einer geschlossenen Oszillatorschaltung und einem Schwingankerhörer, die aus einer Niedervoltbatterie gespeist werden. In Abmessungen und Ausführung ist es der bekannten Schwingankerhörmuschel der modernen Fernsprechgeräte ähnlich. Der vom Strahler abgegebene Ton ist im allgemeinen auf bis zu 152 m Entfernung hörbar, und die Frequenz ist so gewählt, dass sie nicht merkbar durch den Hintergrund maskiert wird.

Die Treiberschaltung ist der eines Stimmgabelgenerators ähnlich. In der magnetischen Schaltung sind zwei getrennte Spulen so angeordnet, dass eine den Schwinganker treibt und die andere auf das induzierte Signal anspricht und es zur Basis des Transistors, der mit zugehörigen Bauelementen die komplette Verstärkerschaltung bildet, zurückkoppelt.

Die Grundarbeitsfrequenz wird im wesentlichen durch die mechanischen Eigenschaften des Schwingankers bestimmt.

Das Gerät hat Anwendungsmöglichkeiten auf vielen Industriegebieten. Es ist vor allem dort als Alarmtonquelle geeignet, wo höhere Zuverlässigkeit als die mit dem herkömmlichen Summer erreichbare erforderlich ist.

EE 01 759 *für weitere Einzelheiten*

Sprechfunkverstärker

The Plessey Co. Ltd, Ilford, Essex

(Abbildung Seite 820)

Im Werk Bridgnorth der Plessey-Elektronikgruppe wurde ein neuer Transistorverstärker für Sprechfunkausrüstungen entwickelt. Ursprünglich war er für die Sprechfunkgeräte 700 und 700R der Firma gedacht und kann mit seiner 14-dB-Verstärkung die Ausgangsleistung auf 8 W erhöhen. Man kann ihn aber auch in bestehende Funkfernsprechverbindungen einbauen. Die Abmessungen sind nur 76 mm hoch, 159 mm breit und 76 mm tief.

Ausser der Grundausführung gibt es auch noch ein Gerät in hermetisch dichtem Gehäuse und tropenfest für Einsatz im Freien in beliebigen Teilen der Welt.

Das für 12 V— ausgelegte Gerät ist auch für Betrieb im Frequenzbereich 68...174 MHz lieferbar.

Mit einem Stromverbrauch von nur 1,6 A ist der neue Verstärker als Zusatz zu beweglichen Ausrüstungen oder Energiequelle zum Treiben von Veraktor-Vervielfachern bis in den GHz-Bereich geeignet.

EE 01 760 *für weitere Einzelheiten*

Stromversorgungen

Vertrieb: Scientific Measurements & Equipment Ltd, 78 Main Street, Queniborough, Leicester

(Abbildung Seite 820)

Diese von Knott Elektronik in München gefertigten Speisegeräte sind für Verwendung mit elektronischen Ausrüstungen bestimmt, die äusserst konstante und genaue Gleichspannungen benötigen, wie beispielsweise GM-Zähler, Proportionalzähler, Ionisationskammern, Bildverstärker, Photovervielfacher, Szintillationszähler, Elektronenstrahlröhren usw.

Hoher Ausgangsstrom und sehr niedriger Innenwiderstand gestatten gleichzeitige Speisung mehrerer Geräte ohne Wechselwirkung. Haupteigenschaften der Geräte sind ihr hohes Regelverhältnis, Langzeitkonstanz, einstellbarer Schutz und Einstellbarkeit auf 1 V. Die Geräte sind im Gehäuse oder für Gestelleinbau lieferbar.

Das Lieferprogramm umfasst Geräte von 0...30 V, 10 A bis zu 0...15 kV, 10 mA. Es gibt auch ein Konstantstromgerät für 0...10 A bei 13 V.

EE 01 761 *für weitere Einzelheiten*

Allstrom-Millivoltmeter

Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorkshire

(Abbildung Seite 820)

Das volltransistorierte Millivoltmeter TM1 hat 12 Bereiche für Wechsel- sowie Gleichstrom mit Skalenendwerten von 1 mV bis zu 300 V in 1-3-10 Folge. Ausserdem gibt es eine dB-Skala von −10...+2 dB (0 dB = 1 mW an 600 Ω).

Die Fehlergrenze ist für Gleichstrom 3% des Skalenendwertes ±100 μV und für Wechselstrom 4% des Skalenendwertes ±100 μV von 20 Hz...100 kHz. Der Gleichstromeingangswiderstand ist 1 MΩ/V bis zu 10 V und konstant bei 10 MΩ für höhere Bereiche; die Eingangsimpedanz bei Wechselstrom ist 100 kΩ von 1 bis 30 mV, 1 MΩ von 100 bis 300 mV und 10 MΩ von 1 bis 300 V. Die höchste Parallelkapazität ist 40 pF.

Für Gleichstrom hat das Millivoltmeter einen Widerstands-Eingangsabschwächer mit nachgeschaltetem Gleichstromverstärker mit Messinstrument und Rückkopplungsregelung. Im Wechselstromteil speist ein frequenzkompensierter Abschwächer einen galvanisch gekoppelten Verstärker mit Rückkopplungsschleife und Messinstrument. In den vier untersten Bereichen hält ein impedanzwandelnder Verstärker die Eingangsimpedanz bei 100 kΩ.

Das Gerät wird aus zwei Batterien PP11 gespeist und hat einen eingebauten Spannungstest. Es ist 165 mm hoch, 218 mm breit und 220 mm tief und wiegt ohne Batterien 2,86 kg.

EE 01 762 *für weitere Einzelheiten*

Anzeigegerät für Messleitungen

Marconi Instruments Ltd, Sanders Division, Gunnels Wood Road, Stevenage, Hertfordshire

(Abbildung Seite 821)

Der von der Sanders Division der Marconi Instruments Ltd angekündigte neue Stehwellenverhältnis-Anzeiger 6596 ist ein rauscharmer, preiswerter selektiver Verstärker. Wegen seiner Empfindlichkeit und seinem Verstärkungsbereich eignet er sich für die meisten Mikrowellenmessungen, in denen Kristalldetektoren angewendet werden. Das HF-Signal wird mit 1 kHz moduliert.

Grob- und Feinregelung der Verstärkung geben einen für demodulierte Signale bis zu 1 mV_{eff} ausreichenden Regelbereich.

Die Skala des Messwerkes ist in Stehverhältnissen geeicht, was die Annahme voraussetzt, dass der Detektor im quadratischen Gebiet der Kurve arbeitet, d.h. dass die Ausgangsspannung der Mikrowellenleistung proportional ist.

Die Höchstempfindlichkeit ist besser als 1 μV Vollausschlag und das Rauschen niedriger als 0,1 μV. Ein Eingangswähler gestattet getrenntes Anschalten jedes der zwei Eingangssignale. Für Anwendung einer extraempfindlichen Mikrowellenbrückentechnik kann auch die Differenz der beiden Signale angeschaltet werden.

Das Gerät eignet sich ideal für Einsatz mit dem Sanders-Arbeitsplatz 599 für Ausbildungszwecke.

EE 01 763 *für weitere Einzelheiten*

Miniatur-Keramikkondensatoren

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

(Abbildung Seite 821)

Mullard hat eine neue Baureihe hochwertiger Miniaturkondensatoren in rechteckiger Form und nur 1,9 mm dick angekündigt, die auf Druckschaltungsplatten mit 2,54 mm Rastermass hohe Packdichten gestatten. Die Kondensatoren sind sowohl für Unterhaltungs- wie Industrieanwendung geeignet. Sie bestehen aus einer dünnen, metallisierten Keramikplatte, die mit Schutzlack isoliert ist und ihre guten Eigenschaften unter den feuchtesten Bedingungen beibehält.

Jeder Kondensator der neuen Baureihe (C 333) hat ohne die 13,5 mm langen Anschlüsse Abmessungen von 5 × 8,5 × 1,9 mm. Der Kapazitätsbereich ist 3,9 bis 150 pF mit einer Toleranz von 0,5 pF oder ±2%, wobei jeweils die grössere gilt. Die Betriebsspannung ist über den Temperaturbereich −25°... +85°C 40 V und die bei 10 V gemessene Isolation besser als 1 GΩ. Der Temperaturkoeffizient hängt vom Kapazitätswert ab und liegt zwischen Null und −750 × 10^{-6}.

Durch enge Toleranzen und hohe Konstanz sind die neuen Kondensatoren für Verwendung in ZF-Trafos für Fernseher, abgestimmte Kreise und andere Anwendungsgebiete geeignet, in denen niedrige Verluste und hohe Leistung wichtig sind.

EE 01 764 *für weitere Einzelheiten*

Präzisionsmessbrücke

The Wayne Kerr Co. Ltd, Sycamore Grove, New Malden, Surrey

(*Abbildung Seite 821*)

Die neuste Ergänzung des Wayne-Kerr-Programmes für Autobalance-Messbrücken ist das Modell B331, in dem die in Normallabors erforderliche Genauigkeit mit der Geschwindigkeit und einfachen Arbeitsweise der elektronischen Nullung vereint ist. Zwei Messinstrumente zeigen gleichzeitig die gleichphasige und Querkomponente beliebiger Bauelemente oder komplexer Impedanzen an. Mit beleuchteten einzeiligen Anzeigen zusammenwirkende Drucktasten gestatten Kompensation beider Anzeigen durch drei Dekaden, was in allen Bereichen ein 6stelliges Auflösungsvermögen gibt.

Sonderzweckschaltungen kompensieren automatisch für die Impedanz der Messleitungen, die mit den neusten Kelvin-Klemmen ausgerüstet sind. Die Ausgänge können Digitalvoltmeter, Drucker, Schreiber, Klassiereinrichtungen und Steuerschaltungen treiben. Für Vergleichsmessugen lassen sich externe Normale über vorhandene Buchsen anschliessen, wobei man mit Feinregelung ein Unterscheidungsvermögen von 10 × 10^{-6} erreichen kann.

Die interne Quelle und der Demodulator arbeiten bei 1,00 kHz, können aber mit externen Zusätzen im Bereich 50 Hz...20kHz betrieben werden. Gesamtmessumfang bei 1 kHz ist 0,0001 pF...0,25 F, 1 pS...1 kS, 1 mΩ...1 TΩ und 100 nH...250 MH. Zwischen 1 pF und 10 μF sowie 10 nS und 100 mS ist die Messunsicherheit 0,01 Prozent. Das Gerät ist 19″ (483 mm) breit, 305 mm hoch, 229 mm tief und wiegt etwa 22,5 kg.

EE 01 765 *für weitere Einzelheiten*

Zungenrelais

Hendrey Relays & Electrical Equipment Ltd. 390-394 Bath Road, Slough, Buckinghamshire

(*Abbildung Seite 821*)

Das Zungenrelais 5858 für Bord- oder Bodeneinsatz entspricht den Anforderungen der britischen Norm BS. 2G.100; seine Grenzwerte sind 5 Hz... 2 kHz bei 10 g und 50 g Beschleunigung in beliebiger Richtung. Erregerspule und Zungenaggregat sind in ein Stahlgehäuse gekapselt, das mechanischen sowie magnetischen Schutz gibt. Die Vergussmasse besteht aus Epoxydharz, aus dem auch der die Anschlussstifte tragende Relaissockel hergestellt ist. Der Baustein ist daher sehr robust, feuchtigkeitsfest und für Einsatz unter rauhesten Bedingungen mit Umgebungstemperaturen von −60°C ...+120°C geeignet.

Sorgfältige Auswahl und Formung der Zungen vor Verwendung garantieren mindestens 2 × 10^6 Spiele bei Vollast, eine Kontaktansprechzeit von besser als 2 ms mit vernachlässigbarem Prellen beim Schliessen und Öffnen.

Höchstbelastbarkeit ist 0,5 A oder 30 V— oder 10 W bei reinohmscher Last. Induktive Lasten müssen in geeigneter Weise gelöscht werden. Es gibt drei Standardkontaktpakete: (a) zwei Arbeitskontakte, (b) vier Arbeitskontakte und (c) zwei Arbeits- und zwei Ruhekontakte. Spulen können für jede Gleichspannung bis zu 100 V gewickelt werden.

Das Relais ist mit Löthaken oder freien Zuleitungen, für Montage auf Träger oder mittels Schelle, oder mit Stiften für Einsteckfassungen lieferbar.

EE 01 766 *für weitere Einzelheiten*

Verzögerungsschalter

Solid State Controls Ltd, 30-40 Dalling Road, London, W.6

(*Abbildung Seite 822*)

Der "Radian" ist ein preisgünstiger, hochwertiger Ersatz für thermische Verzögerungsschalter, einfache Kondensatoren-Relaiszeitgeber und Synchronmotoreinrichtungen, wenn es auf hohe Wiederholungsgenauigkeit, Langzeitkonstanz und sehr schnelle Freiwerdezeit ankommt. Ungleich thermischen Einrichtungen braucht die Radian-Relaisverzögerung keine Abkühlzeit, sondern kann sofort zurückgestellt werden und gibt äusserst reproduzierbare Zeiten. Der Baustein ist für 24 V— und 24 V∼ Betriebsspannung lieferbar und kann mit einer breiten Auswahl von Relais und Kontaktgebern bestückt werden.

Der "Radian" ist ein Einsteckbaustein, dessen Verzögerung entweder bei Erregen oder Abschalten beginnen kann und mit zwei Zeitbereichen 1...60 s und 1...5 min lieferbar ist.

Einschwingvorgänge an den Speiseleitungen, Schwankungen der Betriebsspannung um ±15 Prozent und der Temperatur zwischen −10° und +55°C beeinflussen den Baustein nicht.

EE 01 767 *für weitere Einzelheiten*

Konstantstromversorgung

Servomex Controls Ltd, Crowborough, Sussex

(*Abbildung Seite 822*)

Für Anwendungen, die keinen Gestelleinbau erfordern, hat Servomex Controls Ltd Ausführungen ihrer Spannungskonstanthalter AC2 und AC7 entwickelt. Die neuen Modelle AC2 Mk 111A und AC7 (industriell) sind einfacher zu installieren und warten, da sie für Installation durch Betriebselektriker anstelle von Elektronikern ausgelegt sind. Alle periodisch zu wartenden Teile sind nach Entfernung der Abdeckung leicht zugänglich, ohne dass das ganze Instrument—wie bei den Modellen für Gestelleinbau—ausgebaut werden muss, um von der Rückseite Zugang zu erlangen.

Die einzelnen elektrischen Bauelemente sind mit denen der Gestelleinbaumodelle identisch, und das gleiche gilt für die Qualität und Facharbeit der Herstellung dieser neuen Typen. Modell AC2 Mk 11A ist kleiner und handlicher als der Gestelleinbautyp, während Modell AC7 (industriell) auf Laufrollen auf dem Boden steht und schwallwassersowie staubgeschützt ist. Es ist daher für Montage im Keller oder einem oft gefegten Durchgang ideal geeignet.

Beide Geräte arbeiten mit Eingangsspannungen von 200...250 V und geben eine hochkonstante Ausgangsspannung ab, die nicht mehr als ±0,1 Prozent von der Nennspannung des Netzes abweicht. Sie bleiben durch Gleichrichterbelastung oder Änderung des Leistungsfaktors von reinohmisch auf reinreaktiv einschliesslich kapazitativer Belastung unbeeinflusst.

Die Konstanthalter arbeiten mit einem Regeltransformator, der je nach Bedarf Zusatz- oder Kompensationsspannungen erzeugt, um die Eingangsspannung auf den gewünschten Wert zu bringen. Der Regeltrafo wird durch einen Zweiphasen-Stellmotor mit gekapseltem Getriebe und Drehmomentbegrenzer getrieben. Änderungen der Ausgangsspannung werden durch eine Bolometerbrücke entdeckt, deren Ausgang in einem Zweiröhren-Regelverstärker verstärkt und an den Stellmotor gelegt wird.

Die Geräte sind ungewöhnlich robust gebaut und können Stösse bis zu 40 g in jeder Richtung aushalten. Weder Relais noch Thyratrons oder Elektrolytkondensatoren finden Verwendung.

Die neuen Typen sind nicht nur der einfacheren Gehäuse wegen billiger, sondern auch weil Voltmeter, Amperemeter, Ein-Ausschalter und Hauptsicherung weggelassen wurden. Die letzteren beiden Bauelemente sollen—zusammen mit einem auf Wunsch lieferbaren externen Voltmeter—an die Wand montiert werden.

EE 01 768 *für weitere Einzelheiten*

Digital-Dehnungsanzeiger

S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Abbildung Seite 822)

Der Digital-Dehnungsanzeiger SE.601 ist ein tragbares Instrument für dynamische und statische Dehnungsmessungen mit vielen neuen Eigenschaften. Die Digitalanzeige ist für den Bedienenden am bequemsten und gibt mit Hilfe des Schalters "Zuzählen" einen Messumfang von ±50 000 Mikrodehnung mit einer Messunsicherheit von ±0,1 Prozent der Anzeige oder ±5 μE, wobei jeweils der grössere Wert gilt (k = 2,0).

