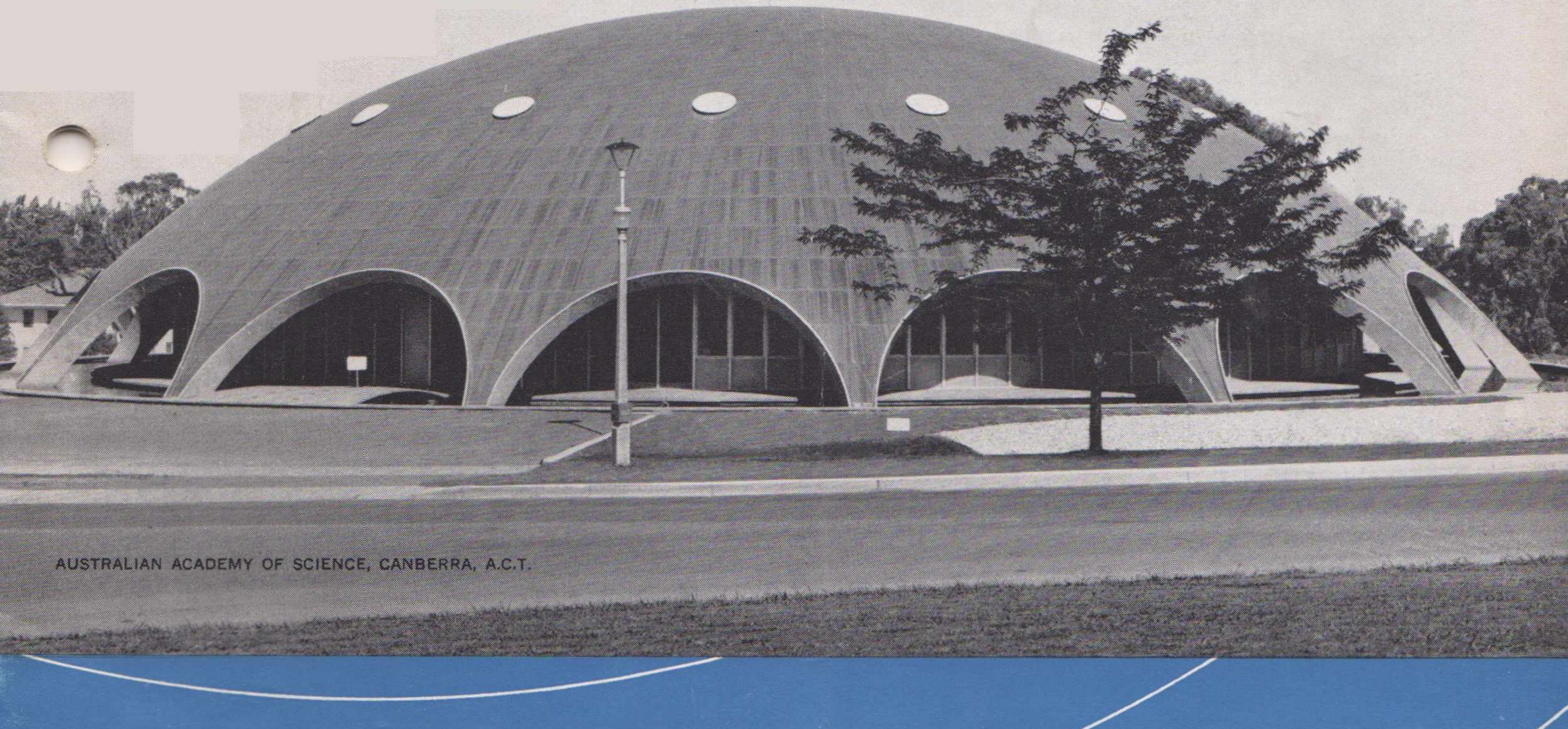
# Mullard Outlook

AUSTRALIAN EDITION

## 1965 I.R.E.E. CONVENTION ISSUE



AUSTRALIAN ACADEMY OF SCIENCE, CANBERRA, A.C.T.

MARCH-APRIL, 1965
VOL. 8, No. 2

MULLARD-AUSTRALIA PTY. LTD.

VOL. 8 — No. 2 MARCH-APRIL, 1965

Editorial Office:
35-43 Clarence Street, Sydney.
Telephone: 29 2006

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JOERN BORK





TABLE OF CONTENTS



1965 I.R.E.E. CONVENTION ISSUE

*The modern headquarters building of the Australian Academy of Science in Canberra A.C.T. Canberra is the venue of the 1965 I.R.E.E. Convention.*

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*". . . . . . . . . . . we will deny or defer to no one, right and justice."*

from the Great Charter 1215.

It is claimed and is self-evident that good government demands the intelligent interest of all citizens and it is a long, long time since John bowed to the barons at Runnymede.

The provisions against wrong and extortion which they drew up as against the King for themselves, they drew up as against themselves for their tenants . . . surely the significant point, to be followed fifty years later by Simon de Montfort's bold and happy innovation, a Parliament of Knights, Citizens and Burgesses which left to all later reformers nothing to do but improve it in detail—and Simon was by birth a stranger to England!

The Parliament in Canberra, our Parliament of the same blood line, when sitting is broadcast to all the people, communications in the free and noble sense and something Simon would have enthusiastically approved.

It is fitting therefore that Canberra with its fine University, its Academy of Science and seat of Government and its need to develop and keep good communications with the other peoples of the world is the venue for the 1965 Convention of the Institution of Radio and Electronics Engineers Australia.

May the Convention be complementary to the highly successful Dunrossil Memorial Lecture recently delivered by his Royal Highness the Duke of Edinburgh to honour a previous Patron of the Institution and Governor General of the Commonwealth of Australia, the late Viscount Dunrossil, onetime Speaker of the House of Commons, with its unbroken succession since Simon's Parliament of 1265.

M.A.B.

# VIEWPOINT WITH MULLARD

# MULLARD EDUCATIONAL SERVICE

Many Outlook readers are aware of the Mullard Educational Service and have made use of the teaching aids available, the booklets and pamphlets on particular electronic devices and their application in discrete end products. In this issue we detail the field covered in that greater use may be made of this Service and its contribution to the technical education of Australian youth. The Mullard Educational service in Australia has grown from the Educational Service offered by our Parent Company in the United Kingdom, this service is now in its tenth year.

**Tenth Anniversary of Mullard Educational Service**

To mark the occasion a special exhibition was arranged at Mullard House to show how it assists schools, technical colleges, universities, and industrial training schools, to teach physics and electronic engineering. The exhibition was opened on November 11, 1964, by Lord Bowden, Minister of State, Department of Education and Science, Her Majesty's United Kingdom Government.

Over the ten years, more than 8,000 teachers and lecturers in Great Britain and the Commonwealth have made use of the facilities and in Great Britain alone, since 1954, the number of individual enquiries received from teachers has risen rapidly to the present rate of 7,000 a year.

The Service provides aid to firms' training units, with material specially designed to teach electronics to students in industry. Help has been given to The Royal College of Surgeons in preparing a course to teach medical staff the fundamentals of electronics —a timely step when electronic equipment is being used increasingly in many branches of medicine.

The Service also acts as adviser on electronics courses, co-operates with the publishers of children's books and so on.

The exhibition showed the range of the material teachers can draw on, one section featuring more than 70 demonstration experiments devised by the Service's engineers and published in a series of booklets.

*Mr. B. P. A. Beresford.*

**Audio/Visual Teaching Aids**

Visual aids include films, filmstrips, slides, the Mullard Educational Catalogue now listing 28 films and 67 filmstrips and a host of publications and booklets describing simple measurement techniques, the design and construction of classroom apparatus. The Service is administered by Mr. B. P. A. Beresford, Manager of our Technical Service Department, an enthusiastic supporter of youth training in the broadest sense. The Service operates from our Head Office in Sydney and indirectly through our branch offices in each State and the facilities extend from basic electronics experiments for schoolboys and schoolgirls right through to the training of university students in nuclear physics.