Interne aufladbare Zellen gestatten langzeitigen Einsatz im Aussendienst. Ein weiteres Merkmal ist der Regler "Empfindlichkeitsfaktor", den man auch benutzen kann, wenn direkte Anzeigen von Lastdosen, Druck- und Weggebern usw. gewünscht werden.

Rechteckwellenerregung gestattet Einsatz mit Dehnstreifenschaltungen, die bis zu 0,01 μF Kapazität haben, ohne merkbare Reduktion der Messgenauigkeit. Auf Wunsch lässt sich auch sinusförmige Erregung vorsehen.

Das Instrument ist für Dehnungsstreifen mit 50 Ω ... 2 kΩ Widerstand geeignet. Getrennten Buchsen für Galvanometer- und Oszillografenanschluss können 1 mA und 2 V zum Treiben herkömmlicher Anzeiger und Schreiber entnommen werden.

EE 01 769 *für weitere Einzelheiten*

Selbsthaltende Magnetspule

H. E. & B. S. Benson Ltd, Exning Road, Newmarket, Suffolk

(Abbildung Seite 822)

Eine völlig neue selbsthaltende Magnetspule wird von H. E. & B. S. Benson Ltd angekündigt. Der Tauchkern besteht aus Dauermagnetmaterial. Wenn er durch einen kurzen Impuls eingezogen wird, bleibt er für eine unbegrenzte Periode in dieser Position und trägt eine Last, ohne Strom zu entnehmen. Ein Umkehrimpuls lässt den Tauchkern abfallen.

Jeder gleichgerichtete Wechselstrom oder Gleichstrom (einschliesslich Trockenbatterie) kann diese Magnetspule betätigen. Es gibt eine Wechselstromschaltung, die nur eine Diode und einen kleinen Widerstand erfordert, die in den Betätigungsschalter gelötet werden können.

Vorteile sind u.a. Abwesenheit eines Temperaturanstiegs, daher maximale Betriebssicherheit (auch nützlich in hoher Umgebungstemperatur); eine breite Auswahl von Kraftkenndaten, die alle nach momentanem Schalten dauernd halten und selbst bei Ausfall der Erregerspannung weiter halten.

Die selbsthaltenden Magnetspulen PMB sind äusserlich den bestehenden Typen B.2, 3, 4 und 5 gleich. PMBs können Stromstossrelais ersetzen und haben nach Ausrüstung mit einer Rückholfeder das definitive Zwei-Positions-Verhalten einer teureren doppelwirkenden Anordnung mit Doppelspule.

EE 01 770 *für weitere Einzelheiten*

Digitaluhr

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

(Abbildung Seite 823)

Venner Electronics Ltd hat eine Allsilizium-Digitaluhr TSA 6686 mit Glasfaser-Druckschaltungskarten eingeführt, die auf sechs gasgefüllten Ziffernanzeigeröhren Stunden, Minuten und Sekunden darstellt. Die Uhr ist je nach Wunsch mit 12- oder 24-Stunden-Basis lieferbar und mit elektrischen Ausgängen für Anschluss eines Druckers oder Streifenlochers ausgerüstet.

Start-Stoppeinrichtungen sind entweder durch negative Impulse von 4 V Amplitude oder durch Drucktasten auf der Frontplatte zu betätigen. Die Anzeige kann manuell für unabhängige Anzeige von Stunden, Minuten und Sekunden eingestellt werden und hat eine Rückstelleinrichtung auf Null.

Die für einen Temperaturbereich von −5°C...+60°C konstruierte Uhr arbeitet mit 2×10^{-6} Genauigkeit.

Sie ist für Netzanschluss von 120... 125 V∼ oder 200...250 V∼, 50 Hz sowie Einbau in ein 19"-Gestell ausgelegt, 133 mm hoch und 241 mm tief.

EE 01 771 *für weitere Einzelheiten*

Wanderfeldröhre

The M-O Valve Co. Ltd, Brook Green Works, London, W.6

(Abbildung Seite 823)

Die M-O Valve Co Ltd hat einen neuen Hochleistungs-Impulswanderfeldverstärker TWX16 für Anwendung in Radarsystemen eingeführt. Er gibt bei 5 kW Spitzenausgangsleistung (30 W mittlere Leistung und 5...20 kW gesättigte Spitzenleistung) 40 dB Verstärkung über 500 MHz Bandbreite im Bereich 8...9,3 GHz.

Die Röhre ist in Metall-Keramiktechnik ausgeführt und arbeitet mit einer Ring- und Stabstruktur für die langsame Welle, die Störschwingungsfreiheit gewährleistet. Fokussiert wird in einem Magnetspulenzusammenbau SMX16, der auch die HF-Kopplungen und die Wärmeableitung enthält.

Zur Vereinfachung der Impulsspeiseanforderungen wird der Wendel des TWX16 auf einem Gleichspannungspegel gehalten.

Die 1-Watt-Wanderfeldröhre M-O V-TWX8 eignet sich für Einsatz als Treiber.

EE 01 772 *für weitere Einzelheiten*

Klystron-Stromversorgung

Vertrieb: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(Abbildung Seite 823)

Die hochkonstante Stromversorgung LS 525R von Oltronix (Schweden) wurde für den Betrieb von Reflexklystrons entwickelt. Strahl-, Reflektor-, Gitter- und Heizerspannungen werden getrennt geregelt. Die Reflektorspannung kann für Dauerstrichbetrieb des Klystrons unmoduliert sein, oder sie kann mit einer intern erzeugten Rechteck-, Impuls-sägezahn- oder Sinuswelle oder mit einem externen Signal moduliert werden. Die Gitterspannung lässt sich intern mit Rechteckwellen oder Impulsen modulieren. Die Rechteckwellen- und Impuls-Modulationsspannungen sind durch den Dauerstrichpegel des Gitters oder Reflektors begrenzt.

Die Strahlspannung ist zwischen −200 V und −3600 V bei 0 ... 125 mA regelbar. Eine Netzspannungsänderung um 10 % verursacht eine Ausgangsänderung von unter 300 mV und eine Änderung von Leerlauf zu Vollast eine von weniger als 200 mV.

Der Ausgang ist kontinuierlich in vier Bereichen regelbar: 200 ... 1200 V, 1000 ... 2000 V, 1800 ... 2800 V und 2600 ... 3600 V. Das Gerät ist mit einer zwischen 10 und 125 mA einstellbaren elektronischen Sicherung völlig gegen Überlastung geschützt.

Das Gerät ist auch für Anwendung mit Photovervielfachern geeignet, und zwar wenn über 100 Photovervielfacher von derselben Stromversorgung betrieben werden können.

In der Abbildung wird oben das LS 525R und unten ein zusätzliches Treibergerät für Klystrons gezeigt.

EE 01 773 *für weitere Einzelheiten*

Breitband-Millivoltmeter

Vertrieb: Livingston Laboratories Ltd, Livingston House, Greycaines Road, North Watford, Hertfordshire

(Abbildung Seite 823)

Radiometer in Kopenhagen hat einen Nachfolger des Röhrenvoltmeters RV33 angekündigt. Es handelt sich um das neue Breitband-Millivoltmeter RV35 mit Überlastungsschutz bis zu 500 V— oder ∼ und einem Messumfang von 10 μV ...

300 V. Über den grösseren Teil des Frequenzbereiches 10 Hz ... 6 MHz ist die Messunsicherheit 2 Prozent.

Ausserdem lässt sich Modell RV35 auch als Verstärker mit hoher Verstärkung mit einem Ausgangsspannungsendwert von etwa 80 mV an 75 Ω einsetzen.

EE 01 774 *für weitere Einzelheiten*

Keramische Elektronenstrahlröhre

Ferranti Ltd, Glen Mill, Oldham, Lancashire

(*Abbildung Seite 823*)

Die wahrscheinlich erste völlig elektrostatisch fokussierte keramische Elektronenstrahlröhre wurde vom Geschäftsbereich Elektronik der Ferranti Ltd eingeführt. Die für Widerstandsfähigkeit unter rauhesten Vibrationsbedingeungen konstruierte neue Röhre vereint hohes Auflösungsvermögen mit einem nutzbaren Schirmdurchmesser von 12,7 mm.

Die Gesamtlänge der Röhre ist 114,3 mm, der Körperdurchmesser 15,2 mm. Der Schirmträger hat einen nutzbaren Schirmdurchmesser von 12,7 mm und ist aus speziell für Verwendung mit keramischen Kolben entwickeltem Qualitätsglas hergestellt und mittels Sonderzweckverfahren mit der Keramik verschmolzen. Eine typische Spannung für die erste Anode ist 300 V und die Gesamtspannung 8 kV. Bei 450 V an der ersten und 7 kV an der Endanode ist die Zeilenhelligkeit in der Grössenordnung von 65 000 Lux. Die Röhre wiegt mit Ablenkspulen und MU-Metallschirm unter 110 g.

Die Teile des Strahl- und Fokussiersystems sind mit sehr engen Toleranzen bearbeitet, was ermöglicht, sehr gute Bildschärfe und ein Auflösungsvermögen von besser als 500 Zeilen zu erzielen und eine Informationsmenge darzustellen, die normalerweise eine Elektronenstrahlröhre mit viel grösserem Schirmdurchmesser erfordert.

EE 01 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Steuerungssystem für einen Molekularvakuummesser

von R. G. Christian

Zusammenfassung des Beitrages auf Seite 772-777

Ein mit einem Schwingplatten-Molekularvakuummesser verwendetes elektronisches Steuergerät wird beschrieben. Die an die Steuerung des Messers, dessen Messfühler und Treiberanordnung als Differentialkondensator ausgebildet sind, zu stellenden Anforderungen werden besprochen.

Nach der angewandten Messfühlermethode wird die Kapazitätsänderung gemessen, während die Platte mit Hilfe eines Frequenzmodulationssystems, das eine Spannung abgibt, die eine Funktion der Schwingungsamplitude ist, schwingt. Rückkopplung dieser Spannung über geeignete Schaltungen treibt den Messer. Die Schwingungsamplitude wird mittels einer Regelschaltung stabilisiert, in der die Regelspannung eine logarithmische Funktion des Druckes ist. Diese Spannung gibt eine logarithmische Druckskala, die mittels eines Röhrenvoltmeters, das ausserdem noch andere Aufgaben erfüllt, angezeigt wird. Ein Start-Trigger ermöglicht schnelle Auferregung der Plattenamplitude.

Versuchsmässige Ergebnisse liegen bisher nur für Luft vor, trotzdem der Messer auch erfolgreich für Helium (Gewicht 4) und Hexafluoropropylen (Gewicht 150) sowie Gase mit dazwischenliegendem Molekulargewicht eingesetzt wurde.

Kryostattemperaturmessungen von 0,1°K ... 20°K mit Wien-Brückenoszillator

von P. R. Adby

Zusammenfassung des Beitrages auf Seite 778-781

Die beschriebenen Oszillatoren sind eine Abwandlung des herkömmlichen Wien-Types, in dem die in den Widerständen des Phasenschiebernetzwerkes zerstreute Energie sehr klein ist. Einer der Widerstände ist temperaturabhängig und kann bei einer Temperatur von nur 0,1°K in einen Kryostaten eingeführt werden, ohne dass die thermischen Bedingungen merkbar beeinflusst werden. Die Oszillatorfrequenz wird mit der Temperatur in Zusammenhang gebracht und kann mittels eines Zählers gemessen werden.

Elektronische Kontrolle der Arbeitsweise von Funktionsverstärkern

von A. D. Bond und P. L. Neely

Zusammenfassung des Beitrages auf Seite 782-786

Der Beitrag bespricht den Entwurf eines Gerätes, das notwendig ist, um die bestehenden Analogrechnereinrichtungen so zu erweitern, dass sie die sich wiederholende und Kettenarbeitsweise der Operation einschliessen.

Die Grundlage des Entwurfes ist eine Sechs-Dioden-Brücke in einer Schaltanordnung für die Gegenkopplung.

Die Erzeugung von mit Dreieckpulsbreiten modulierten Wellen

von J. F. Young

Zusammenfassung des Beitrages auf Seite 787-789

Für die praktische Untersuchung impulsbreitenmodulierter Verstärkersysteme ist eine Quelle von Impulsen nützlich, deren Breite entweder linear oder durch Abwandlung des Dreiecks mit der Zeit geändert werden kann. Solch eine Quelle kann durch Bildung einer Schwebung von zwei Rechteckwellen mit etwas unterschiedlicher Frequenz erzeugt werden, wobei entweder UND- oder ein AUSSCHLIESSLICH-ODER-Gatter als Mixer Verwendung findet.

Ein kompensierter Transistor-Gleichstromkonstanthalter von M. Pacak

Zusammenfassung des Beitrages auf Seite 789-791

Eine beschriebene Konstanthalterschaltung ermöglicht hochwertige Stabilisierung von über einen breiten Bereich regelbaren Gleichströmen und Gleichspannungen mit $\pm 1{,}5 \times 10^{-4}$ Langzeitkonstanz nach acht Stunden und $\pm 2{,}5 \times 10^{-6}$ Kurzzeitkonstanz nach einigen Minuten. Die für die Hauptstöreffekte kompensierte Schaltung arbeitet mit Mitkopplung.

Verstärker mit regelbarer Verstärkung von S. Ghosh

Zusammenfassung des Beitrages auf Seite 792-793

Ein Verstärker wird beschrieben, dessen Verstärkung durch Änderung nur eines Widerstandswertes über einen Bereich von etwa 60 dB geregelt werden kann. Das wird ohne merkbare Änderung des Frequenzganges, der Ein- und Ausgangsimpedanzen und der Schleifenverstärkung des Verstärkers erreicht.

Entwurf eines Binär-Paralleladdierers von J. B. Earnshaw und P. M. Fenwick

Zusammenfassung des Beitrages auf Seite 794-796

Der Beitrag beschreibt den Grundentwurf eines Binär-Paralleladdierers mit sehr schneller Übertragfortpflanzung. Der Entwurf beruht auf einer zweistufigen Transistor-Dioden-Logikschaltung mit Rückkopplung, die eine Verzögerung von etwa 1 ns je logische Stufe hat. Anfängliche Messungen lassen darauf schliessen, dass die Summenbildungszeit für Worte mit 16 Bit etwa 50 ns beträgt.

Optimierung eines Impulstreiberentwurfs für schlechteste Betriebsbedingungen von P. F. Jones

Zusammenfassung des Beitrages auf Seite 797-799

Ein Optimierungsverfahren für schlechteste Betriebsbedingungen berücksichtigt im Entwurfsstadium die Gesamtdrift von Bauelementen während ihrer Lebensdauer. Nach dem Verfahren werden die Grenzwerte für jedes Bauelement direkt von den Grenzwerten der Parameter der Anfangsleistungsfähigkeit berechnet. Bauelemente mit bevorzugten Nennwerten, die innerhalb der berechneten Grenzwerte liegen, werden dann gewählt. Das Verfahren formuliert einen Satz von Ungleichheiten von den Leistungsparametern und löst diese gleichzeitig, wodurch sich ein zweiter Satz ergibt, der die Grenzwerte für jedes einzelne Bauelement bestimmt.

Binär-quinärer Dekadenzähler mit Widerstandslogik von R. Parshad und S. P. Suri

Zusammenfassung des Beitrages auf Seite 800-801

Dieser Beitrag beschreibt einen binär-quinären Dekadenzähler mit Widerstandslogik. Die besprochene Schaltung benötigt weniger Transistoren als konventionelle. Die Digitalanzeige ist einfach und sparsam in der Anwendung von Bauelementen. Mit der Schaltung kann auch ohne Schwierigkeiten Zählen in beiden Richtungen erreicht werden.

Ein mit einem neuen Feldeffekttransistor bestückter Chopper-Verstärker mit sehr hoher Eingangsimpedanz von R. Verrill

Zusammenfassung des Beitrages auf Seite 802-804

Der neuentwickelte Feldeffekttransistor EXP380 hat einen sehr niedrigen Gate-Reststrom in der Grössenordnung von 10^{-10}A bei 25°C und eine sehr niedrige Gate-Kapazität von etwa 2, 5 pF. Das Element hat sich in einem versuchsmässigen Serienparallel-Chopper-Gleichstromverstärker, für den niedriger Gate-Reststrom und niedrige Kapazität erforderlich sind, als sehr erfolgreich erwiesen. Der beschriebene Verstärker hat einen Eingangswiderstand von 100 MΩ, und die Empfindlichkeit für Vollausschlag bei ±100 μA am im Ausgang liegenden Drehspulinstrument war ±1 mV. Die Eingangsdriftverschiebung mit Temperatur entsprach etwa 1 μV/°C und 15 pA/°C, ohne dass Zusammenpassen von Transistoren erforderlich war.

Digitaldarstellung der Lufttemperatur zwischen —17,8°C und +37,2°C von W. V. Dromgoole

Zusammenfassung des Beitrages auf Seite 805-807

Dieses Gerät wurde entwickelt, um die Lufttemperatur in eine Digitalform umzusetzen, die Anzeige in Zehnern und Einheiten auf einem Lampenanzeigesystem gestattet.

Die Lampenanzeige ist Teil eines Systems, das jede Minute 47 Sekunden lang die Zeit und 10 Sekunden lang die Lufttemperatur anzeigt und in Städten, in denen das System installiert ist, auf weite Entfernung sichtbar ist.

Ein Transistor-RC-Oszillator mit negativen Impedanzen von S. Pasupathy

Zusammenfassung des Beitrages auf Seite 808-809

In diesem Beitrag wird eine verallgemeinerte Form von RC-Oszillatoren mit negativer Impedanz beschrieben und gezeigt, dass Wien-Brückenschaltungen einen Sonderfall derselben darstellen. Eine direkte Synthese dieses Netzwerkes mit einem negativen Impedanzwandler ergibt einen neuartigen Impedanzoszillator. Die Oszillatorschaltung, ihre zwei Hauptarbeitsweisen und einige Sondereigenschaften werden besprochen.

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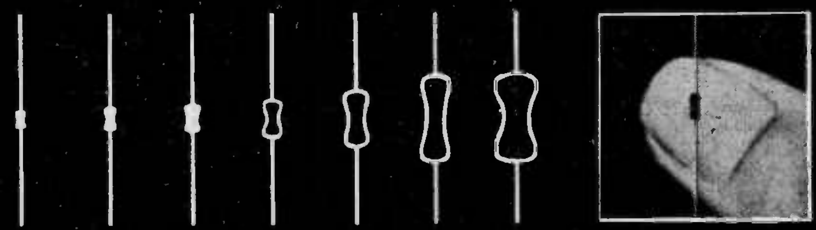

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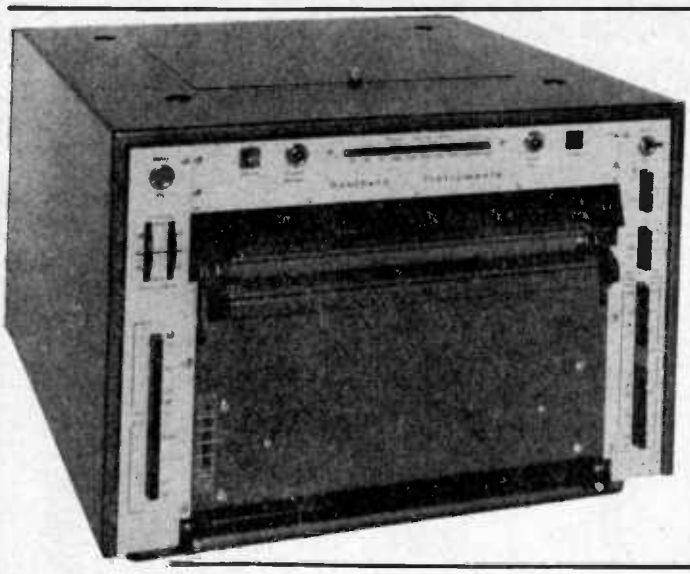

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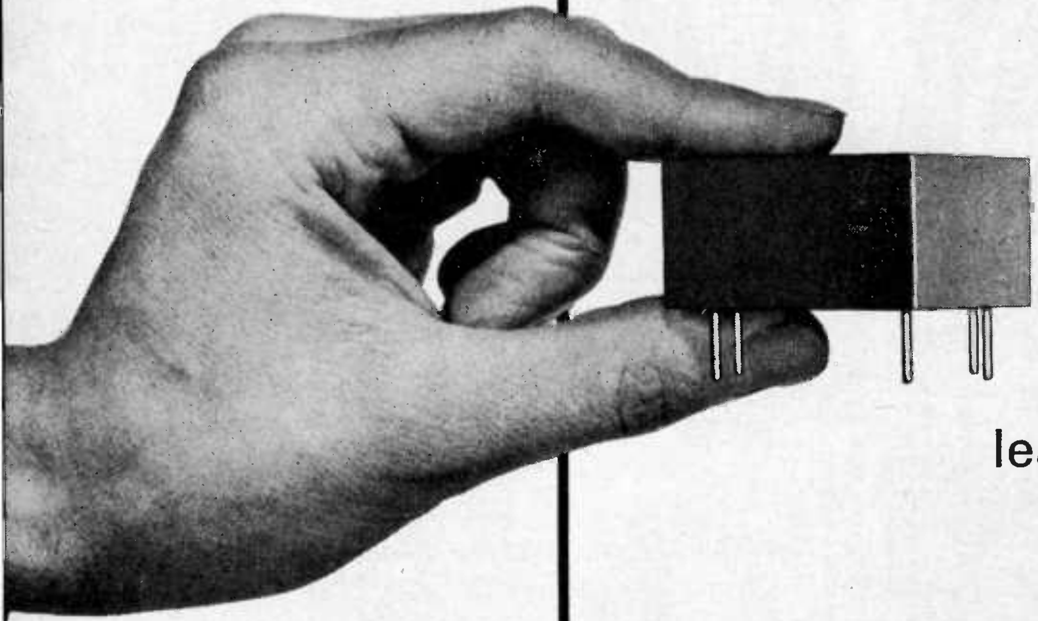

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REX 1S	30–50V	1A	3⅛"		7⅜"	£37. 15.
REX 2.5S	30-50V	2.5A	4⅛"		12⅜"	£55. 15.
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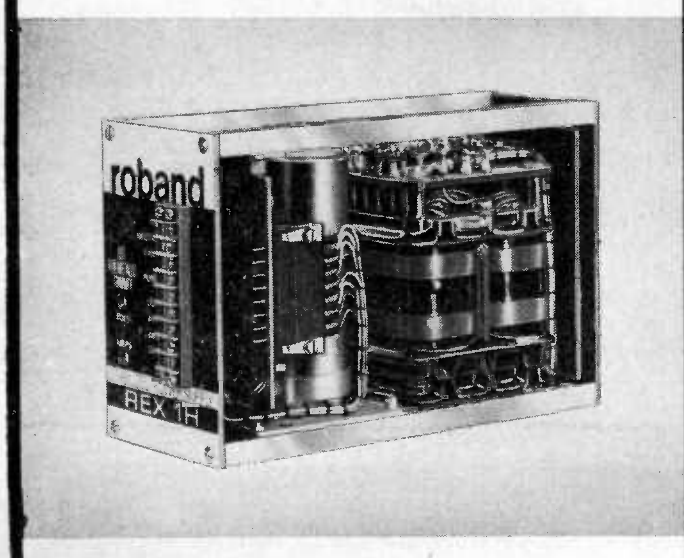
PERFORMANCE Input Voltage: 100-125/200-250 ± 7½%. 48-400 cps. Stability (15% mains variation): 0.01%. Stability (0 to full load): 0.05%. Ambient Temperature: —10°C to +60°C. Temperature Coefficient: 0.01%°C. Protection: Self-resetting current limiter preset to operate at 110% of maximum current rating. Ripple and noise: 500 microvolts, peak to peak.



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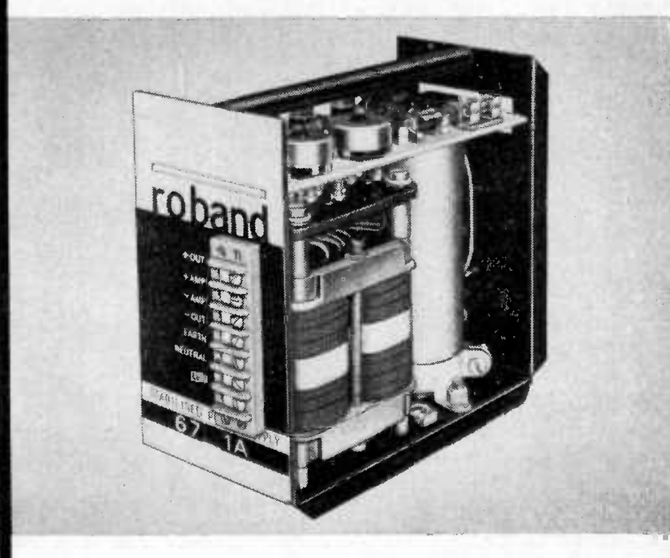
67–Twin 5A	6 –24V (X2)	5A–5A	5½"	6 21/32"	12"	£76

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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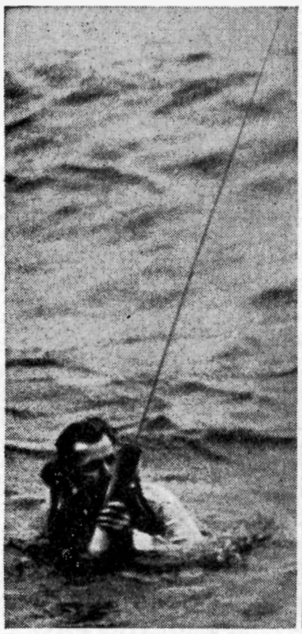

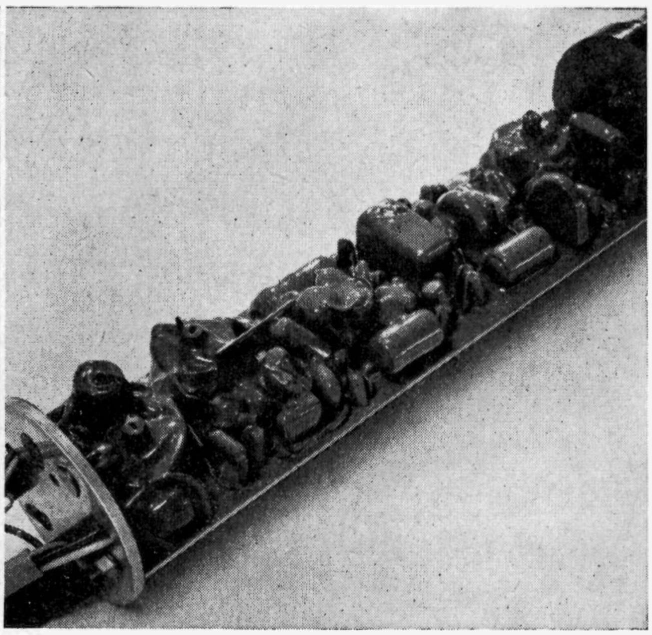

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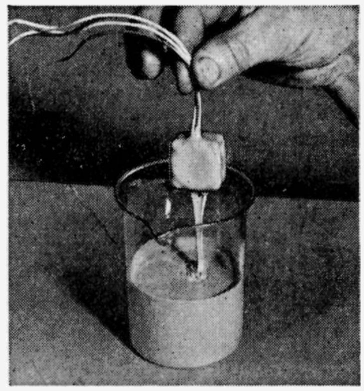
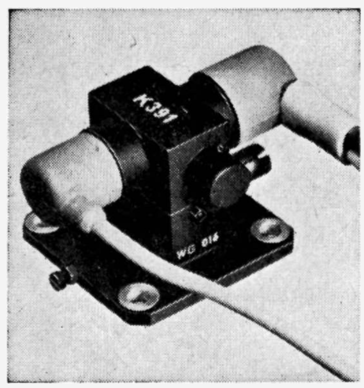
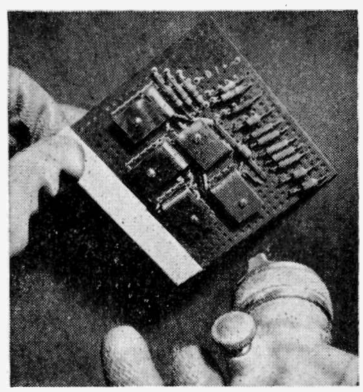


Dip-encapsulating a transformer in 'Silcoset'. The coating quickly cures to a resilient silicone rubber.

Miniature cable assembly: connector/cable junctions are potted in 'Silcoset' to protect from moisture, shock and vibration. Courtesy Amphenol-Borg (Electronics) Ltd., Northern Division.



'Silcoset' encapsulation provides a good hermetic seal for this klystron. The seal does not become brittle even at very low temperatures. Courtesy English Electric Valve Company Ltd.



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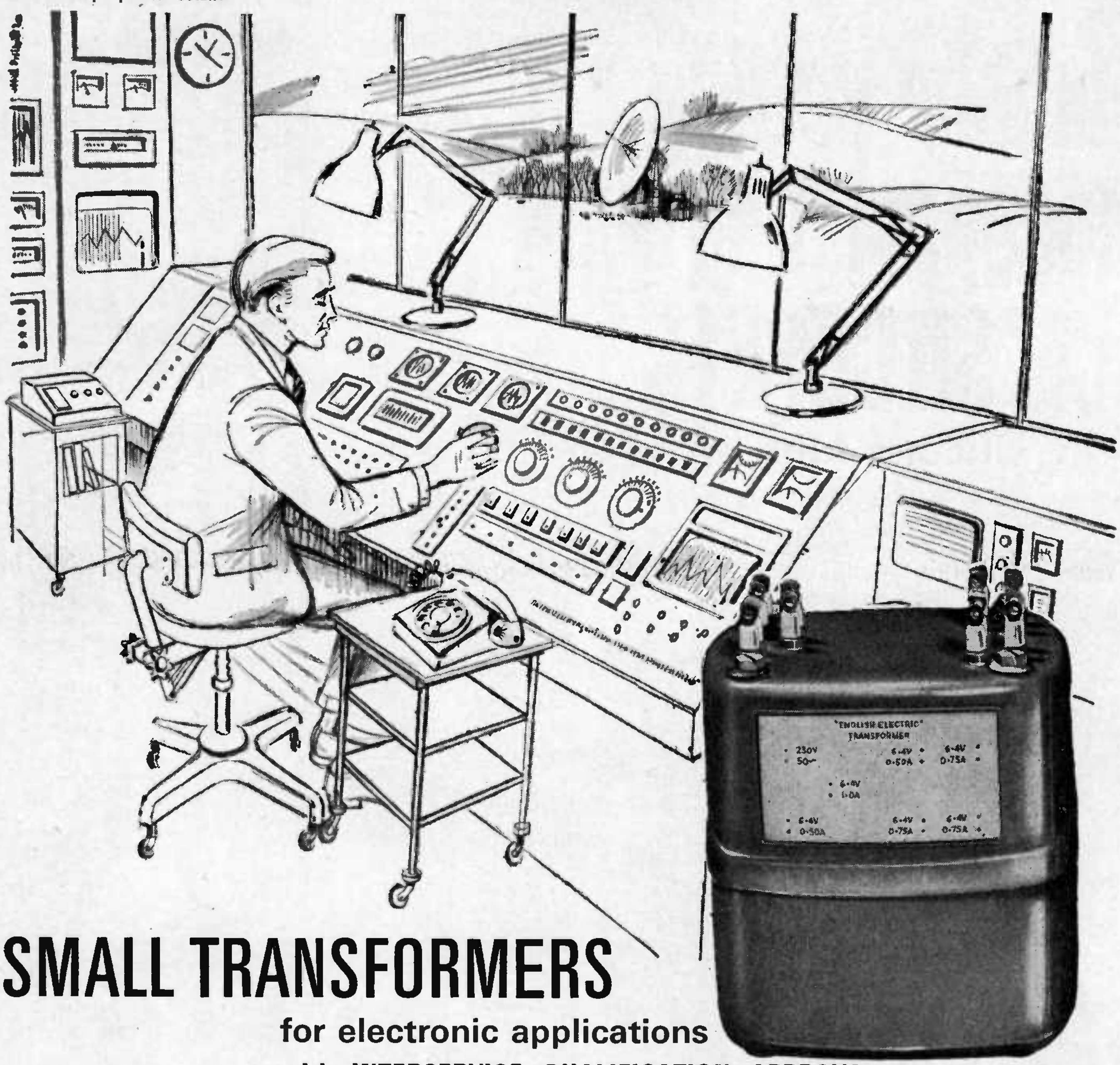

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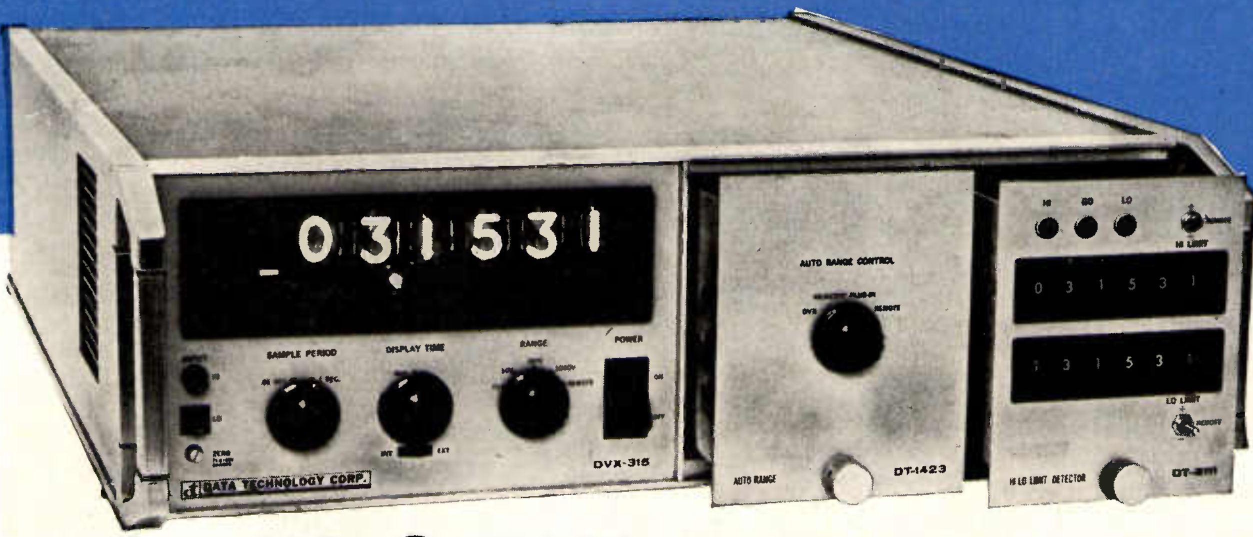