The catalogue of 16mm sound films and 35mm filmstrips and lecture notes will gladly be forwarded on request, the films are available on loan or outright purchase.

*Lord Bowden, Minister of State, at the opening of the anniversary Exhibition at Mullard House, seen talking to Dr. F. E. Jones and Captain S. R. Mullard, founder and Director of the Company.*

**Constructional Projects and Equipment Designs**

Detailed in specific leaflets are practical experiments and demonstrations that can be constructed and operated at schools and technical colleges. It should be pointed out that the Mullard Educational Service supplies the idea—circuit details, a lay-out photograph, theory, calibration figures and operating instructions. We do not provide any of the equipment or kits but we are only too willing to answer any technical queries which may arise and to encourage the manufacture of our designs by other commercial companies.

**Wall Charts and Show Cases**

Selected wall charts are provided and at specific request, for technical colleges and the like, show cases are prepared for permanent displays of an educational value, for example the construction of valves, special electron tubes and semiconductors, and, more recently, examples of thin film circuit techniques.

**Publications**

Whilst the Mullard Educational Service embraces a wide group of publications and leaflets in its own right, these are complementary to the many technical publications of our Parent Company, Mullard Limited, London—as also the numerous publications produced locally in Australia by the Technical Service Department.

Practical experiments suggested by the Mullard Educational Service are discussed in this issue of Outlook on pages 34 to 40.

Some of these publications are available free to selected training establishments and special bulk purchase arrangements are →page 16

## MULLARD EDUCATIONAL SERVICE

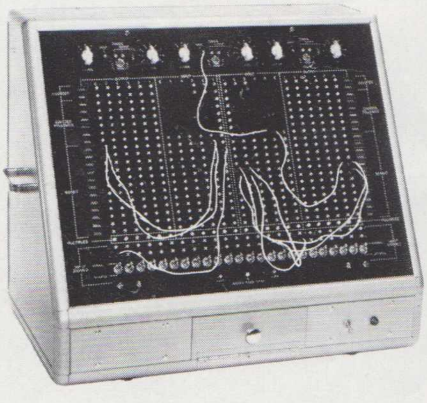
← page 15
available for books, booklets and more advanced technical publications.

### Technical Enquiries Service

The Enquiries Service operates in co-operation and at specific request from commercial concerns, institutions and various training establishments towards the suggestion and recommendation of a particular training approach, insofar as it is related to, say, electronic aids to research in medicine or a particular industry, for example, the chemical industry or the practical engineering approach to, say, static switching with solid state switching devices (Norbits). It is essential in the training of industrial control and automation technologists, that their syllabus of training include a comprehensive study and practical application of solid state switching.

### Mullard Norbit Simulator

For this purpose training establishments are shown how to demonstrate Norbits by applying these in representative practical arrangements augmented by the Mullard Norbit Simulator, a self-contained equipment. Although specifically intended for instructional purposes, the Simulator provides a means of demonstrating and checking the steps involved in solving switching logic problems as related to Norbit application.



*The Mullard Norbit Simulatar.*

### High-power Electromagnets for Universities and Technical Colleges

Whilst our normal commercial range of magnets extends from tiny Ticonal and Magnadur magnets for small instruments, hearing aids,' television receivers and the like—and extending right through to large high-power magnets, weighing many tons, for industry and research, Mullard offer a special 7″ electromagnet for universities, technical colleges and similar educational establishments to enable them to demonstrate phenomena associated with the fundamental physical properties of matter, such as nuclear resonance, Hall effect and magnetic susceptibility. A feature of this air-cooled magnet type EE1032 is the ease of access to the air gap made possible by keeping the distance between the coils as great as possible, so as to leave the gap area free from obstruction. The magnet can therefore be used in cryogenic experiments, involving the use of a Dewar flask.

A range of special pole pieces enables gap geometry to be varied and a series of accessories are available to extend the magnet's range of applications. All pole pieces are fully adjustable for optimum field uniformity and are made from specially selected flaw-free high quality iron.

The EE1032 is a low impedance electromagnet suitable for applications demanding high field uniformity and stability such as electron spin resonance spectroscopy and low resolution nuclear magnetic resonance.

### Equipment for Nuclear Physics Experiments

Mindful of the ever-increasing demand for advanced technological instruction, particularly in areas where students and graduates do not have the facilities of a nuclear reactor, linear accelerator or other high energy resource, we are pleased to introduce in this issue of Outlook, the Visiflux Neutron Howitzer. The device has been developed especially for training in Nuclear Science and Engineering. It is a product of imaginative design and extensive testing by leading educators in radio isotope technology, neutron physics and nuclear engineering. With the device is an experiment manual showing how it can be applied to eight specific experiments in radiochemical analysis and fifteen specific experiments in nuclear engineering and neutron physics.

A smaller unit is available with somewhat restricted facilities. The neutron source for both units being Americium-beryllium, Am-Be.

### Neutron Activation Foils

In addition to the Howitzer a full range of neutron active foils are available for training and research. Activation foils are probably the most widely used type of neutron detector, these are extremely reliable and a convenient method of obtaining neutron flux levels and spectra. An Experiments Manual is furnished with the educational foils, each Manual containing the following material — thermal, intermediate and fast neutron flux measurement, counting techniques and corrections, cad-

→page 35

*The Neutron Howitzer.*

# 1000,000,000,000 OHMS

## INPUT RESISTANCE

### Practical Application of the Mullard 95BFY M.O.S.T. (Metal-Oxide-Silicon Transistor)

*A description of the basic principles of the M.O.S.T. was given in Outlook Volume 8, Number 1. These principles are repeated here, for the convenience of readers, together with more detailed application engineering and typical circuitry.*

*The 95BFY is essentially a developmental type, but in view of the intense interest in four-element field-effect transistors, this article is presented purely as general information and in due course readers will be advised when production devices are available for initial equipment.*

One of the latest developments in the Mullard series of field-effect transistors is that of a four element type 95BFY. It is an insulated-gate field-effect transistor, the insulation of the gate being achieved by a silicon dioxide dielectric.

The transistor is thus well suited for use with high input impedance requirements and for circuits requiring low offset voltage and currents, such as chopper circuits, twin-T oscillators, high impedance transducers, timing circuits, etc. As the 95BFY is a majority carrier device it has no transit time effects (as usually found with conventional transistors) and thus the frequency response of the circuit may only be limited by its stray capacitance and the capacitance of the device. Because of its fourth electrode, the 95BFY may also be used as a variable gain amplifier, mixer, level detector, inhibit gate, DC amplifier, etc.

With four electrodes, a very large input resistance in excess of $10^{10}\Omega$, low input capacitance of 4 pF and a mutual conductance of better than 1 mA/V, the metal-oxide-silicon transistor is the nearest solid-state equivalent to the thermionic valve.

**Operation of the Transistor**

The M.O.S.T. 95BFY consists of two diffused super-rich n-regions on a p-type silicon "substrate". The n-regions, which are known as the "source" and the "drain" are separated by part of the p-type substrate. A thin layer of silicon dioxide acts as a dielectric between the substrate and a small metal electrode called the "gate" (see Fig. 1).

When the gate is made positive with respect to the source, electrons are attracted to the surface of the substrate. This forms a n-type layer (inversion layer) near the oxide dielectric. The greater the applied gate voltage, the thicker will be this inversion layer. The inversion layer forms a conducting path between the two super-rich n-regions (drain and source) and hence a current will flow if a potential is applied between them. Because the gate voltage ($V_{GS}$) controls the thickness of the inversion layer it consequently controls the magnitude of the drain current ($I_{DS}$).