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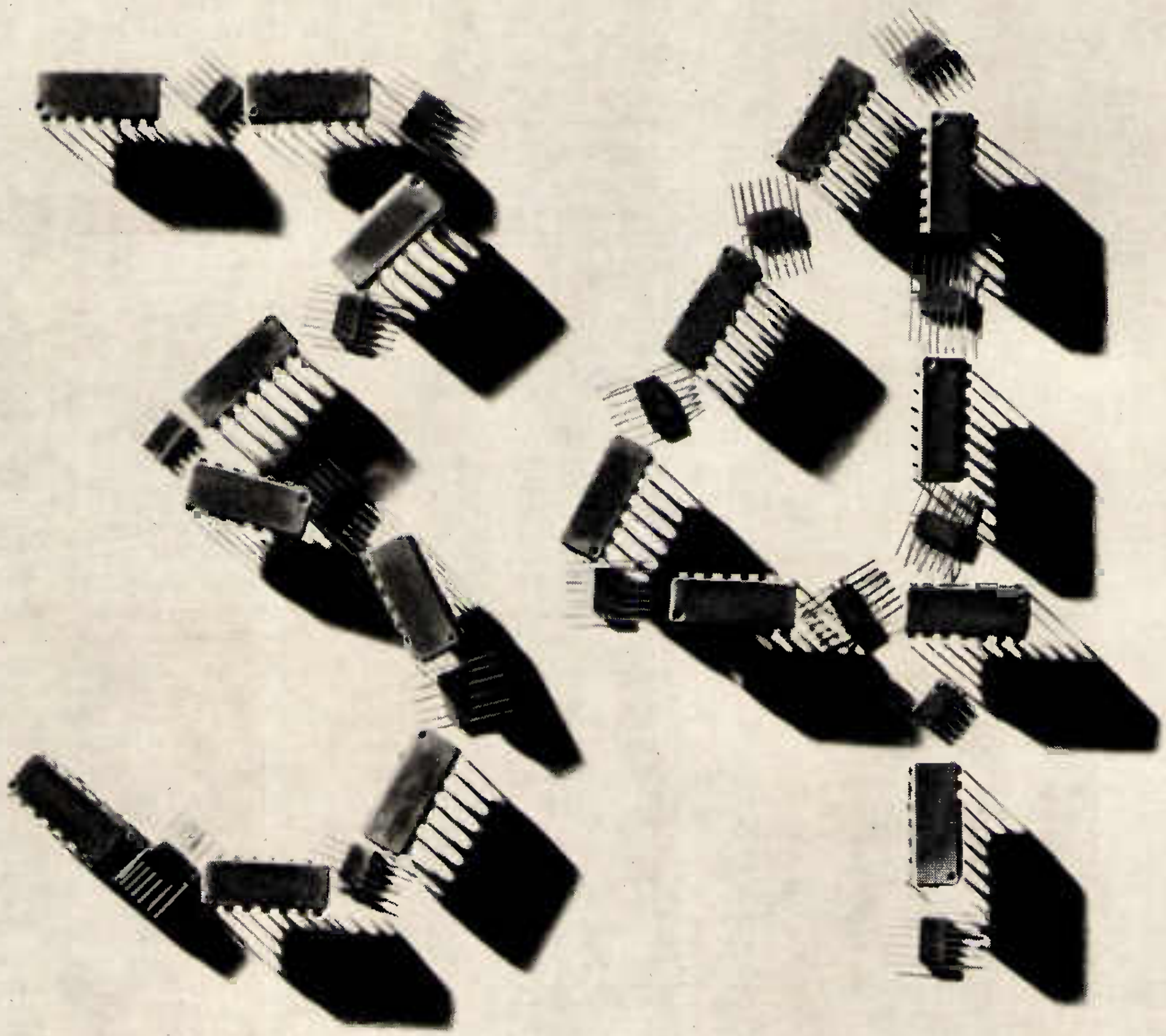
Output Z: Less than 200 m Ohms, at 200 Kc/s.
Ambient Temperature: 55°C.



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FCH141	Triple Nand/Nor gate	FCH142
FCH151	Triple Nand/Nor gate	FCH152
FCH161	Triple Nand/Nor gate	FCH162
FCH171	Triple Nand/Nor gate	FCH172
FCH181	Quadruple Nand/Nor gate	FCH182
FCH191	Quadruple Nand/Nor gate	FCH192
FCH201	Sextuple Nand/Nor gate	FCH202
FCH211	Sextuple Nand/Nor gate	FCH212
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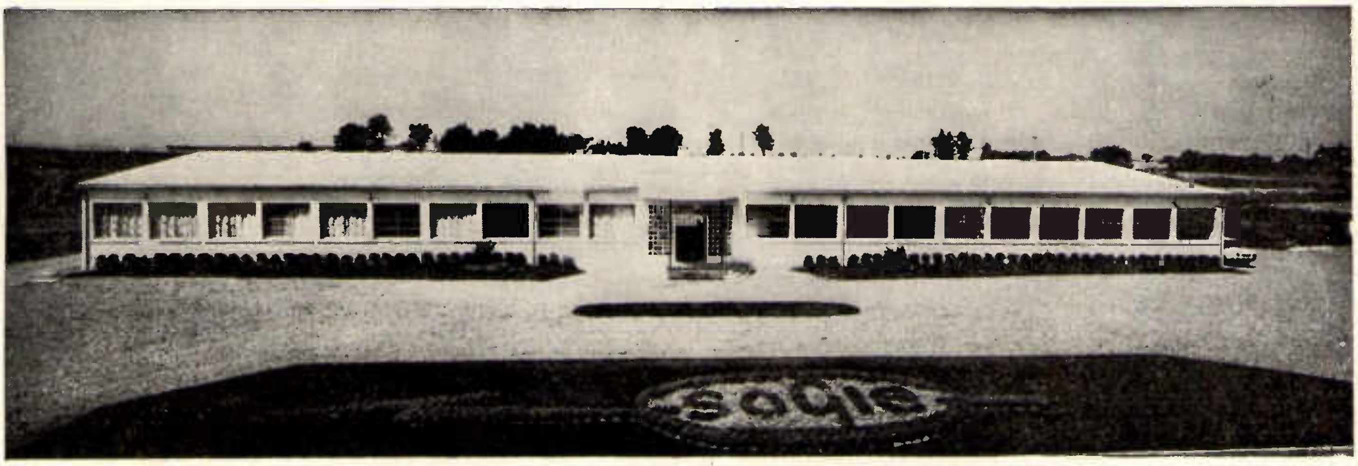

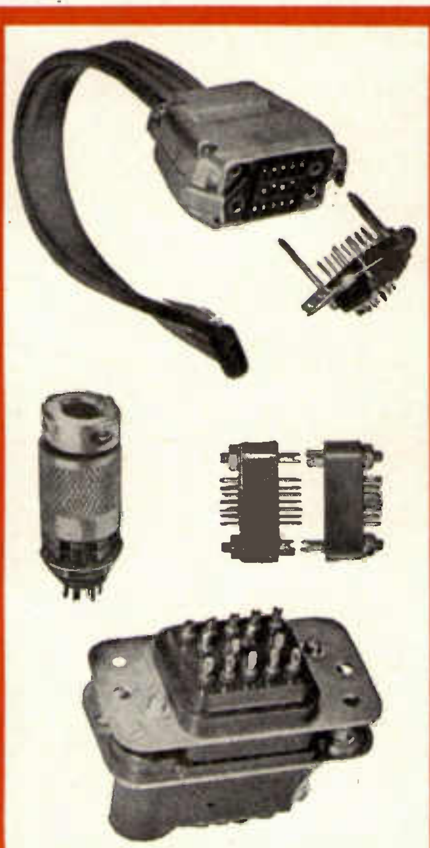

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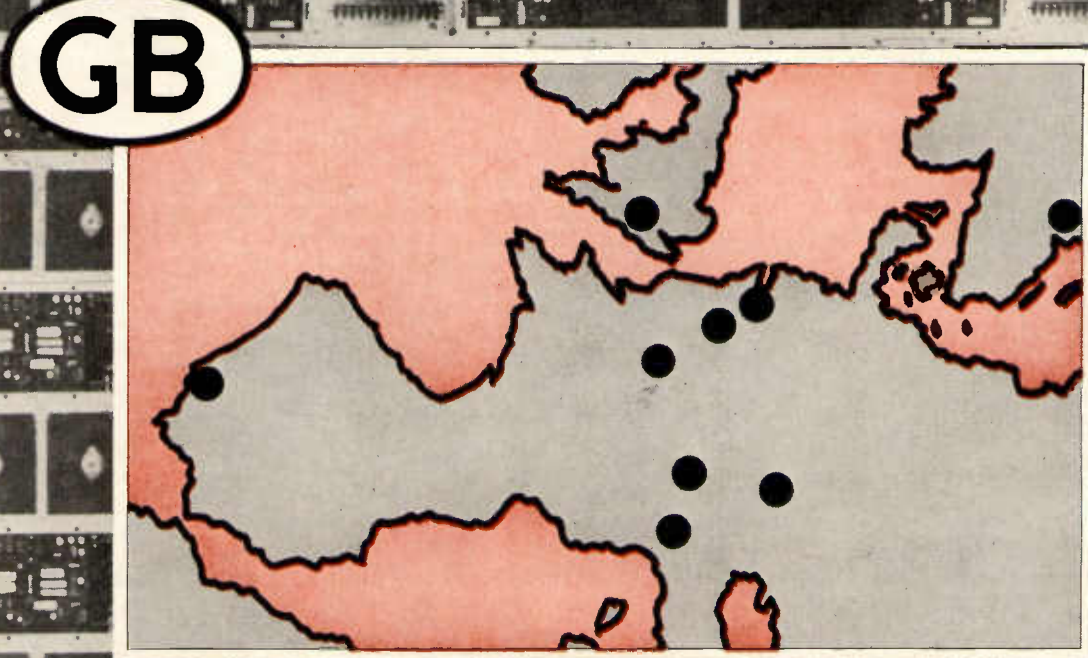

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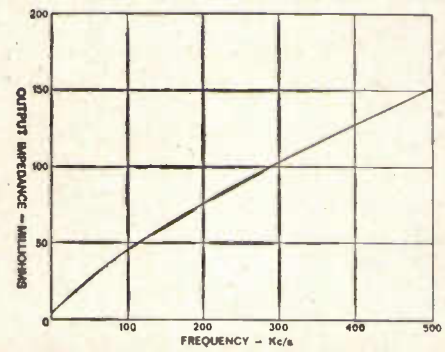
SWEDEN: ab Elektronik Enheter, Torogatan 24, Enskede
SWITZERLAND: Omni Ray, A.G, Dufourstrasse 56, 8008 Zurich
SOUTH AUSTRALIA: W. E. Beveridge & Co., 204 Cross Road, Unley Park, Adelaide
U.S.A.: Dentronics Inc., 60 Oak Street, Hackensack, New Jersey 07601

selling POWER SUPPLIES

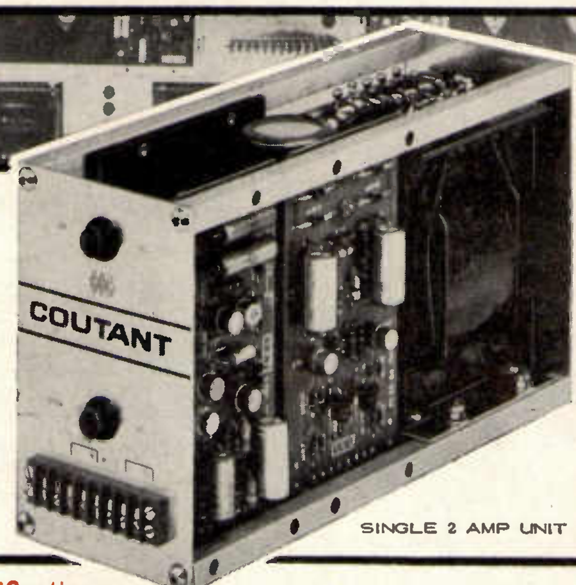
COUTANT FOR PEAK PERFORMANCE

Coutant Electronics' modular power supplies are built with precision, embody the most advanced design techniques and offer you unequalled reliability even when operated continuously at elevated temperatures. A typical specification is given below—and these figures are often bettered substantially in production. For VA per pound weight *and* VA per cu. in. COUTANT specs lead the field.



COUTANT for the widest choice

Manufacturing more than 70 different models of power supply unit COUTANT offer the widest and highest performance range in Europe. The 'E' Series offers from ½-amp to 2×5A in 19 versions—7 of which cater for 5-14V output needs. There are 23 standard modules in the high-temperature rated 'K' series—and there are unstabilised versions too. Whatever your application area, there is a COUTANT supply unit to give you best value for money. NOW—VARIABLE UNITS TOO! Recent additions to our range include variable lab. supply units in bench or modular form: current and voltage variable—sub-units remotely programmable and ideal for systems work. FILL IN THE COUPON NOW!



SINGLE 2 AMP UNIT

typical basic specification

STAB. RATIO	O/P RESISTANCE	RIPPLE p-p	O/P IMPEDANCE 100 kc/s
> 5000 : 1	≤ 0.002 Ω (< 0.0005 Ω ABOVE 5A)	< 200 μV	< 0.1 Ω

C E C

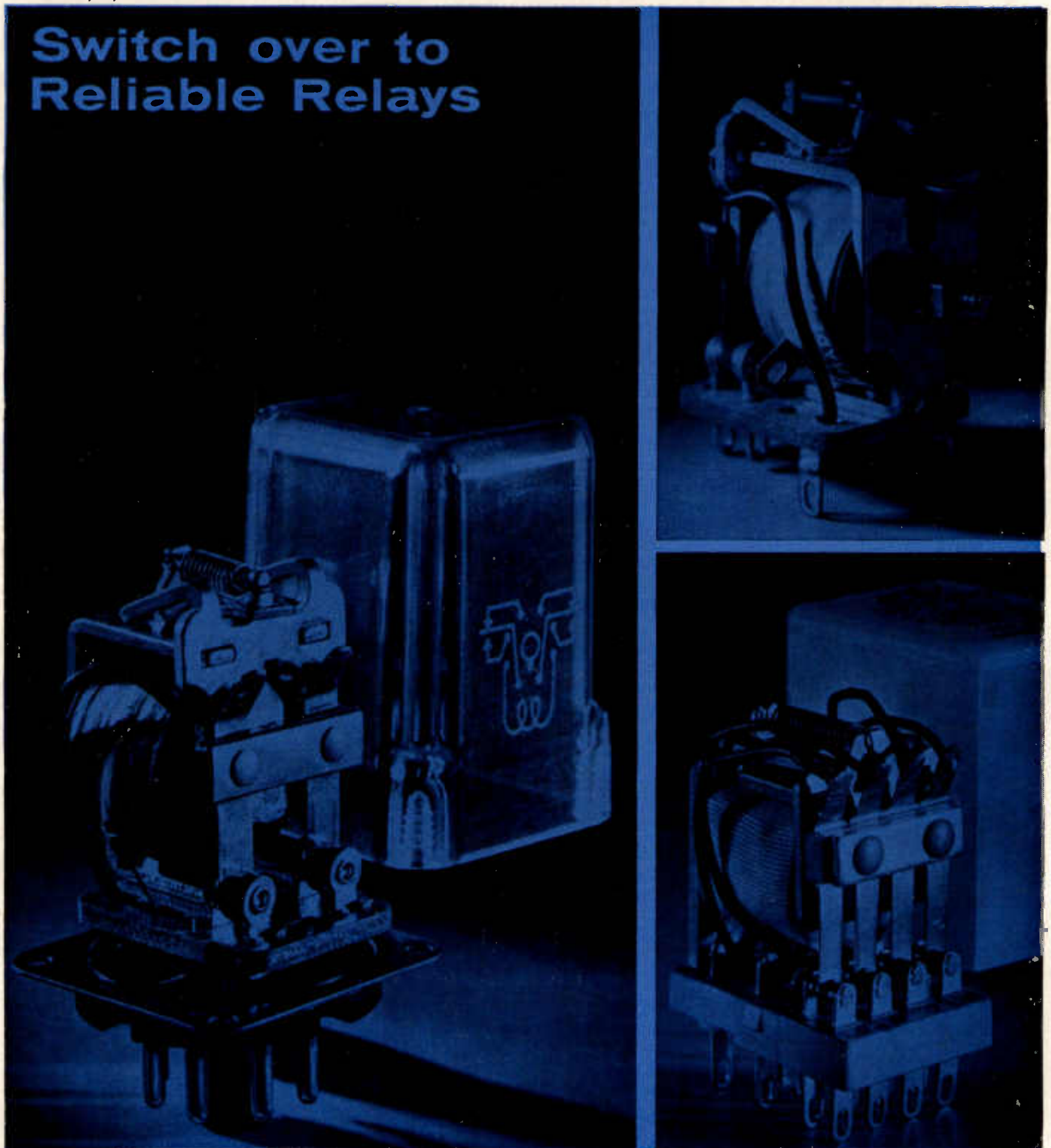
COUTANT ELECTRONICS LIMITED

TRAFFORD ROAD
RICHFIELD INDUSTRIAL ESTATE · READING · BERKSHIRE
Tel: READING 55391-5

electrotech instruments division

EE 01 115 *for further details*

EE 01 116 *for further details*



Radiospares for long-life relays! They're made to the highest standards of quality throughout and all types are available "by return of post".