Since the inversion layer is formed in the substrate the gate voltage affecting the inversion layer may be off-set by selecting a suitably doped substrate so that an inversion layer still exists when the voltage on the gate is either negative, positive or zero with respect to the source. The level of voltage thus required to completely remove the inversion layer (hence cutting off the drain current ($I_{DS}$) almost to zero) is called the "pinch-off voltage" $V_{GS(P)}$. With the gate voltage ($V_{GS}$) below $V_{GS(P)}$ the M.O.S.T. is well in the cut-off region (see Fig. 2).

The drain and source form P-N junctions with the substrate. Hence the source-to-drain cut-off current ($I_{DSX}$) is of the same order as the collector-to-base leakage current in a conventional silicon planar transistor. Thus $I_{DSX}$ is the leakage current in the drain-to-substrate N-P junction. The leakage between gate and substrate is extremely low, as the gate is insulated from the substrate by the silicon dioxide layer. Consequently the gate current is extremely low and the gate input resistance is of the order of $10^{12}\Omega$.

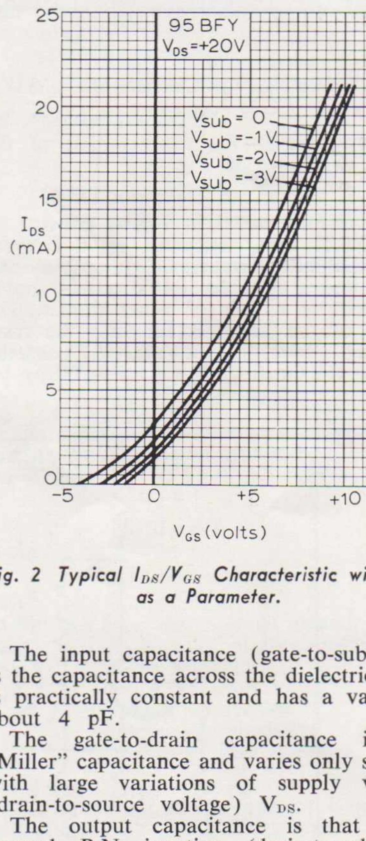



*Fig. 2 Typical $I_{DS}/V_{GS}$ Characteristic with $V_{sub}$ as a Parameter.*

The input capacitance (gate-to-substrate) is the capacitance across the dielectric. This is practically constant and has a value of about 4 pF.

The gate-to-drain capacitance is the "Miller" capacitance and varies only slightly with large variations of supply voltage (drain-to-source voltage) $V_{DS}$.

The output capacitance is that of a normal P-N junction (drain-to-substrate) and varies with $V_{DS}$.

All capacitances are independent of drain current $I_{DS}$.

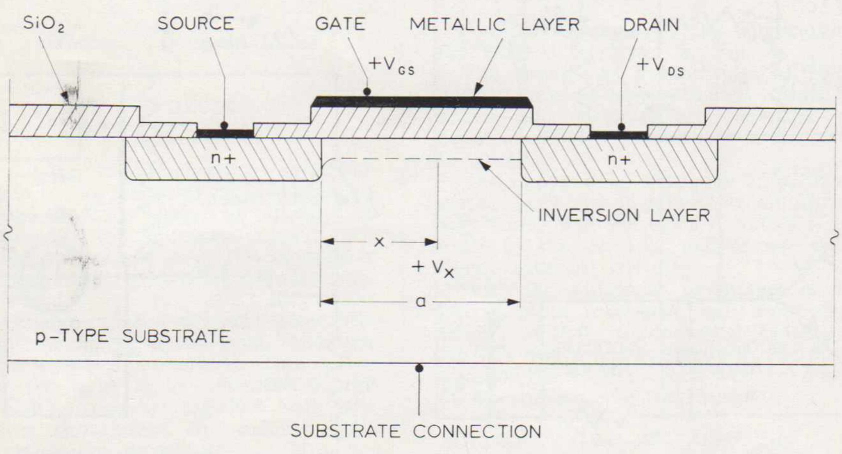



*Fig. 1 Diagrammatic Representation of a Metal-Oxide-Silicon Transistor.*

**Application**

Whilst the complete range of applications of the M.O.S.T. are too numerous to be covered in this article, some typical examples are briefly discussed. In many applications of the 95BFY the substrate is connected to the source, but if the substrate has a voltage negative with respect to the source the $I_{DS} - V_{GS}$ characteristics is shifted as in Fig. 2.

→page 18

# 1000,000,000,000 OHMS INPUT RESISTANCE

← page 17

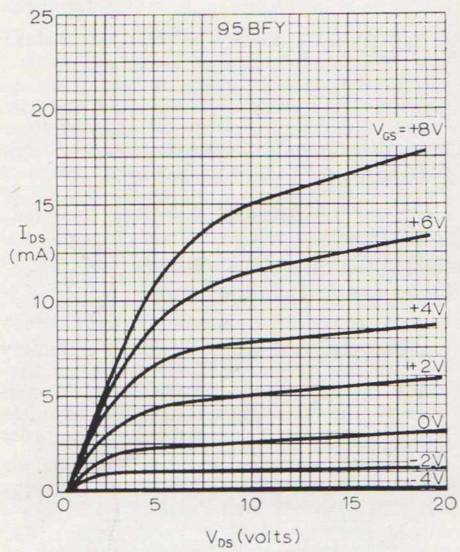



*Fig. 3 Typical $I_{DS}/V_{DS}$ Characteristic with $V_{GS}$ as a Parameter.*

A typical $I_{DS} - V_{DS}$ characteristic of the 95BFY is shown in Fig. 3.

**M.O.S.T. Timer and Dekatron Drive**

The circuit shown in Fig. 4 consists of a M.O.S.T. timer followed by a conventional Schmitt trigger, DC amplifier and keyer, and a suitable +15V power supply regulated by a pair of zener diodes. The output of the keyer is fed into the transistorised drive circuit of the dekatron counter via the 0·47μF capacitor $C_2$.

When the M.O.S.T. timer has triggered TR1 is bottomed leaving the keyer open to receive the 100c/s mains rectified pulses for a pre-determined period governed by the M.O.S.T. timer. The length of this period is displayed by the dekatron counter.

Other forms of M.O.S.T. timers are: monostable multivibrators, astable multivibrators, twin-T oscillators, etc. A monostable multivibrator is discussed below.

**M.O.S.T. Monostable Multivibrator Timer**

Fig. 5 shows a monostable multivibrator with an extremely long delay. The basic operation is the same as that of a conventional transistor monostable. With the 95BFY, however, the gate current remains virtually constant and does not increase sufficiently to hold the voltage across the capacitor $C_o$ constant once the M.O.S.T. has bottomed. Thus D1 serves as the clamping diode and charging path for $C_o$.

The delay is approximately given by:

$$t = C_o \cdot R_o \cdot \ln\left[\frac{2E_1 - V_{CE} - V_D}{E_1 - V_K}\right] \text{ secs.}$$

where $V_{CE}$ = TR1 collector-emitter saturation voltage

$V_K$ = determined from 95BFY characteristics and is a function of $R_L$ and $V_{GS(P)}$ (C in farads and R in ohms.)