Miniature plug-in relays—12V. or 24V. D.C. coils, 115V. A.C. coils.
4 P.C.O. contacts rated 3A. A.C.
Open series—12V. D.C. or 230V. A.C. coils 2 P.C.O. contacts rated 250V. 5A. A.C. Also 6V. D.C. coils with 115V. 2A. A.C. contacts.
Plug-in relays—D.C. and A.C. coils with 2 or 3 P.C.O. contacts rated at 2A. or 5A. or 10A. 250V. A.C.

Radiospares Ltd.

RS

4-8 MAPLE STREET · LONDON W.1 · ENGLAND · TELEPHONE: EUSton 7232 (8 LINES)
TELEGRAMS and CABLES: RADOSPERES, LONDON W.1

RS 9/1

You know - but do you PLESSEY

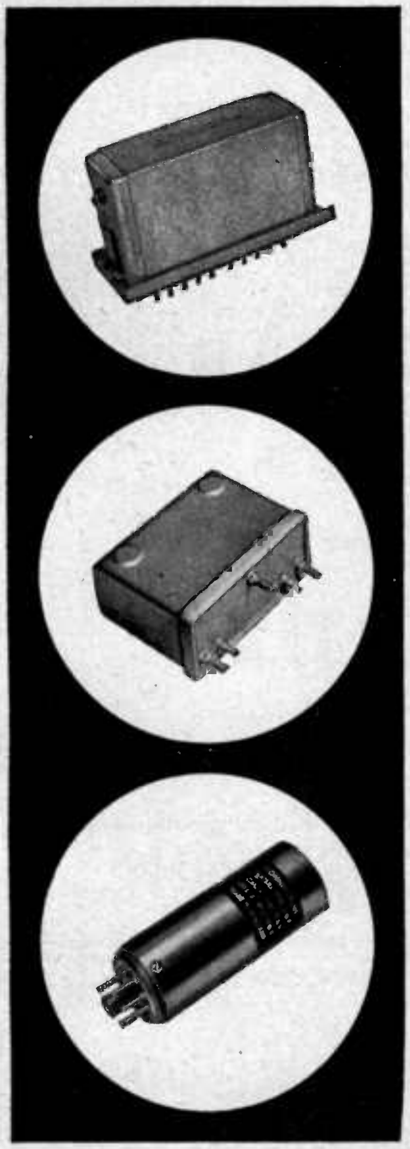
Who we are

The Plessey Electronics Group is one of the five Product Groups of The Plessey Company Limited. The Group functions as an independent and fully decentralised unit with its own research, development, production and marketing organisations.
Plessey Electronics employs 7,500 people in 11 different centres with a total floor space of 1.3 million square feet. One person in every four is engaged in research and development in one of the separate research establishments or in laboratories associated with production units.
The outstanding features of Plessey Electronics are the breadth of its product range and the depth of its system capability. Within the Divisions that constitute the Group is a wealth of experience in the development, production, planning and installation of major electronic projects of international importance.

What we do

The activities of Plessey Electronics range over the design, development and production of electronic equipment in the areas of radar, radio communications, carrier transmission, avionics, oceanography and national defence.
It is the major British manufacturer of HF, VHF and UHF communications systems. Its line-integrated radio telephone systems bring telephone service to isolated communities. It manufactures air surveillance, coastal and meteorological radars and complete air traffic control and air defence systems. Advanced programmes of work are in hand on underwater weapons systems and in the vital and expanding science of oceanography.
To complete the total system capability in communications, the Electronics Group is engaged in forward looking programmes of microwave and space communication development.

Plessey Electronics...for men



QUARTZ STABILITY

crystals

Marconi have more than twenty years' experience in the manufacture of high quality quartz crystals.
A wide range of crystals can be supplied from 1 kc/s upwards.
Special low aging types are available at 5 Mc/s and 2.5 Mc/s for use in high stability oscillators.

crystal filters

A selection of crystal filters, both s.s.b and bandpass, are available.
New designs are constantly being added to the range and can be produced to meet special requirements.

crystal ovens

Marconi crystal ovens accommodate various types of crystals.
Models are available which can accommodate up to ten crystals.
Outstanding in the range is the Marconi 'change of state' oven which needs no thermostat and has a switching differential of $\pm 0.0014°C$.

Marconi specialized components are designed and manufactured only when the precision and high performance required is otherwise unobtainable.
The Specialized Components catalogue lists the full range.

Marconi specialized components

The Marconi Company Limited, Specialized Components Division,
Billericay Works, Radford Crescent, Billericay, Essex, England. Telephone: Billericay 3431

LTD/F54

EE 01 125 *for further details*

ARE YOUR INSTRUMENTS ACCURATELY CALIBRATED?

The Bradley Calibrator 130B provides D.C. voltages up to 3,000 and A.C. voltages up to 511 at an accuracy of 0.2% for the rapid and easy calibration of all types of instruments.

This Calibrator provides a wide range of A.C. and D.C. precision voltage steps for testing of avometers, meter movements, transistor and valve voltmeters, oscilloscopes etc. All outputs have % error indication and are short circuit proof.

To enable you to test the 130B Calibrator yourself, we shall be glad to provide one on loan. In addition, the Bradley range includes precision current sources and programming facilities. May we send you full details of the Bradley range of calibrators? Write for the 'Calibration Equipment' booklet, publication No. 102.

See your Electronic Engineering Index Library for full technical data

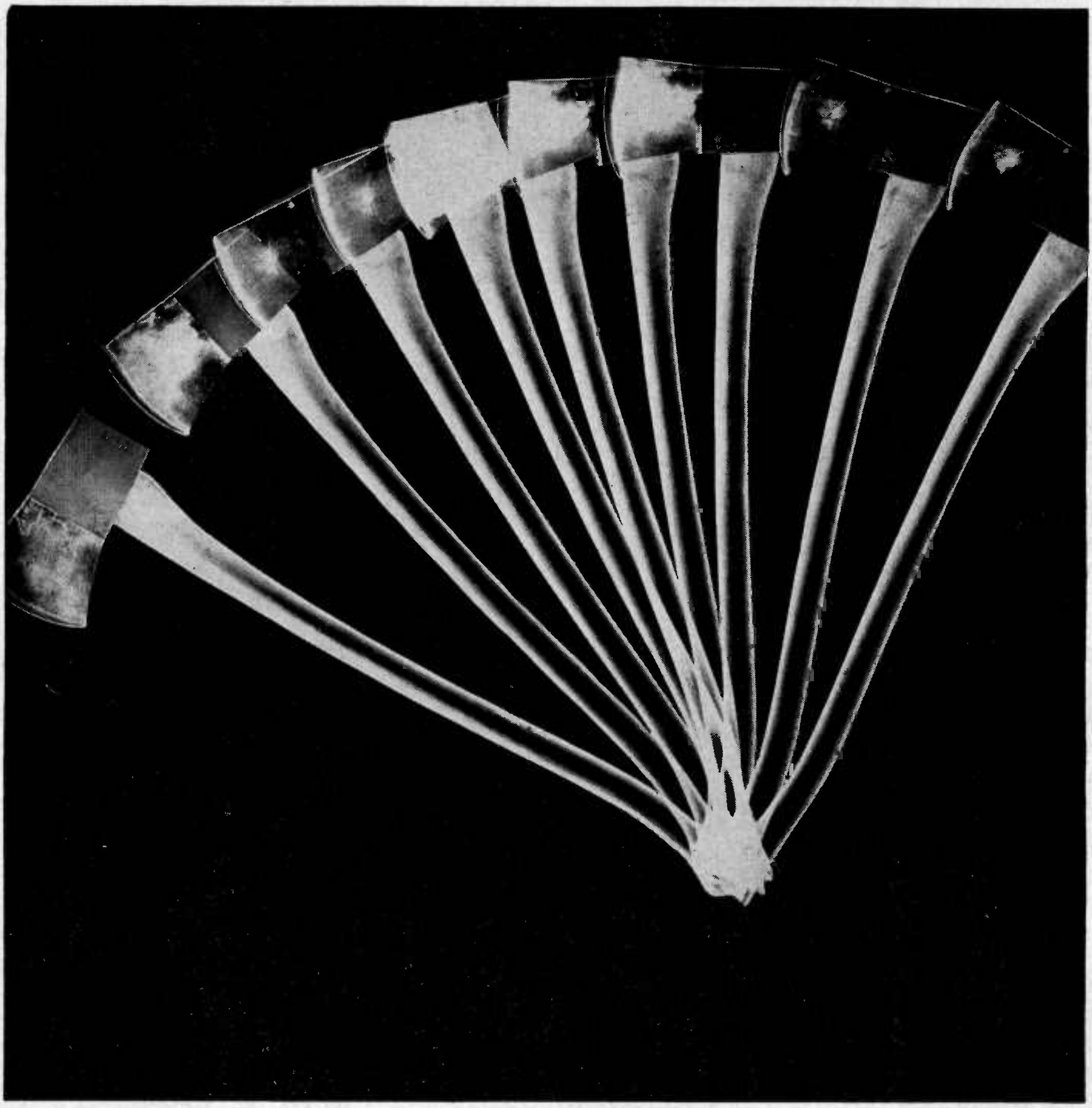
GEB

G. & E. BRADLEY LIMITED

Electral House, Neasden Lane, London, N.W.10. Telephone: DOLlis Hill 7811

Telegrams: Bradelec London N.W.10. Telex: 25583

EE 01 126 for further details



15 AMPS IN ONLY 7¾″ DIA

0-275 VOLTS

MODEL 74 REGAVOLT

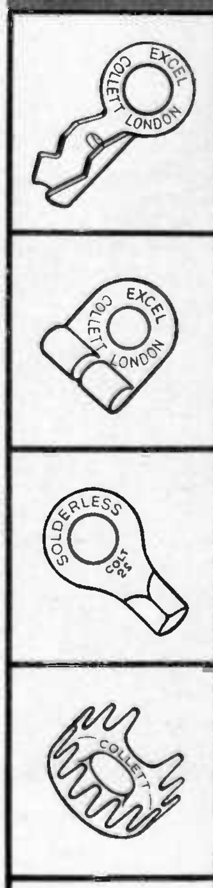
variable transformer with trouble-free brush gear

Another winner for BERCO DESIGNERS... Model 74 delivers twice as much KVA as any other variable transformer of equivalent size... only £25.15.0... delivery from stock. Please phone Mr. Love, or write for list 625H/S for full details.

BERCO
group of companies

THE BRITISH ELECTRIC RESISTANCE CO LIMITED
Queensway · Enfield · Middx · Tel: HOW 2411
Grams: Vitrohm Enfield

46B



SMALL QUANTITIES OFF-THE-SHELF

Over 200 different types of SOLDER and SOLDERLESS TERMINALS, TAGS and CLIPS are available in the COLLETT range.

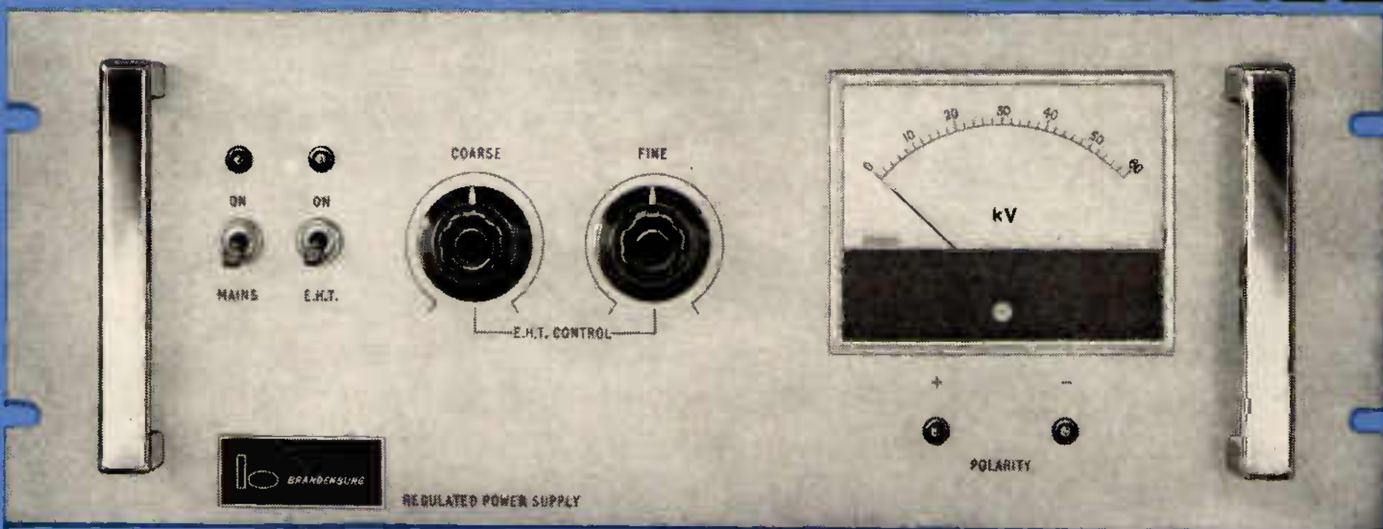
COLLETT specialise in supplying SMALL QUANTITIES. Boxes of 100 only—available of any type. The complete range is always in stock and can be supplied immediately OFF-THE-SHELF.

Send for fully illustrated catalogue and price list today.

COLLETT TERMINALS
347/349, GOSWELL ROAD,
LONDON, E.C.1.
TERMINUS 2584

EE 01 129 *for further details*

60kV 7" HIGH



THE NEW BETA RANGE MODELS 905 & 906

Extending the recently introduced BETA Range of high voltage stabilised power supplies, the new Models 905 & 906 provide 400 µA at 6-60kV — and still within a 19 in. panel, 7 in. high, just 17 in. deep.

Despite their compact proportions, the Models 905 & 906 compromise with neither reliability nor specification; this outstanding example of size reduction depends largely on careful attention to detail in high voltage head design and the manufacture of specially moulded and encapsulated components—an area in which the Brandenburg design team leads the world.

Brief Specification:

Input	200-250 volts 50/60 c/s or 100-125 volts to special order.
Output	Model 905: 6kV to 60kV at 400µA positive. Model 906: 6kV to 60kV at 400µA negative.
Ripple	100V peak to peak.
Regulation	0.25% at full voltage.
Stability	0.1% against +10% and −7½% mains change.
Polarity	Either positive or negative (not reversible).
Price	£245.

and models 705 and 800

(identical in size)

Brief Specification:

Output	Model 705: 2mA at 1.5kV to 15kV. Model 800: 1mA at 3kV to 30kV.
Ripple	0.1%.
Regulation	0.25% at full voltage.
Stability	0.1%, against +10% and −7½% mains change.
Polarity	Reversible.
Price	£150.

brandenburg limited

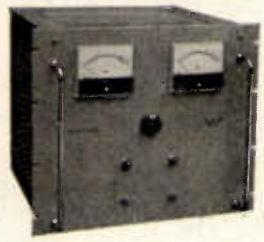
139 Sanderstead Road, South Croydon, Surrey. Tel: Sanderstead 0225.

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

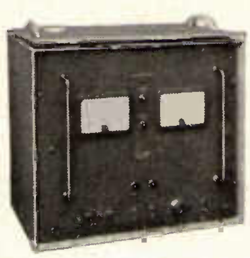
Output	100-2,500 volts d.c.
Output Current	5mA d.c. maximum.
Polarity	Reversible.
Ripple	Less than 10mV.
Stability ±7% Mains Change	Not worse than 0.03%.
Drift with Time	Less than 0.01% per hour.
Voltage Control	5 Switched Ranges with overlapping fine control.
Input	200/250 volts 50 c/s.
Size	19″ wide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

EE 01 140 *for further details*

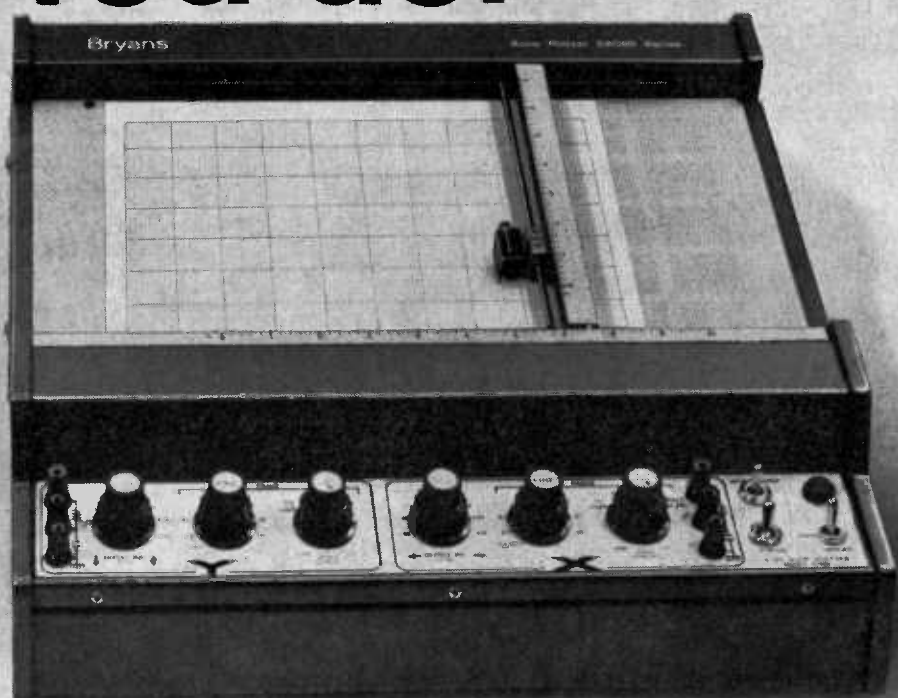
Who needs 1200 in/sec^2?

You do.

Bryans 22000 series, the world's highest performance plotters in full production, offer 45 in/sec. writing speed and 3g maximum acceleration. The pen follows your rapidly varying and transient inputs more accurately than any other.
The 22000 series is simple to operate –no manual gain control, no external unit needed for AC conversion— and is fitted with a well designed rear plug-in panel for plotter accessories and computer readout connections.

Let us demonstrate the 22000 under your conditions.

Bryans Ltd., Willow Lane, Mitcham, Surrey. Mitcham 5134.



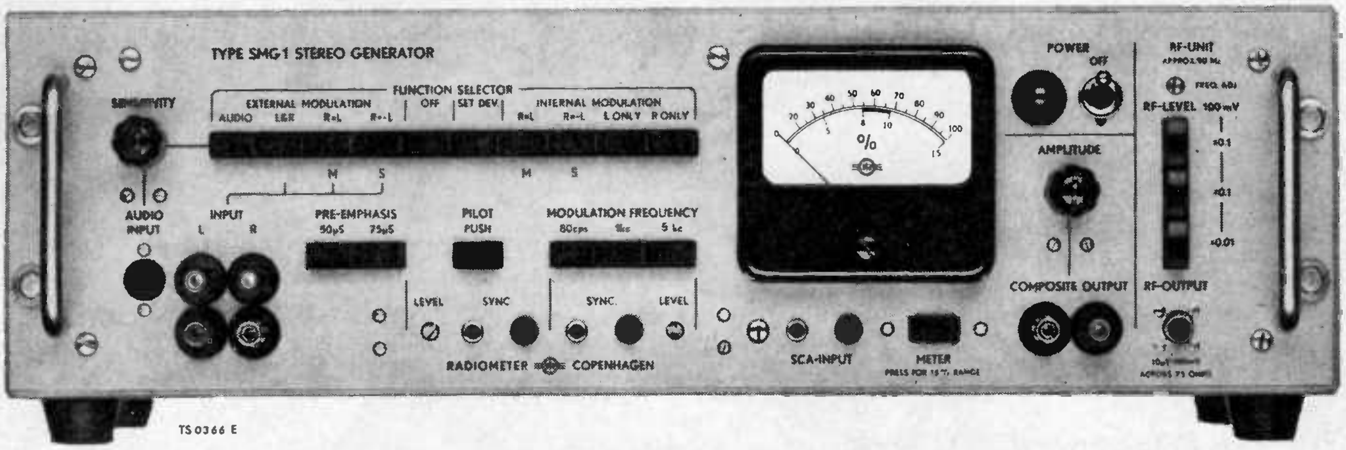
Bryans

22000 plotters

BY36

EE 01 143 *for further details*

STEREO GENERATOR TYPE SMG1



The type SMG1 Stereo Generator is a transistorized signal source that complies with the standards for stereophonic broadcasting approved by the American FCC and recommended by the European Broadcasting Union (EBU).

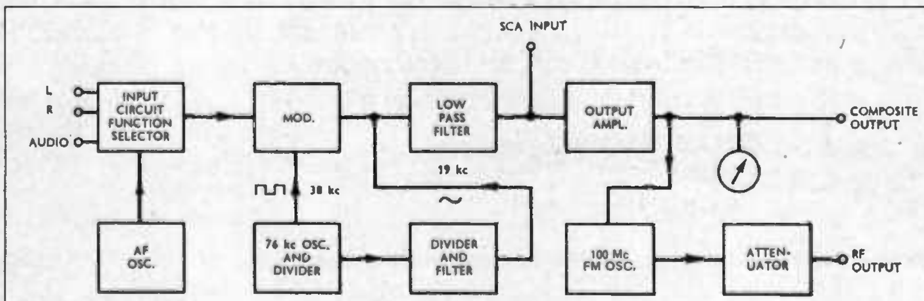
The Stereo Generator operates according to the time-multiplex system (PAM – System), and gives the following output signals: A composite signal for testing stereo adapters or driving FM generators to full 75 kc deviation, and a 100Mc output signal which is frequency-modulated by the composite signal. In connection with a built-in step attenuator in the 100Mc output, the Stereo Generator is self-contained for testing stereo receivers and stereo tuners. The modulation can be generated either by means of a built-in oscillator with 3 fixed AF frequencies, or by means of an external generator. Two input circuits are available for connecting the external modulation signal source, namely one suited for connection to a stereo record-player or tape recorder, and one for connection to separate R and L signals. Provision is made for modulating with pure M (L = R), (monophonic channel), pure S (L = –R), (stereophonic sub-channel), L-only and R-only.

The Stereo Generator, type SMG1, features:

▶ LR and MS separation $>$40 dB. ▶ Distortion $<$ 0.2%. ▶ Extended frequency range (54 to 210 Mc), when used with the Radiometer FM-AM Standard Signal Generator, type MS 26. ▶ 10 μV to 100 mV RF output. ▶ Push-button operation.

For Great Britain:

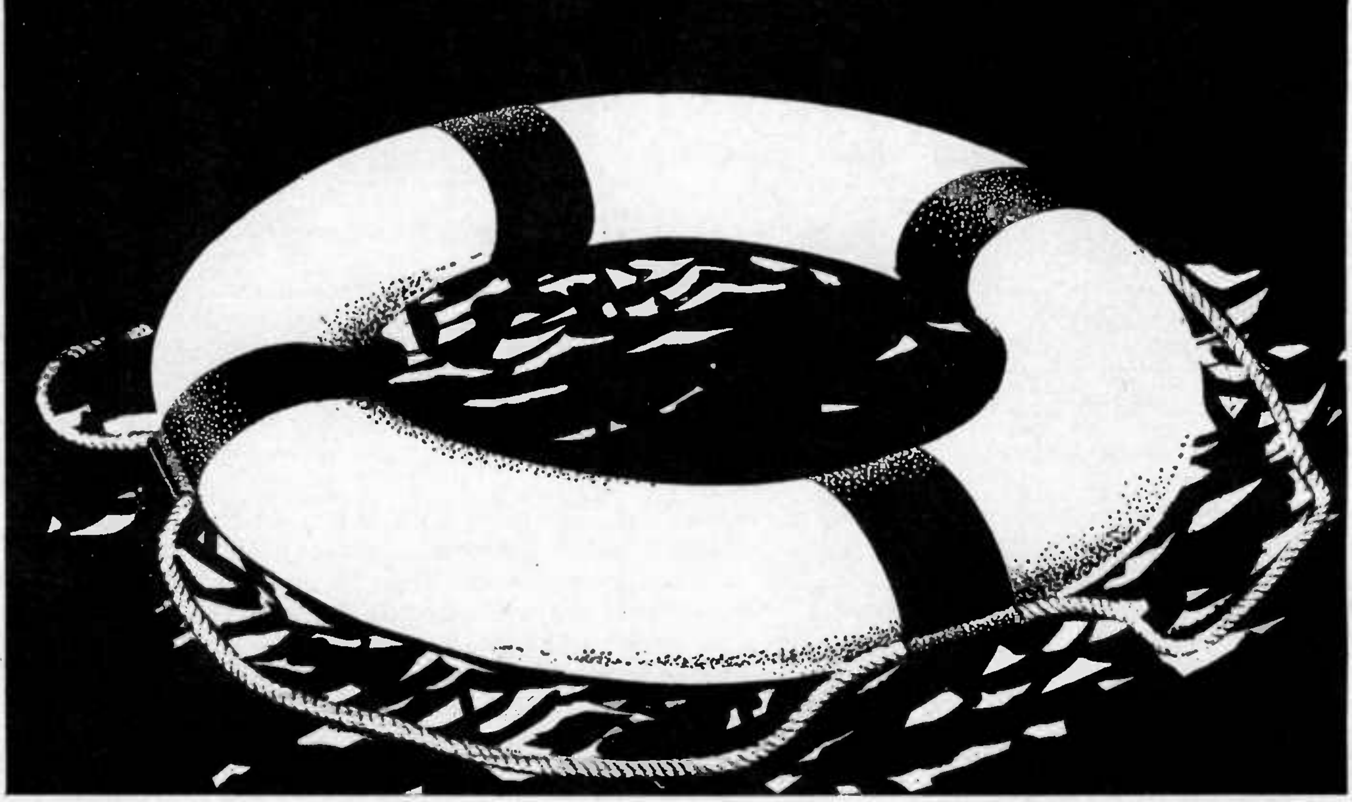
Livingston Laboratories LTD.
Livingston House
Greycaines Road
North Watford
Herts



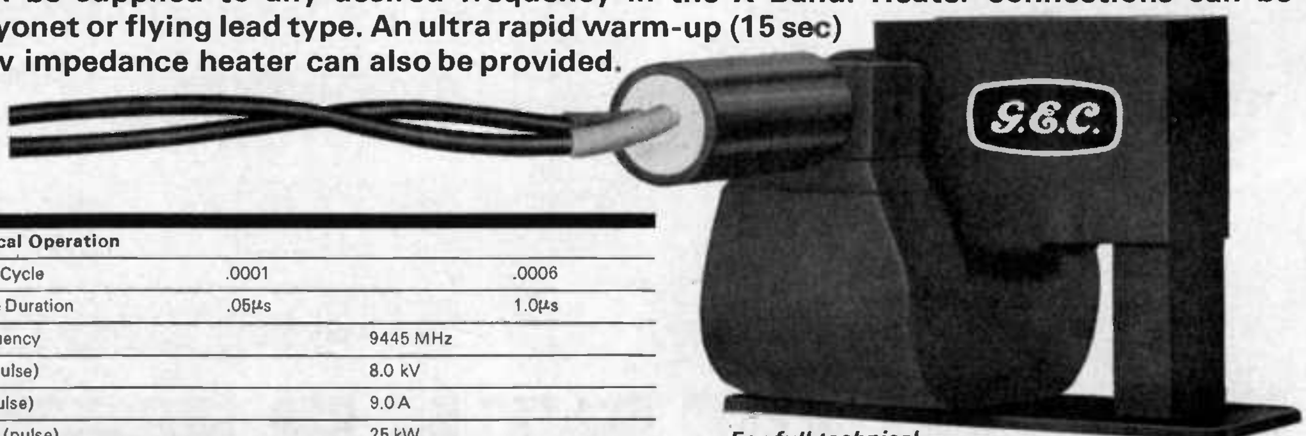
RADIOMETER A/S, 72 EMDRUPVEJ, COPENHAGEN NV, DENMARK

RADIOMETER COPENHAGEN

If there's a next time he'll know better



In fogbound, densely packed harbour approaches, reliable high-performance Radar is vital. But a ship's Radar set is only as good as its Magnetron. If it becomes unserviceable, The Captain's next order could be "All hands to rig collision mats" – (after keel-hauling the radar man.) Next time, if there is a next time, he'll know better.
With M-O V's new MAG16, reliability comes first. It has a predicted reliability in excess of 5,000 hours – (2,000 is guaranteed).
The MAG16 is a compact, Pulsed Magnetron, designed for Marine Radar applications. It can be supplied to any desired frequency in the X-Band. Heater connections can be bayonet or flying lead type. An ultra rapid warm-up (15 sec) low impedance heater can also be provided.



Typical Operation			
Duty Cycle	.0001		.0006
Pulse Duration	.05μs		1.0μs
Frequency		9445 MHz	
V_a (pulse)		8.0 kV	
I_a (pulse)		9.0 A	
P out (pulse)		25 kW	

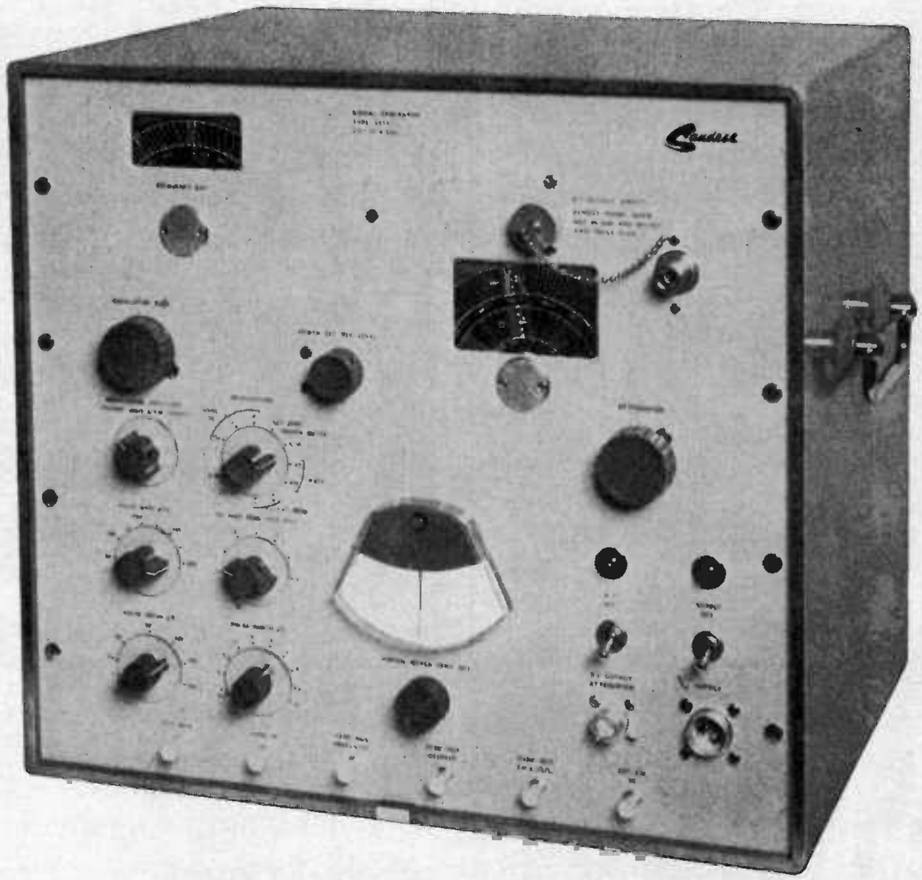
For full technical specification write or phone:

G.E.C.

The M-O Valve Company Limited

BROOK GREEN WORKS
HAMMERSMITH · LONDON W6
TEL: 01.603.3431 TELEX 23435

EE 01 152 for further details



Widest frequency range of any generator available

The new Sanders Microwave Signal Generator type 6459 has the following characteristics:—

- ✱ **Extremely wide frequency range, 3.5 to 12.0 GHz.**
- ✱ **Direct reading frequency dial.**
- ✱ **Direct reading Attenuator, calibrated in voltage and dBm.**
- ✱ **High stability and accuracy.**
- ✱ **Internal FM, CW, pulsed or square wave modulation.**
- ✱ **Optional conversion to rack mounting**

FM, CW, square wave and pulse waveforms are generated internally and the instrument has facilities for the application of external modulations. The pulse repetition rate is variable from 40-4000 pps, and pulse width from 0.5 to 10 microseconds. Synchronising signals are simultaneous with the Rf pulse, or between 3 and 300 microseconds in advance of the Rf pulse. The internal pulse generators can also be synchronised by means of external signals.

Power Output: (*a*) direct—an insertion probe couples out power to 80 mw. (*b*) 0.223 volt to 0.1 microvolt (0 to —127 dBm) between 4.5 to 11.0 GHz and —6 dBm to —127 dBm between 3.5 and 4.5 GHz. Above 11.0 GHz the output power is generally better than —6 dBm into a 50 ohm load. Type N co-axial connector.

Power Output Accuracy: within ±2 dB from —7 to —127 dBm when terminated in a 50 ohm load, within the frequency range 3.5 to 11.5 GHz.