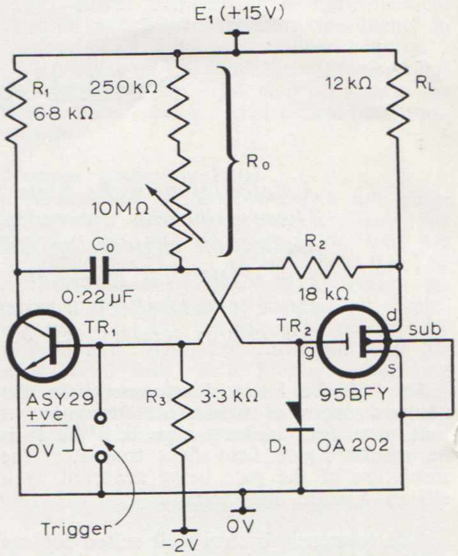



*Fig. 5 M.O.S.T. Monostable Multivibrator Timer.*

**M.O.S.T. Industrial Timer**

A suggested circuit for a simple timer is shown in Fig. 6.

When MS is closed D1 is reverse-biased allowing $C_o$ to charge up via $R_o$ towards 0V. The relay A will remain energised once the delay time has elapsed. The timer is reset by depressing the reset button.

Neglecting leakage currents in components, the time delay is approximately given by

$$t = C_o \cdot R_o \cdot \ln\left[\frac{E_2 - V_D}{E_1 - V_K}\right] \text{ seconds}$$

where $V_D$ = forward drop across D1

$V_K$ = a function of $V_{GS(P)}$ and the transistor load resistance

$V_{GS(P)}$ = pinch off voltage

Without using electrolytic capacitors time delays up to 30 seconds are readily possible.

→page 19

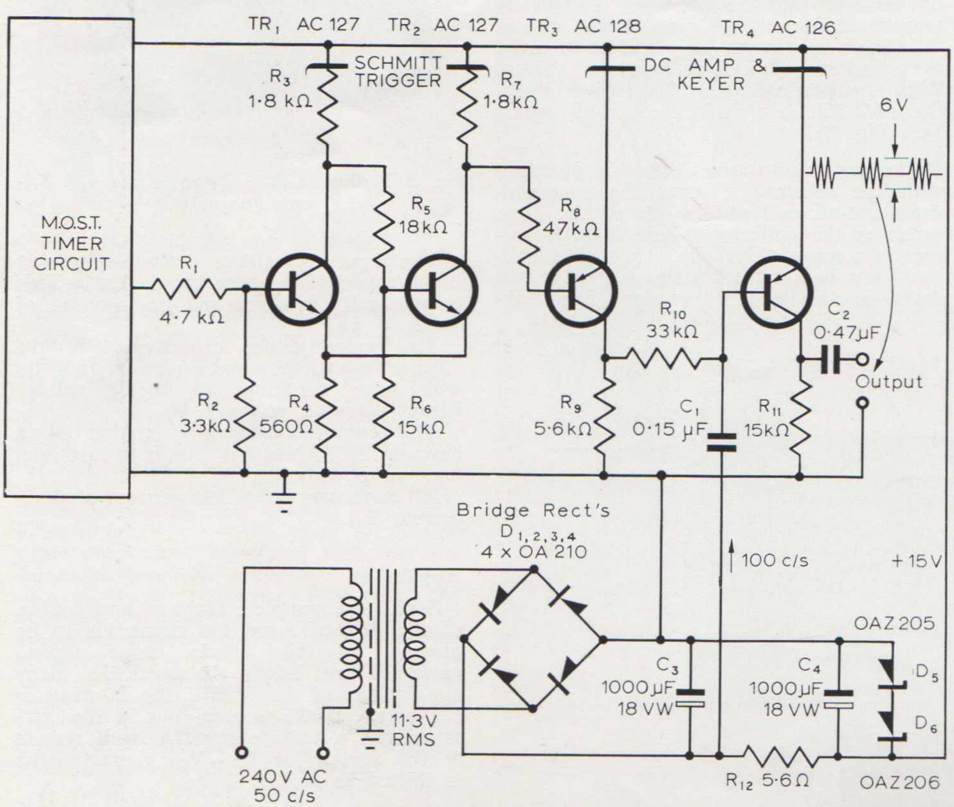



*Fig. 4 M.O.S.T. Timer and Dekatron Drive.*

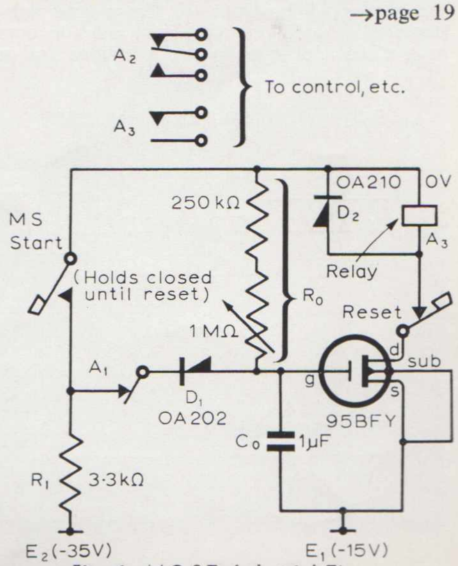



*Fig. 6 M.O.S.T. Industrial Timer.*

# 1000,000,000,000 OHMS INPUT RESISTANCE

← page 18

**A Simple Source Follower**

Fig. 7 shows a simple circuit of a source follower with its substrate connected to the source. This circuit is equivalent to a conventional valve (cathode) follower with C1 as the coupling capacitor and R1 the gate (grid) leak resistor.

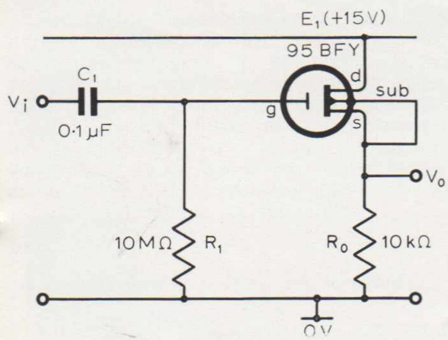



*Fig. 7 A Simple Source Follower.*

$V_i$ and $V_o$ are input and output voltages respectively.

The effective input is given by

$$V_i = RI_{DS} + V_{GS(P)}$$

$$V_o = RI_{DS}$$

$$I_{DS} = \left[\frac{V_i - V_{GS(P)}}{R} + \frac{1}{\beta R^2}\right] - \left[\frac{2\,(V_i - V_{GS(P)})}{\beta R^2} + \frac{1}{\beta R^2}\right]^{\frac{1}{2}}$$

The gain of the source follower is given by

$$\frac{d\,(RI_{DS})}{d\,V_i} = 1 - \left[\frac{1}{2\beta R\,(V_i - V_{GS(P)}) + 1}\right]^{\frac{1}{2}}$$

and is always less than unity.

Where $\beta$ = gain factor

As an example, typical values for the 95BFY are

$$V_{GS(P)} = -6V$$

$$\beta = 0.4\ \text{mA/V}^2 \text{ at } I_{DS} = 3\text{mA}$$

$$V_i = 12\ \text{V and } R = 10\text{k}\Omega$$

$$\text{The gain is} = 1 - \left(\frac{1}{49}\right)^{\frac{1}{2}}$$

$$= 0{\cdot}85$$

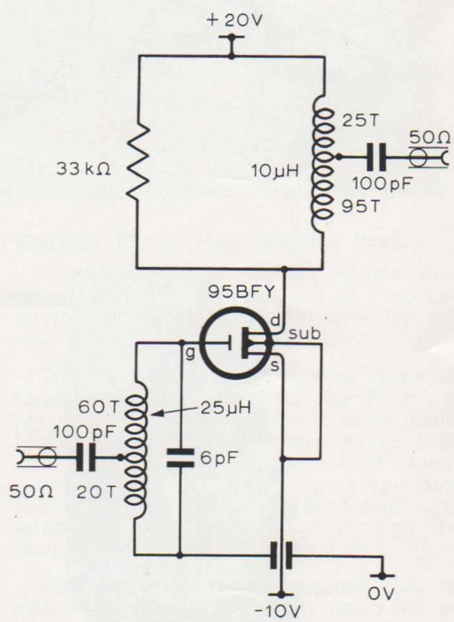



*Fig. 8 10Mc/s IF Amplifier.*

**10Mc/s IF Amplifier**

Another example of the application of the 95BFY is shown in Fig. 8.