For the measurement of Receiver sensitivity, sensitivity or rejection signal—to—noise ratio, aerial gain and transmission line characteristics, in addition to numerous specialised applications, the 6459 is a most versatile instrument for single band operation in the frequency range 3.5 to 12.0 GHz.

Sanders DIVISION

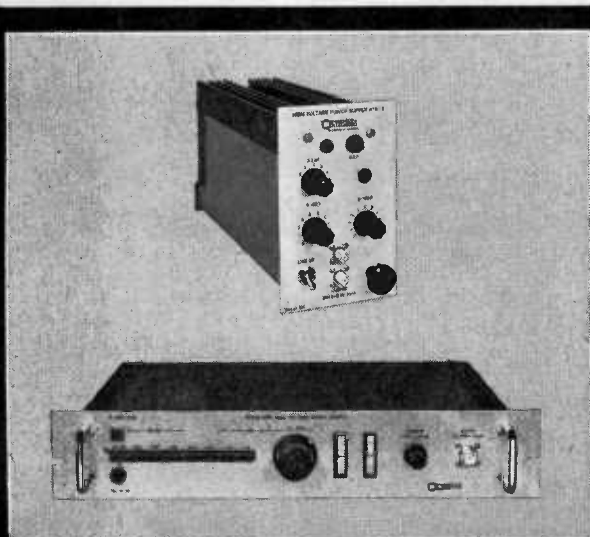
Marconi Instruments Limited—Sanders Division

GUNNELS WOOD ROAD
STEVENAGE, ENGLAND
Tel: Stevenage 2311
Telex: 82159 Sanders Stev.

EE 01 155 for further details

Here's OLTRONIX
New Blue Profile Line

COMPLETE DC POWER SUPPLY PROGRAM



HIGH VOLTAGE MODELS

Two Precision High Voltage DC Power Supplies

- 300–5000V, 2mA
- 0,002 % stability
- Current limit
- Small size

The power supply is made as a plug-in module according to the ESONE system, but can also be used as a separate bench model.

- 10–2500V, 10mA
- 0,001 % stability
- Programmable
- Adj. current limit
- Digital readout
- Reversible output

Meters on front panel for quick indication of output voltage, current and **polarity.**

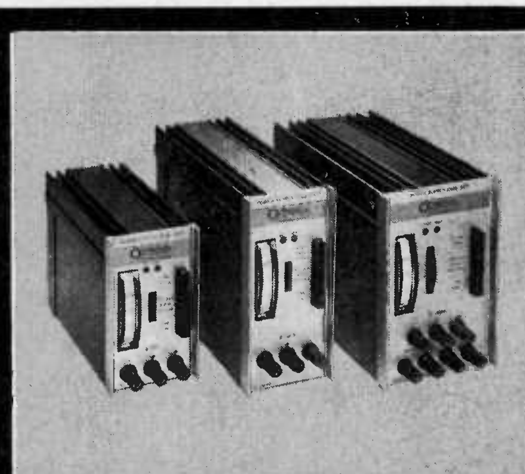
LABPAC

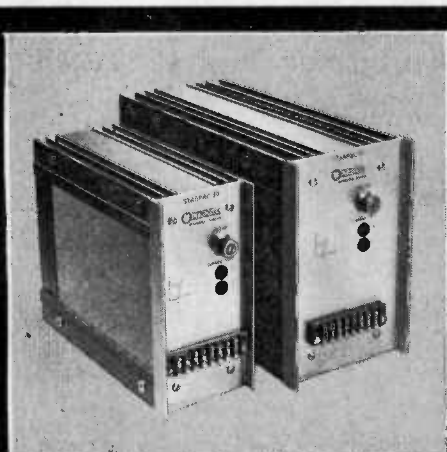
Six Models of Low Cost General Purpose Laboratory Power Supplies

- Programming
- All silicon
- Laboratory or systems use
- 15–30W
- Compact design
- Dual Range* output

"Fold back" current limit and range gives maximum output at minimum price and volume.

*Push a button on the front panel and LABPAC gives double the current at half the voltage.





STABPAC

50 Models of Compact Modular DC Power Supplies 6-500W

- All silicon
- Current limit
- Programming
- Remotse sensing
- Flexibility in mounting
- Rugged, compact design
- Adj. output voltage
- High ambient temp.

Two different referens-amplifier plug-in boards can be choosen to facilitites 0,01 % or 0,25 % line regulation. STABPAC can be mounted on three of its surfaces and also in 19" rack adapters.

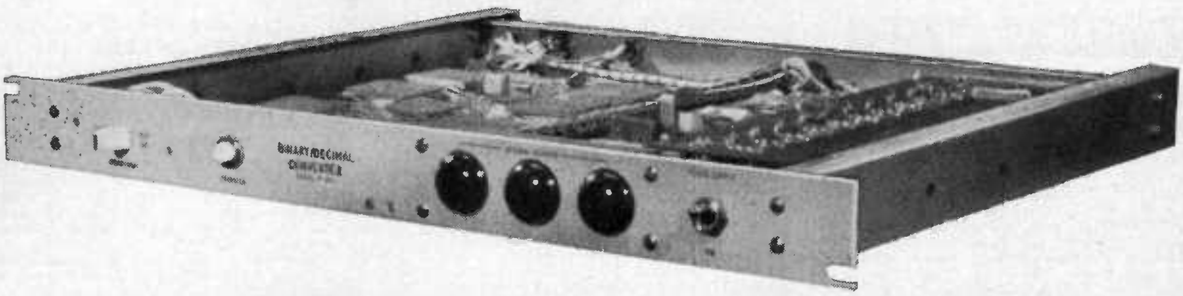
Production for EFTA
OLTRONIX AB
Jämlandsgatan 125, Vällingby, Sweden
Tel.: 08/87 03 30 Telex: 10738

Production for EEC
OLTRONIX-NEDERLAND N.V.
Vredenburgsweg 7, Hoogezand, Holland
Tel.: 05980-2301 Telex: 53301

Argentina Sirex, Buenos Aires. **Austria** Technical Office International, Vienna. **Belgium** Miravox, Brussels. **Danmark** Oltronix A/S, Copenhagen. **England** Miles Hivolt Ltd., Shoreham-by-Sea. **Finland** OY Findip AB, Helsinki. **France** Sorelia, Courbevoie-Paris. **Western Germany** Mütron Müller & Co., Bremen. Wenzel Elektronik, München. **Greece** Marios Dalleggio Representations, Athens. **Holland** Oltronix-Nederland N.V., Hoogezand. **Israel** Palec Ltd., Tel Aviv. **Italy** Societa 'Elettronica Lombarda, Milano. **Norway** Morgenstierne & Co., Oslo. **South Africa** Vitaphone Ltd., Johannesburg. **Spain** Ataio Ingenieros, Madrid. **Switzerland** Société d'Electronique Nucléaire, Geneva.

OLTRONIX

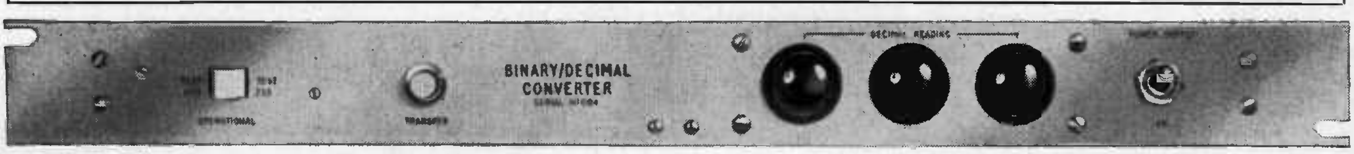
Ether Engineering
'Low-Line'
Digital Equipment



The use of standard thin film modules together with printed circuit techniques, enable Ether Engineering to offer a series of 1¾ in high, digital equipment for 19 in rack mounting, in addition to bench models.

Binary-Decimal Converter Model MA 1101
- ☐ Rapid decimal conversion: three neon numerical indicators
- ☐ Input: up to 8 parallel binary digits
- ☐ Logic: '0' = 0V to +0.5V
  '1' = +5V to +12V
  (other logic levels available)
- ☐ Conversion Instruct Signal: 5V positive going
- ☐ Conversion Time: 204mSec max.
- ☐ Aperture Time: ≤ 200μSec
- ☐ Power Supply: 12V at 600mA

Analogue-Digital Converter Model MA 1102
- ☐ Full scale analogue input: 10V (corresponding to 2^{10} digits)
- ☐ Resolution: 1 in 1024
- ☐ Accuracy: ± 0.1% ±½ digit
- ☐ Input resistance: 10kΩ
- ☐ Digital output: 10 parallel lines, binary
- ☐ Output levels: +5V through 8kΩ
- ☐ Conversion Time: 500μSec
- ☐ Sampling: automatic, manual or external
- ☐ Power Supply: 240V 50 c/s



The Ether Engineering series of 'Low-Line' equipment embodies reliable, robust, stable circuitry designed down to size, to get the utmost out of rack space.

Ether Engineering Ltd

PARK AVENUE · BUSHEY · HERTS
TELEPHONE: WATFORD 28566

E.E.8

CLASSIFIED ADVERTISEMENTS

RUN-ON AND SEMI-DISPLAYED The charge for these advertisements is 6s. per line of approximately 6 words (Minimum 24s.), for Run-on advertisements. For Semi-displayed advertisements the charge is 72s. per single column inch. Use of Box Number 3s. extra. Copy is required by 14th of the month for insertion in the next issue.

FULLY DISPLAYED (With border rules, larger type faces and blocks if required). The charge is 80s. per single column inch—minimum 3 inches. Eighth page £15 0s. 0d. Quarter page £30 0s. 0d. Half page £57 10s. 0d. Full page £115 0s. 0d. Use of Box Number 3s. Blocks must be supplied mounted. Copy is required by 5th of the month (if proofs are wanted), or by 10th of the month (without proofs) for insertion in next issue.

Advertisements for publication and replies to box numbers should be addressed to:

Classified Advertisement Dept., **"Electronic Engineering", 28 Essex Street, Strand, London, W.C.2.** *CENtral 6565*

OFFICIAL APPOINTMENTS

IMPERIAL COLLEGE OF SCIENCE & TECHNOLOGY

EXPERIMENTAL OFFICER

Experimental Officer required in Engineering in Medicine Laboratory; duties to assist research in medical instrumentation and medical applications of digital computers. Experience in instrumentation or data processing an advantage. Qualifications: H.N.C. or equivalent. Salary as appropriate to age and experience in Scales E.O. £1,335—£1,745 or A.E.O. (with prospects of promotion) £600—£1,300 p.a. Applications in writing to J. C. Ganley, Electrical Engineering Department, Imperial College, S.W.7. W 8467

LEICESTERSHIRE

LOUGHBOROUGH TECHNICAL COLLEGE

PRINCIPAL: F. LESTER, B.Sc., Ph.D., F.R.I.C.

DEPARTMENT OF ELECTRICAL ENGINEERING

Applications are invited for the post of Assistant Lecturer Grade 'B' in the Department of Electrical Engineering. Duties will include the teaching of electronics and electrical technology in National Certificate and Electrical Technicians' courses.
Applicants should be suitably qualified and should be familiar with both valve and transistor circuitry. A good background of industrial experience is required and teaching experience would be an advantage.
Salary will be in accordance with the Burnham Further Education Award, 1965, viz: Assistant Lecturer Grade 'B', £955—£1,845, with placing on the appropriate scale according to qualifications and experience.
Further particulars may be obtained from the Principal, Loughborough Technical College, Radmoor, Loughborough, Leicestershire, to whom completed applications should be returned by 21st December, 1966. W 8465

THE UNIVERSITY OF MANCHESTER INSTITUTE OF SCIENCE AND TECHNOLOGY

CHIEF TECHNICIAN

A Chief Technician is required in the Department of Electrical Engineering and Electronics for duties as follows.
(i) Construction, design and maintenance of digital equipment associated with the simulation and control of road traffic.
(ii) Servicing, maintenance, and extension of the on-line data-link facilities between the Institute and the Atlas Computer at the University of Manchester.

AGE LIMIT: 25–54 years.
Salary Scale
£1,242—£1,423 per annum according to experience, rising to £1,656 per annum in special circumstances.
5 Day–37½ hour week—Superannuation Scheme—Canteen Facilities.
Requests for Application Form quoting Reference No. EE2 should be sent to the Bursar, The University of Manchester Institute of Science and Technology, Room M2, Velvet House, Sackville Street, Manchester 1. Completed Forms to be returned as soon as possible. W 8445

SCIENTIFIC CIVIL SERVICE

EXPERIMENTAL OFFICER CLASS

Careers in SCIENCE and TECHNOLOGY are offered to men and women with qualifications in MATHEMATICS, PHYSICS, CHEMISTRY, BIOLOGY, METEOROLOGY, ENGINEERING and GEOLOGY as Assistant Experimental Officers and Experimental Officers, in Government Scientific Establishments.

QUALIFICATIONS: University Degree, Dip. Tech., H.N.C. etc.,—candidates under age 22 may offer lower qualifications (minimum: G.C.E. in five subjects; including two scientific/mathematical subjects at 'A' level or equivalent level.)

SALARY and AGE LIMITS:
Experimental Officer (age 26–30) £1,365—£1,734.
Assistant Experimental Officer (ages 18–27) £568 (at 18)—
£1,017 (at 26 or over)—£1,243.
Salaries supplemented in London area. Aid given for further education. Promotion prospects. Non-contributory pension.

APPLICATION FORMS AND BOOKLET from Civil Service Commission, Savile Row, London, W.1. Please write quoting S/579–580/66. Closing date extended to 31st December 1966. Candidates who have already applied need not do so again. W 8462

CITY OF LIVERPOOL

EDUCATION COMMITTEE

LIVERPOOL REGIONAL COLLEGE OF TECHNOLOGY BYROM STREET, LIVERPOOL, 3

DEPARTMENT OF CHEMISTRY & BIOLOGY

TECHNICAL ASSISTANT FOR INSTRUMENTATION

Applications are invited for the above post from persons with a practical knowledge of electronics and the physics of instrumentation, to be responsible for the maintenance and assembly of chemical instruments. Specific knowledge of chemical instruments is desirable but not essential. Chemical knowledge is not necessary.
Salary in accordance with Scale APT I (£735—£960). Commencing salary to be determined at interview.
Application forms (returnable to the Principal by 7th November 1966) from the Director of Education, 14 Sir Thomas Street, Liverpool, 3.

THOMAS ALKER,
Town Clerk.
W 8443

THE HOSPITAL FOR SICK CHILDREN

GREAT ORMOND STREET, LONDON, W.C.I.

CHIEF TECHNICIAN

Chief technician required to take charge of medical electronics workshop (staff of three including chief technician), engaged in development and maintenance of hospital and research apparatus. H.N.C. in electronic engineering. Equivalent salary £1,073 to £1,383 per annum plus £75 London Weighting. Application forms from the Assistant Secretary. W 8453

UNIVERSITY COLLEGE OF NORTH WALES, BANGOR

SCHOOL OF ENGINEERING SCIENCE

CONTROL ENGINEERING DIVISION

Applications are invited for the post of Lecturer in Control Engineering in the School of Engineering Science. Applicants should have interests in the theoretical or experimental aspects of modern control techniques. Honours graduates with a keen research interest in the field of automatic control are invited to apply.
Salary will be according to qualifications and experience on the scale £1,470—£2,630 and in addition there are F.S.S.U. benefits.
Applications (two copies), giving details of age, qualifications and experience together with the names of three referees should be sent by November 30th to the Registrar from whom further particulars may be obtained. W 8449

UNIVERSITY COLLEGE OF NORTH WALES, BANGOR

SCHOOL OF ENGINEERING SCIENCE

RESEARCH FELLOWSHIP

APPLICATIONS ARE INVITED for the Turner and Newall Research Fellowship in Applied Science. The research will be concerned with novel types of electrical machine of special significance to Control Engineering. Applicants should have had two or three years postgraduate experience, or industrial experience, and should preferably have some knowledge of semiconductor devices.
The appointment, which will take effect from a date to be mutually agreed with the successful applicant, will be for three years in the first instance, and the award will be in the range £1,000 to £1,600 per annum, with F.S.S.U. benefits.
Two copies of the application should reach the REGISTRAR, from whom further particulars may be obtained, as soon as possible, but not later than December 15th, 1966.
W 8460

THE UNIVERSITY OF WARWICK

SCHOOL OF ENGINEERING SCIENCE

RESEARCH ASSOCIATE IN AUTOMATIC CONTROL

Applications are invited for a post of Research Associate in the School of Engineering Science, to be concerned with programming a new on-line digital computer used for research in optimisation techniques applied to a wide range of systems. Some previous programming experience will be an advantage, and a degree or equivalent qualification is essential.
The appointment is for a period of up to two years, and the salary will be up to £1,000 p.a., together with F.S.S.U. benefits.
Further details and application forms are obtainable from the Registrar, University of Warwick, Coventry, Warwickshire. W 8456

CLASSIFIEDS
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OFFICIAL APPOINTMENTS

THE UNIVERSITY OF SUSSEX

LABORATORY OF EXPERIMENTAL PSYCHOLOGY PATTERN RECOGNITION

ELECTRONIC TECHNICIAN required to develop novel circuits for learning to recognize patterns and with application to sensory aids for the Blind; there are good laboratory facilities for this rewarding work.