The circuit has a bandwidth of 200kc/s and a gain of approximately 24dB. The source and substrate are connected to a negative supply to limit the drain current to 10mA.

The above brief examples of application merely serve to illustrate, in an introductory form, that the 95BFY M.O.S.T. lends itself to the already well-known conventional transistor circuitry.

With the additional advantages of $10^{12}\Omega$ typical input resistance, 4pF input capacitance, mutual conductance better than 1mA/V and a fourth electrode, the applications of the 95BFY reach well beyond that of the conventional transistor.■

R. ZENGER,
Applications Laboratory, Sydney.

## WARNING

Under no circumstances must the gate of the transistor be connected to any apparatus operated directly or indirectly by the mains supply; such as a soldering iron or test equipment unless the apparatus is effectively earthed.

If this precaution is not taken, the gate insulation is destroyed, rendering the transistor inoperative. This is due to the large input resistance of the 95BFY coupled together with the stray capacitance and induced voltages of the apparatus.

# HIGH FREQUENCY INDUCTORS

The modern trend in communications systems requires each route or link to carry a vastly increased number of channels, which means that frequencies higher than ever before are now being utilised. To cater for 600, 1,200 and 2,400 communication channels in one system, carrier frequencies are now used in the range of 2 to 12 Mc/s.

Frequency selective circuits in this range require compact, high quality, stable inductors and these can be manufactured from the Mullard Blue Range of Vinkor adjustable pot cores.

Inductor design in the 2 to 12 Mc/s range involves somewhat different problems than are encountered at lower frequencies. The type of wire (essentially stranded or bunched), the type of bobbin and the position of the wire within the bobbin assume paramount importance in relation to the inductor 'Q' factor. The principal losses that reduce the 'Q' factor of the inductor are shown below.

**DC WINDING LOSS**

$$\tan \delta_o = \frac{R_o}{\omega L} \qquad (1)$$

where $R_o$ = DC resistance of winding.

**EDDY CURRENT LOSS IN THE WINDING**

The principal eddy current loss in the winding is that due to proximity effect. The loss angle is given by

$$\tan \delta_{pe} = \frac{K_e f d\ 4\ nN}{\mu_e} \qquad (2)$$

where d = bare diameter of strand in cm
n = the number of insulated strands in the conductor (= 1 for solid wire)
N = number of turns
$\mu_e$ = effective permeability
$k_e$ = proximity effect constant.

The proximity effect constant is a function of the core and winding geometry and the value of $\mu_e$.

**DIELECTRIC LOSS DUE TO WINDING SELF CAPACITANCE**

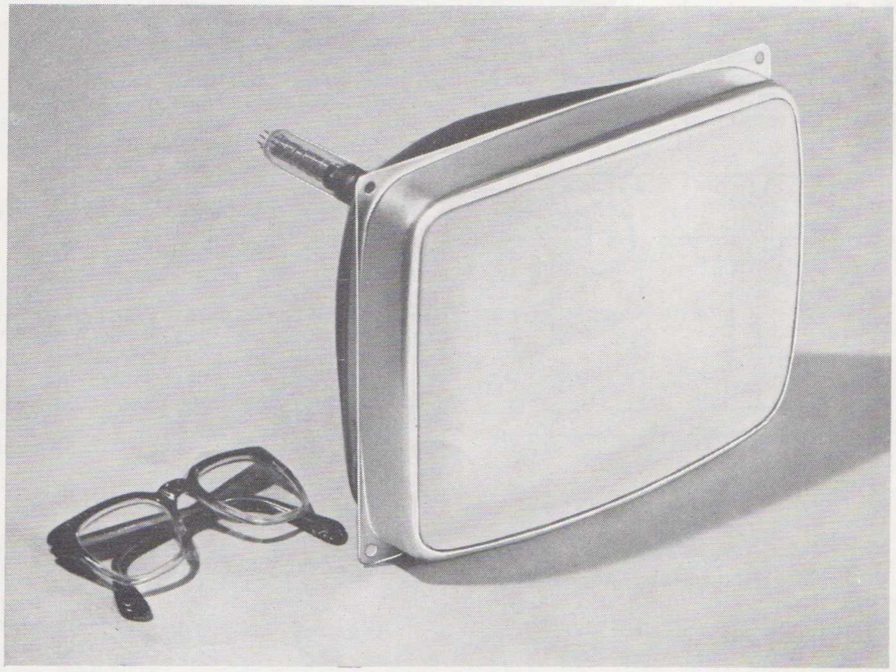
The loss factor is given by

$$\tan \delta_{Cd} = \omega^2 LC_s \tan \delta_d \qquad (3)$$

where $C_s$ = total self capacitance of the coil
$\tan \delta_d$ = loss factor of associated dielectric.

→page 24

# NEW COMPONENTS FOR PORTABLE TELEVISION RECEIVERS

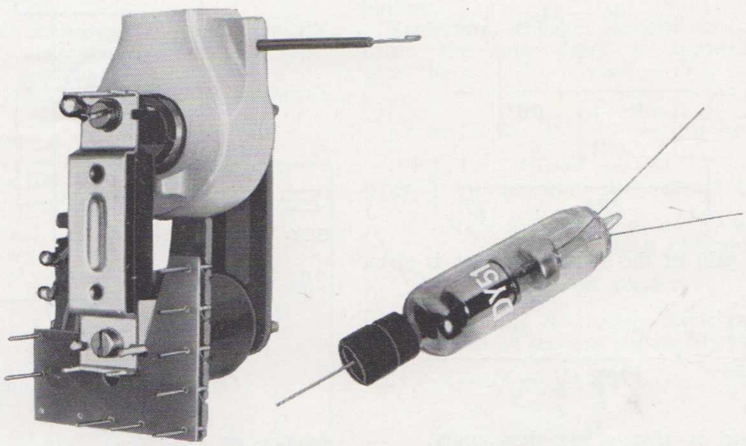


The A28-13W is a rectangular, 11 inch, long-life Panorama television picture tube which features an aluminised, grey glass almost flat screen. Because of the deep-drawn metal-rim reinforcement, it may be used without additional face-plate or safety glass.

This screen size was selected for good entertainment value commensurate with a sensible ratio of picture tube size and portability.

The A28-13W is designed for an EHT voltage of 11kV and this factor, together with a deflection angle of 90°, results in very little deflection power being required. The narrow, 20mm neck (as against 28.6mm for conventional picture tubes) reduces the required deflection power by a further 30% and this low power consumption, together with a voltage swing at the cathode of only 45V for a good contrast picture, renders the A28-13W eminently suitable for transistor drive.

The heater ratings are 11V ± 15% at 68mA. The heater voltage may be derived from a conventional power transformer, a storage battery or from a winding on the horizontal output transformer.



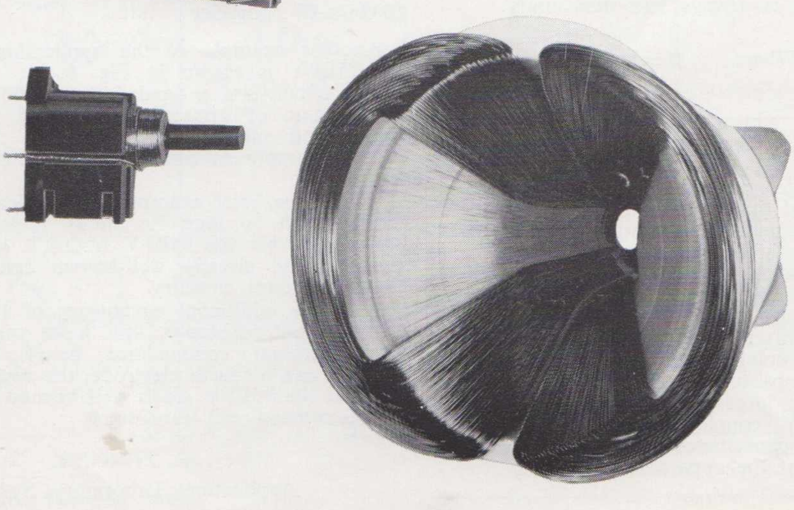

**Horizontal Output Transformer**

The horizontal output transformers may be obtained for use with valves (AT2043) or for transistor operation (AT2042). The terminals on the transformers are brought out to one side and shaped in such a way that the transformers may be used with a printed wiring board.

**Deflection Yoke**

The high-efficiency deflection yokes AT1020 (for use with valves) and AT1021 (for use with transistors) are designed to be used with the long-life Panorama picture tube A28-13W and feature high deflection sensitivity and light weight, together with a very low temperature rise.

The horizontal linearity coils AT4037 (valves) and AT4036 (transistors) are designed for use in conjunction with their relevant horizontal output transformers and deflection yokes.



**EHT Rectifier Valve**

The DY51 is an indirectly heated (1.4V at 0.55A) EHT rectifier valve with flying leads, a maximum peak inverse voltage of 15kV and a maximum peak anode current of 350μA. Because of its high reliability and light-weight construction it may be soldered directly onto the printed wiring board.■

## MAILING LIST

If you change your location, don't forget to let us know in good time, otherwise your Outlook may reach you late or never. And please, when you change, quote both your old and your new address. We can then be sure of destroying the obsolete mailing plate.

# SILICON PLANAR INTEGRATED CIRCUITS

*The story behind the development and manufacture of integrated circuits is briefly outlined in this article by a simplified step by step description of the manufacturing process.*

One of the most remarkable features of integrated circuits is their extremely small size (approximately 1 sq mm) making possible the manufacture of complex equipment in very small lightweight packages. Another feature of the integrated circuit is the small number of external connections which have to be made by the equipment manufacturer, thus reducing the assembly cost and improving the reliability.

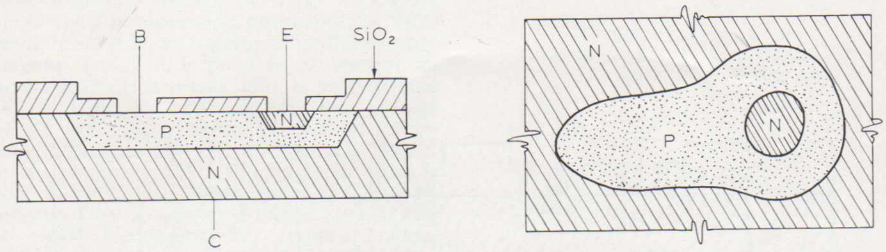



*Fig. 1. Construction of silicon planar transistor.*

Integrated circuits are most commonly manufactured by a lithographic pattern technique on a silicon substrate, coupled with a metal-over-oxide technique, to form a network of interconnected passive and active elements, diffused in the same substrate. This is known as the "planar" process.

**Basic Principle of Construction**

The "planar" process consists of layers of P-type or N-type additive, diffused into a slice cut from silicon crystal. The diffusion pattern or windows are obtained by photo-chemically etching an oxide layer previously grown over the upper surface of the substrate.

This oxide layer is an important feature of the silicon planar process as it provides a hard protective coating to the junction. It is interesting to note that in order to retain this protection during manufacture the oxide is allowed to reform after each etching and diffusion process step.

Fig. 1 illustrates the construction of an N-P-N transistor or an N-type substrate. It will be noted that two successive diffusions are required in order to form the base and emitter, and hence two subsequent growths of oxide layer and window etching.

**Circuit Isolation**

If several diodes and transistors are formed on a common N-type substrate the diode cathodes and transistor collectors would be connected via the substrate. Therefore, in order to create an integrated circuit it is necessary to isolate the elements. This is achieved by using a silicon P-type substrate onto which is grown an epitaxial N-type layer. The P-type substrate with its N-type epitaxial layer is subjected, via etched windows in the oxide layer, to a deep P diffusion reaching down to the P substrate. In this way a number of N-type islands are obtained (see Fig. 2).

Isolation between islands is maintained by reverse bias of the P-N diodes thus obtained, however due to leakage and depletion capacitance complete isolation is not possible.

**Epitaxial Planar Manufacturing Process**

A discrete circuit element is now processed within each island and the required interconnection, in order to form the circuit, must now be provided for.

This is achieved by etching windows through the oxide layer to the P and N regions to which connection is to be made. Metal is then evaporated over the windows and micro-alloyed into the crystal. Interconnection is achieved by evaporating aluminium over the oxide insulating layer through a photo mask, thus forming the desired circuit.

The planar transistor commonly has its collector connection on the underside of the substrate, however as it is necessary to use an insulating substrate and as interconnection is required to be evaporated on the top surface, it is necessary to bring the collector connection also to the top. This results in a long path through the N material resulting in high collector resistance. A well known method of reducing this undesirable resistance consists of making an additional N + (low resistivity) diffusion through the collector contact window prior to evaporating the metallic contact. Moreover, in those cases where low bottoming voltage is required an N + layer may be made through an appropriate window before the epitaxial layer is grown.

1. A silicon crystal containing a P-type impurity (Boron) is sliced with a diamond saw into wafers of approximately ·012″. The wafer is then lapped with a very fine abrasive and chemically etched to form an extremely smooth shiny surface. The thickness of the finished substrate is approximately ·005″.

2. The substrate is now placed in a furnace containing an oxidising atmosphere at 1200°C to form a surface film of silicon dioxide.

3. **Formation of the Collector Layer.** The oxidised substrate is coated with a photo-resist and exposed through a high resolution mask, the unexposed regions are then washed away and the substrate is etched in hydrofluoric acid to remove the silicon dioxide layer.

4. **N + Collector Diffusion**. After etching, the wafer is placed in a special high temperature furnace containing a phosphorous atmosphere. The temperature of the furnace is raised and the phosphorous impurity diffuses into the unprotected sections of the substrate forming an N + region of low resistivity (approximately ·002Ωcm).

→page 22

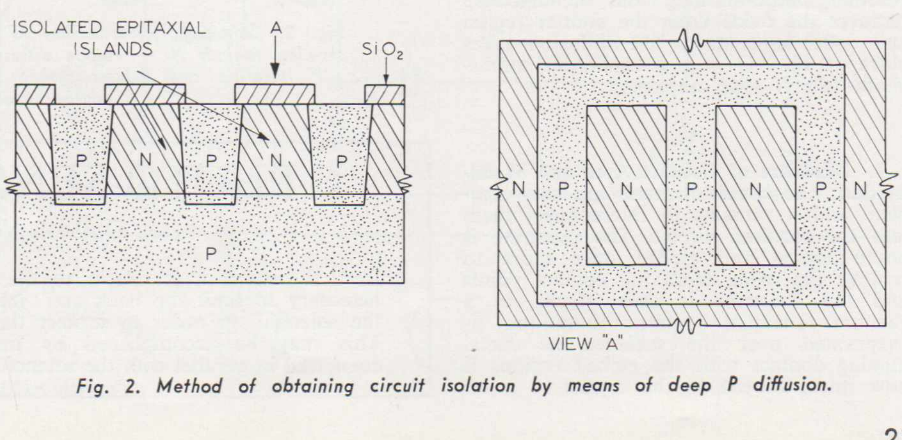



*Fig. 2. Method of obtaining circuit isolation by means of deep P diffusion.*

# SILICON PLANAR INTEGRATED CIRCUITS

← page 21

5. **Epitaxial Layer Growth.** Now that the N + layer is completed, the silicon dioxide layer is again removed from the substrate by etching with hydrofluoric acid, and is then placed in a furnace where an N type epitaxial layer is grown. This is achieved at a temperature of 1250°C when a mixture of silicon tetrachloride ($SiCl_4$) and hydrogen is allowed to flow over the surface. The reduction mechanism allows the slow growth of silicon onto the substrate and conditions are usually chosen so that the layer grows at a rate of 1 $\mu$/min. In this way layers of from 1 or 2 to 50 $\mu$ can be grown with good control, and by adding an N type impurity at a controlled rate, an N type layer is built having the required resistivity, held within close limits. The resistivity of the epitaxial layer is usually about 1Ωcm.