Salary within the range £683 to £968 per annum. Three weeks' annual holiday. Superannuation scheme.

Applications (four copies) should give full details of age, qualifications, experience, present salary and names of two referees and be sent to the Assistant Registrar (Establishment), the University of Sussex, Stanmer House, Stanmer, Brighton, as soon as possible. W 8471

MINISTRY OF DEFENCE (ARMY DEPARTMENT)

ROYAL ARMAMENT RESEARCH AND DEVELOPMENT ESTABLISHMENT

POTTON ISLAND, GREAT WAKERING, SOUTHEND-ON-SEA

EXPERIMENTAL OFFICER, or ASSISTANT EXPERIMENTAL OFFICER aged at least 22, to develop instrumentation and ancillary equipment associated with measurement of transient phenomena. Experience of general electronics, the application of electronic principles and new developments to meet a wide variety of experimental requirements required; experience of high speed instrumentation desirable. QUALIFICATIONS: Degree, Dip. Tech., H.N.C.. or equivalent in appropriate subject.
SALARY: A.E.O. £803 (at 22)—£1,017 (at 26 or over)—£1,243; E.O. (minimum age 26) £1,365—£1,734.
Prospects of permanent pensionable appointment. APPLICATIONS to Ministry of Defence, CE2(F) (AD), London, W.C.2. W 8457

UNIVERSITY OF CAMBRIDGE

CAVENDISH LABORATORY

VHF AND MICROWAVE DEVELOPMENT

Assistant required for circuit development, mainly in applications to radio astronomy. 1–2 years experience in VHF, UHF or microwave fields desirable.
Day-release granted for approved study. Appointment in University Scientific Assistant Grade (salary, if suitably qualified, £590—£849 p.a.) or Technical Assistant Grade, (salary £796—£1,068 p.a.) Initial salary according to age, experience and qualifications.
Applications should be sent to The Secretary, Cavendish Laboratory, Free School Lane, Cambridge. W 8454

THE CITY UNIVERSITY

ST. JOHN STREET, LONDON, E.C.I.

DEPARTMENT OF PRODUCTION TECHNOLOGY AND CONTROL ENGINEERING

Appointment of
EXPERIMENTAL OFFICER

Applications for this post are invited from holders of a British university degree or of qualifications to satisfy the examination requirements for the status of Chartered Engineer.

Salary scale: £1,300 × £75—£1,825 per annum.

The duties will include the development and supervision of construction, installation and commissioning of apparatus and rigs for research, post-graduate instruction and for use in the last years of the honours degree courses. Form of application and further particulars from Head of Department. W 8455

IMPERIAL COLLEGE

LONDON, S.W.7.

ELECTRONICS TECHNICIANS

Vacancy for Technician or Senior Technician to join lively development group concerned with application of electronics techniques to Chemical Engineering and Technology research. This is an excellent opportunity for an ambitious young man to gain experience in transistor circuitry, instrument design and systems engineering. Good pension scheme and conditions. Salary: Technician £713—£1,078, Senior Technician £972—£1,290 depending on qualifications and experience.

Write in confidence to Professor G. R. Hall, Department of Chemical Engineering and Chemical Technology, Imperial College, Prince Consort Road, London, S.W.7. W 8458

IMPERIAL COLLEGE

CHIEF TECHNICIAN FOR ELECTRONICS DEPARTMENT

CHIEF ELECTRONICS TECHNICIAN. Essentially a practical post with plenty of scope for original design in conventional transistor circuitry, micrologic and advanced digital and analogue systems for applications in Engineering and Technological research.

Good pension scheme and working conditions.
Salary in range £1,302—£1,483.

Write in confidence to Professor G. R. Hall, Department of Chemical Engineering and Chemical Technology, Imperial College, Prince Consort Road, London, S.W.7. W 8459

RADIOLOGICAL PROTECTION SERVICE

MINISTRY OF HEALTH AND MEDICAL RESEARCH COUNCIL

TECHNICIAN/JUNIOR TECHNICAL OFFICER

Technician/Junior Technical Officer required by the Birmingham Regional Centre situated at the Queen Elizabeth Hospital. The Laboratory serves the Midlands, East Anglia, South West England and South Wales. It gives advice on radiological hazards in industry and research establishments etc. and carries out experimental work in that connection.

The person appointed will devote most of his time to the development and construction of electronic instruments and on maintaining and improving existing equipment.

Applications will be considered from persons with A levels in scientific subjects, O.N.C. or H.N.C. and possessing relevant experience. 5 day week, superannuation and other M.R.C. conditions. Salary within the ranges £577—£1,201 according to age, experience and qualifications.

Application (two referees) to the Honorary Director, R. F. Farr, Esq., M.A., F.Inst.P., Radiological Protection Service, Queen Elizabeth Hospital, Edgbaston, Birmingham 15. W 8446

CLASSIFIEDS

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SITUATIONS VACANT

FARE PAID TO USA for EXPERIENCED SEMICONDUCTOR DEVICE ENGINEER for work on diode lasers, will train/educate, interesting position, small company. Seed Electronics Corporation, 258 East Street, Lexington, Mass. 02173, USA. W 3083

TEST AND SERVICE ENGINEER experienced in testing, commissioning and servicing of relay and solid state logic systems required by company manufacturing Machine Tool Controls London W.C.1. Applicants who are prepared to travel in U.K. and abroad should write to Box W 8451, c/o ELECTRONIC ENGINEERING.

DAVY FARADAY RESEARCH LABORATORY, ROYAL INSTITUTION OF GREAT BRITAIN

EXPERIMENTAL OFFICER (ELECTRONICS)

An Experimental Officer (Electronics) is required for the design and maintenance of equipment for the study of fast chemical reactions, using nanosecond light flashes and laser techniques. Degree desirable but not essential. Salary and conditions similar to those of universities. Applications with names of two referees to Professor G. Porter, F.R.S., Director, The Royal Institution, 21 Albemarle Street, London, W.1. W 8450

THE BRITISH NON-FERROUS METALS RESEARCH ASSOCIATION

INSTRUMENTATION ENGINEER

Instrumentation Engineer, preferably with at least H.N.C. or equivalent qualification in electronics, electrical engineering or physics. Responsibilities will include development of practical instruments from new ideas. Age 20–30. Apply: Secretary/RE, The British Non-Ferrous Metals Research Association Euston Street, N.W.1. W 8448

UNIVERSITY GRADUATE

in Electronics, preferably with industrial experience in research or development, to train as Patent Agent with London firm of Chartered Patent Agents. Preferred age 22–26. Patent Agency provides an interesting and rewarding career. Financial prospects within the profession are exceptionally good. Write, with brief particulars, Frank B. Dehn & Co., Imperial House, 15–19, Kingsway, London, W.C.2. W 8461

ELECTRONIC CIRCUIT DESIGNER

PERA has a vacancy for an Electronic Circuit Designer to work on original equipment design in a wide variety of fields including R.F., A.F. and Digital. This is a responsible position requiring close collaboration with industry. Excellent opportunities for advancement, generous leave allowance, superannuation under F.S.S.U., attractive commencing salary. Full details in confidence to the Secretary (R.1923/67), Production Engineering Research Association, Melton Mowbray, Leics. W 8452

CANADA, AUSTRALIA, SOUTH AFRICA

We have a steady flow of enquiries for draughtsmen and engineers. To be ready for anything that comes we keep a register of men seeking posts overseas.
Write to Major ENgagements Ltd., Sheepy Parva, Atherstone, Warwickshire. W 8469

SITUATIONS VACANT

MMMMMMMMMMMM

ADVANCED VISUAL DISPLAY

If you are a circuit engineer and would like to join a team concerned with advanced visual display systems, Mullard Research Laboratories would like to hear from you.

We have a senior vacancy in this team, the work of which includes colour television display. The man we wish to appoint will be of graduate or equivalent level, able to assume personal responsibility for his own work under the overall supervision of the project leader.

We are looking for candidates whose approach is sufficiently flexible to be able to work on other television or related circuit problems, as our programmes of work change.

We are also looking for one or two circuit engineers at a more junior level for work on associated projects.

If you are interested in joining the Laboratories, please write giving a brief outline of qualifications and relevant experience and quoting reference 849/EE to:-

Dr. S. A. Ahern,
Mullard Research Laboratories,
Redhill, Surrey.

Mullard

W 8472

MMMMMMMMMMMM

SITUATIONS VACANT

BERRY'S RADIO require SALES ASSISTANT with good technical knowledge. 5-day week. Top salary—permanency. 25, HIGH HOLBORN, W.C.1 W 8417

PATENTS

THE PROPRIETORS OF BRITISH PATENT No. 930901 for "Method of Manufacturing Hollow Cylindrical Members with Apertured Internal Partitions." desire to enter into negotiations with a firm or firms for the sale of the patent, or for the grant of licences thereunder. Further particulars may be obtained from Marks & Clerk, 57/58 Lincoln's Inn Fields, London, W.C.2. W 8466

THE PROPRIETORS OF BRITISH PATENT No. 933205 for "Solenoid Controlled Distributing Slide Valve" desires to enter into negotiations for the sale of the patent, or for the grant of licences thereunder. Further particulars may be obtained from Marks & Clerk, 57/58 Lincoln's Inn Fields, London, W.C.2. W 8447

CAPACITY AVAILABLE

CAPACITY AVAILABLE wiring of electronic equipment, prototypes and quantities: AUTO-GUARD PRODUCTS, 25, Wynnstay Road, Old Colwyn, COLWYN BAY. W 3080

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

EXTRA PRODUCTION FACILITIES: Skilled workers + floor space + expert supervision + extra production facilities for you. See page 108. W 374

FAST SERVICE for design and winding coils prototypes quantity. Also considerable facilities for design & production of assemblies, wiring etc. Electricole Ltd., 2-14, Shortlands, Hammersmith, London, W.6. W 378

SERVICE

ELECTRONICON LTD. From drawing board to production with minimum delay. Let us be your research, development, design and prototype department. Pilot production runs a speciality. 176, Lythalls Lane, Coventry. Tel. Nuneaton 2353. W 372

SITUATIONS VACANT CONTINUED

Electronic Technicians

TEKTRONIX GUERNSEY LTD. would like to hear from men who feel that after a short period of familiarisation, they would be competent to test and calibrate a range of the Company's oscilloscopes. Those with previous electronics experience who would like to work for a Company offering good conditions of employment, should write with full details to

TEKTRONIX

The Personnel Manager
P.O. Box 65
Victoria Avenue
St. Sampson's, Guernsey

W 8463

MONOCHROME AND COLOUR TELEVISION

WARWICK ELECTRONICS INC. one of the world's largest manufacturers of television receivers has several outstanding vacancies for qualified engineers.

Permanent career opportunities as a result of long term expansion programmes at our headquarter laboratories for:

ADVANCE DEVELOPMENT ENGINEERS

CIRCUIT DESIGN AND DEVELOPMENT ENGINEERS

Candidates should possess a degree in Electronic Engineering; H.N.C.; its equivalent or specialist experience of a high order together with a minimum of 3 years experience in industry. Familiarity with semi-conductor techniques and/or use of computers in design a decided asset.

Location is in Niles, Illinois—a suburb north of Chicago with outstanding educational, recreational and social facilities. Excellent insurance, medical and pension benefits. Salaries commensurate with qualifications and experience. Profit Sharing and Cash Bonus Plan.

Liberal relocation allowance and housing assistance for successful candidates and their families.

Interviews will be arranged for in London Area between November 29th and December 5th.

Please submit details to Box No. W 3084 c/o Electronic Engineering.

Science Research Council

DARESBURY NUCLEAR PHYSICS LABORATORY

A large Electron Synchrotron (NINA) will be brought into operation this year and will be used for nuclear research by university and resident research teams.

NINA MACHINE GROUP

TECHNICIANS

A number of technicians, at both senior and junior levels, are required to work in the NINA Machine Group. They will play an important role in the Group as members of the operating team running the synchrotron coupled with responsibilities, under the supervision of senior physicists and engineers, for development of various aspects of the synchrotron to improve its performance. These appointments will entail some regular three-shift working.

Applicants should have served a recognised apprenticeship or have had comparable training and in addition should possess O.N.C. for the Grade II and III posts or H.N.C. for the Grade I posts in electrical engineering or applied physics. Several years' experience in instrument technology, electronics, light electrical engineering or power engineering is desirable and knowledge of one of the major fields of accelerator technology such as radio frequency or vacuum would be an added advantage.

The appointments will carry membership of a non-contributory superannuation scheme and will attract salaries in the range:-

Grade III £945 at age 26, rising to £1009 at age 28 and over, to a maximum of £1129.

Grade II £1129-£1288 per annum

Grade I £1288-£1586 per annum

Grade will be according to age and experience.

Please send postcard quoting reference No. DL/118 for application form to:-

**Personnel Officer,
SCIENCE RESEARCH COUNCIL,
DARESBURY NUCLEAR PHYSICS LABORATORY,
Daresbury, Nr. Warrington, Lancashire.**

Closing date for receipt of completed application forms is 9th December, 1966.

W 8444

SERVICE

BINDING OF VOLUMES

Readers can have their copies of ELECTRONIC ENGINEERING bound, complete with index and with advertising pages removed, in a good quality red cloth covered case, lettered in gold on the spine, including packaging and return postage at a cost of £3 10s. per volume.

Home and overseas readers who require their issues for 1966 bound, are asked to comply with the following instructions:

Tie the issues together, enclose a remittance of £3 10s., with the sender's name and address, and despatch carriage paid in a closed parcel to:

The Circulation Dept. (E.E. Binding), 28 Essex Street, Strand, London W.C.2. (Cheques or postal orders should be made payable to Morgan Brothers (Publishers) Ltd).

Readers who wish to make other arrangements for binding their own copies can purchase binding cases (suitable for firm binding only) at a cost of 15/- each including postage.

IBM Laboratories

Mask Technology and Photo Engineer

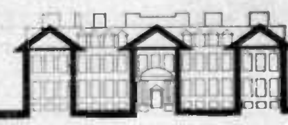
A vacancy exists for an engineer with experience in the latest techniques of integrated circuit and printed circuit photo mask technology. Experience should include advanced Plating Techniques.

The work of the Photo Engineering Laboratory is expanding, and thrusts at the limits of technology, and the very wide range of equipment in use is the most accurate available. A new clean area photo laboratory building is planned.

The appointment offers great scope for further advancement for an engineer with drive, knowledge and initiative and preferably with exempting qualifications for membership of a chartered professional institution.

Full details of the position, the company and its benefits, which include reimbursement of re-location expenses, will be given at interview.

Please write, in confidence, to the Personnel Manager, IBM United Kingdom Laboratories Limited, Hursley Park, Winchester, Hants.



BOROUGH POLYTECHNIC
Borough Road, S.E.1

Applications are invited for the post of

HEAD OF THE DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING

Candidates should possess high academic and professional qualifications in Electrical Engineering and have held a senior teaching post in this subject. Responsible research, industrial or administrative experience would be an added qualification.

SALARY SCALE:-

Head of department (Grade V), £2,970 – £3,210 p.a.

Further particulars and application form obtainable from the Clerk to the Governing Body, to whom completed applications should be returned by 31st December, 1966.

W 8470

Component Evaluation Engineer

A vacancy has arisen for an Engineer in a staff group. Must be familiar with the evaluation of components, including semi-conductors. A knowledge of oscilloscope circuits would be an advantage for work in a product line support function.

Please write for an application form to

The Personnel Manager
Tektronix Guernsey Ltd.,
P.O. Box 65
St. Sampson's, Guernsey

TEKTRONIX

W 8464

E.M.I Electronics Engineers

A number of vancancies exist for Electronics Engineers in the Research and Development Laboratories of E.M.I. Electronics Limited, Feltham, Middlesex.

These positions would be of particular interest to applicants with one or two years experience since qualifying who wish to broaden their experience in a wide and varied field.

CIRCUIT ENGINEERS For the design and development of radar circuits. The duties would include the preparation of specifications and compilation of reports on designs and development undertaken. Preference will be given to applicants with experience in pulse and VHF oscillation techniques.

DEVELOPMENT ENGINEERS For development of missile radar equipment and associated test equipment involving the testing and proving of units under varying environmental conditions.

COMPONENT ENGINEER Working in the Design and Evaluation Department to advise Project Engineers on components, their use, reliability etc. Duties include liaison with Ministry Departments, Services and Manufacturers on information relating to existing components and new development. Knowledge BSI and Industry Component Specifications, preferably with experience of Quality Control, would be an advantage.

Candidates should have a minimum qualification of HNC in Electronics Engineering.

EMI

Please write or telephone for an application form, quoting reference number EE PGB to:-

P. W. E. FOX, PERSONNEL OFFICER,
E.M.I. ELECTRONICS LTD.,
VICTORIA ROAD, FELTHAM, MIDDX.
Telephone No. FELTHAM 3600. Ext. 44.

W 8473

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Electronic Engineering

28 ESSEX STREET · STRAND · LONDON · W.C.2.

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AUTHORS