On completion of the layer growth the substrate is again subjected to an oxidising atmosphere to form a protective silicon dioxide film.

6. **Circuit Isolation, Masking and Diffusion.** The substrate is again coated with photo-resist and again exposed through a photo-mask to form the circuit isolation, these are in the form of bands, as shown in Fig. 2. The unexposed portions of the silicon dioxide are again etched and the wafer placed in a furnace containing Boron atmosphere or P type dopant. The dopant diffuses through the N type epitaxial layer forming a highly concentrated P type region extending through to the P type substrate. Following this diffusion process the oxide film is allowed to regrow over the area and the substrate is now divided into isolated N type regions which will become the transistor and resistor areas.

7. **Base Masking and Diffusion.** The substrate is again masked and etched for simultaneous diffusion of the base regions and resistors. Once again Boron is used as the diffusion impurity, and after the diffusion process exposure to an oxidising atmosphere, passivates the substrate once more.

8. **Emitter Masking and Diffusion.** Another photo-masking and etching step removes the oxide from the emitter region and collector contact. The diffusion takes place at 1200°C in a phosphorous atmosphere (an N type impurity).

9. **Exposure of Contact Area and Metallisation.** The circuit elements are now complete and it remains to interconnect them into the required circuit. The substrate is photo-masked and etched once more to remove the oxide from the contact points and the substrate is then placed in a vacuum chamber where aluminium is evaporated over the surface, the metal making contact with the etched regions is now micro-alloyed to the crystal.

10. **Interconnections.** To this point the substrate has been subjected to a considerable number of process steps, however many substrates can be processed at a time and each wafer contains upwards of two hundred integrated circuits. The wafer is now cut into individual circuit chips ready for packaging, external connection and testing. All outside circuit connections are brought to contact pads on the periphery of the chip to which gold wires are thermo-compression bonded.

Regardless of the complexity of the circuit, the integrated circuit must pass the same process steps as described, however in the interests of best possible production yield and, hence, lowest possible manufacturing cost, it may be advantageous to interconnect a number of basic integrated circuits mounted within the same package. In certain instances where perfect isolation is required and where resistors and capacitors having zero temperature and voltage coefficients are required, it may prove advantageous to use a combination of silicon planar and thin film technique within the same package. However current development will find the best combination of these techniques to provide the ideal balance between cost, performance and reliability.■

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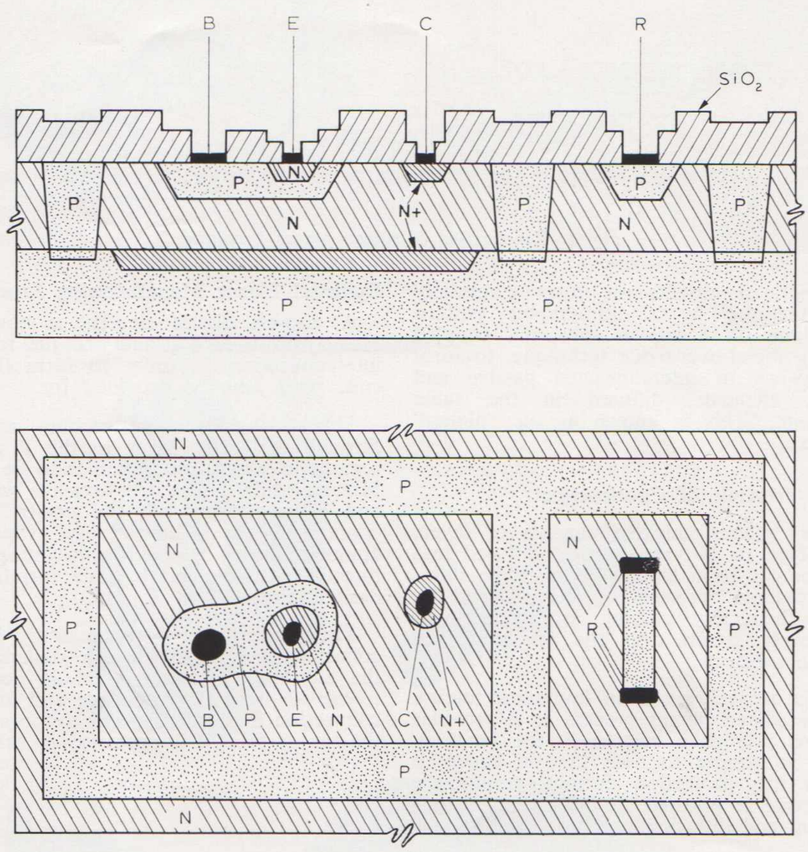


*Fig. 3. Schematic cross-section of a processed silicon transistor and resistor showing regions N + region diffused in P substrate, N + collector contact, P isolation and P-type resistor element. This diagram also illustrates micro-alloyed metallic contacts.*

## PHOTO-SENSITIVE RELAY

(see Outlook Volume 8, Number 1)

Readers' attention is drawn to the fact that with some types of relay it is necessary to limit the back emf, induced when current ceases to flow through the solenoid, in order to protect the switching transistors.
This may be accomplished by installing a Mullard OA91 diode, reverse-connected in parallel with the solenoid.

# VOLTAGE CONTROLLED OSCILLATORS

*In this article two voltage controlled oscillators (VCO) are discussed; one based on a Miller integrator and the other on a free-running multivibrator.*

A voltage controlled oscillator — VCO — is defined as one in which the frequency is a function of the input voltage and the output voltage of the oscillator takes the form of a sine wave, a square wave, a saw tooth or any other periodic function. A typical example of a VCO may be found in the form of the frequency controlling stage in an FM transmitter where the deviation of the output frequency is an almost linear function of a relatively slowly varying input voltage (audio).

In industrial electronics, linear VCOs are employed for measuring purposes, the VCO being used for an analogue-to-digital conversion. Today it is possible to convert all practical measurements into a proportional DC voltage. By applying this DC voltage to the input of a linear VCO a frequency proportional to the DC voltage will appear at the output — i.e., the measured value is now available in digital form and can be registered by a counter. Furthermore, it can be easily transmitted for remote reading which can be in digital form or, after demodulation, in analogue form. This application will normally require a VCO of very low frequency (2 to 5c/s) and a large deviation (1–10 or more).

**VCO Based on Miller Integrator**

A Miller integrator, with the transistors $TR1_A$, $TR1_B$ and TR2, is used to achieve a low frequency. $TR1_A$ and $TR1_B$ are connected to form a high-gain Darlington pair while TR2 is an emitter follower from which the feedback is taken.

The Miller integrator is capable of supplying a very long, linear sweep with small values of R and C. Assuming the voltage gain is A, the input voltage E and the output voltage $E_s$:

$$E_s = AE\left(1 - e^{-\frac{t}{RC(1-A)}}\right)$$

$$\simeq -\frac{E\cdot t}{RC}\left(1 - \frac{t}{2RC|A|}\right)$$

$$\simeq -\frac{E\cdot t}{RC} \text{ for } RC|A| \text{ large}$$

It can be seen from this equation that the capacitor acts as if it had the value C(1−A) and, accordingly, its value can be kept small. With an increase in gain (A), the deviation from linearity will decrease.

An oscillation is obtained if the Miller integrator is reset when the sweep has reached a certain voltage. This oscillation has a frequency proportional to the input voltage, E, since the slope of the sweep will vary linearly with E (for $RC|A|$ large). Since the reset time is finite, an error will be introduced, as the percentage of the total sweep will be higher at high frequencies than at low frequencies. In order to reduce the reset time, the capacitor C can be kept small, as mentioned above. It is discharged through the output impedance of the emitter follower TR2 which must, therefore, have high gain. R1 must be kept as small as possible without losing too much voltage gain. The diode D supplies TR1 with a small positive bias preventing it from being cut-off during the reset cycle.

The complete oscillator consists of the Miller integrator and a feedback loop, a resetting voltage being fed back to the base of TR1 when the output voltage has reached a certain value.

The output voltage itself must be modified as, for resetting, a sharp pulse is required and this is obtained with the aid of a regenerative circuit. Furthermore, it is necessary to introduce a certain amount of delay in the feedback loop and this is achieved by the switching transistor TR3 and the monostable multivibrator TR4 and TR5.

When the output voltage has reached the zener voltage of Z this will conduct and switch on TR3. A positive-going pulse will appear at the collector of TR3 which, via $C_2$, will trigger the multivibrator. A negative-going pulse will appear at the collector of TR5; this pulse is fed back to the base of TR6 via $C_4$, causing it to conduct and →page 24

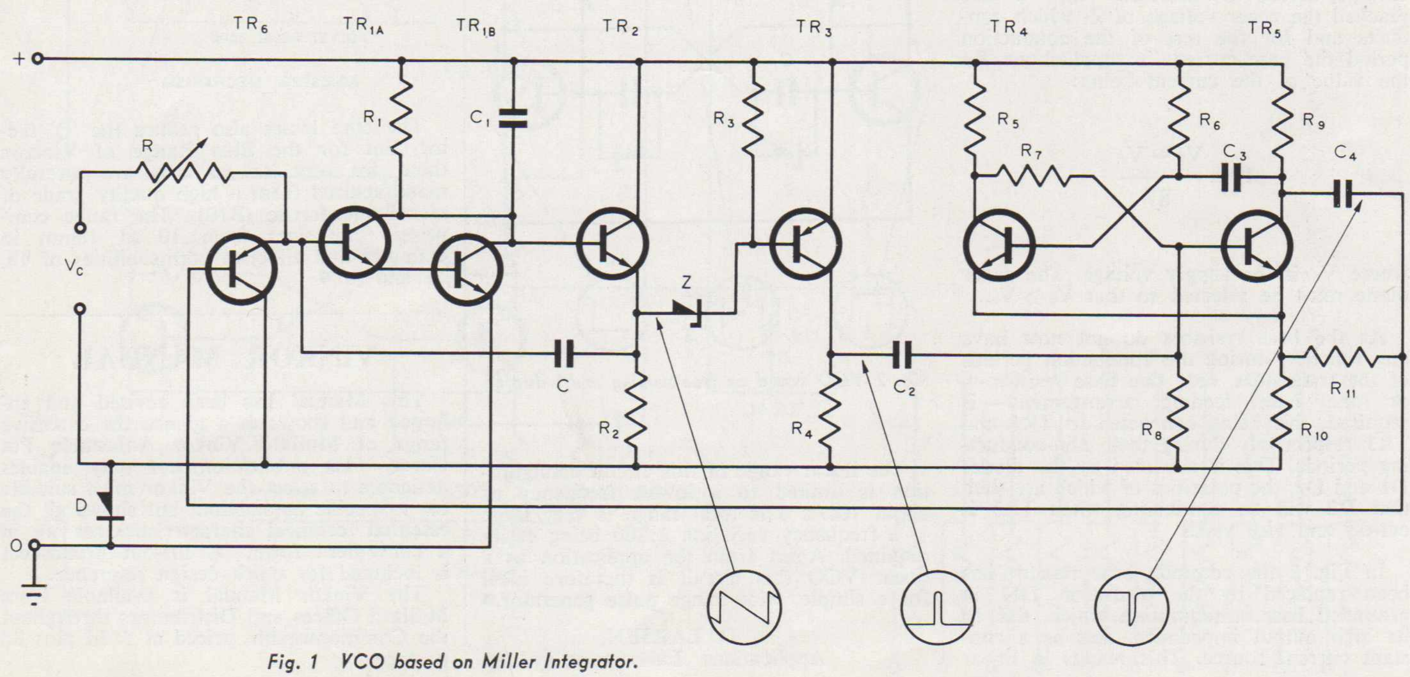


*Fig. 1 VCO based on Miller Integrator.*

# VOLTAGE CONTROLLED OSCILLATORS

← page 23

reset the Miller integrator. To ensure the complete discharge of C the multivibrator pulse width must be at least equal to the reset time of the Miller integrator.

In the range of 2 to 20c/s, an overall linearity better than 1% of the VCO is easily obtained, C being only of the order of 0.5 to 1.0$\mu$F.

**VCO Based on Free Running Multivibrator**

Another VCO which is very useful in applications where linearity at very low frequencies is not required, is shown in Fig. 2. Basically, this VCO is a free-running multivibrator, the frequency of which is controlled by the input voltage $V_C$. The period of a symmetrical free-running multivibrator is:

$$T \simeq 0.7 \cdot R_B \cdot C$$

where C is the cross-coupling capacitor ($C = C_1 = C_2$) and $R_B$ is the base resistor (this resistor does not exist in Fig. 2 for reasons explained later in this article). In order to achieve a low frequency $R_B \cdot C$ must be large. Since the capacitor required is of the order of 1$\mu$F, electrolytic capacitors are unnecessary and indeed should be avoided because of their large temperature coefficient. $R_B$ is limited by the fact that it must be small enough to supply the base current for the conducting transistor.

The limitation placed on the value of $R_B$ can be reduced by supplying base current for the conducting transistor via the opposite collector load resistance. This is achieved by introducing the zener diodes $Z_1$ and $Z_2$. Consider for example the charging of $C_2$ when TR3 is saturating: the voltage across $C_2$ increases until it has reached the zener voltage of $Z_2$ which conducts and for the rest of the conduction period the base current is supplied via $R_1$, the value of the current being:

$$I_b = \frac{V_b - V_z}{R_1}$$

where $V_b$ is the supply voltage. The zener diode must be selected so that $V_b > V_z$.

As the base resistors do not now have any function during the conduction periods of the transistors, only one base resistor — or one base feeding arrangement — is required, this being connected to TR1 and TR3 respectively during their non-conducting periods. This is achieved by the diodes D1 and D2, the polarities of which are such that D2 will be conducting when TR3 is cut-off and vice versa.

In Fig. 2 the common base resistor has been replaced by the transistor TR2 in grounded base configuration which, due to its high output impedance, acts as a constant current source. This results in linear charging of the capacitors $C_1$ and $C_2$ and the following equation is valid:

$$V_{CA} - \frac{1}{C_1} \cdot I_C \cdot \frac{T}{2} = 0$$

where $V_{CA}$ is the maximum voltage across the capacitor ($C_1$ or $C_2$). In this case $V_{CA}$ equals $V_Z$.

Hence

$$T = \frac{2 \cdot V_Z \cdot C_1}{I_C}$$

or

$$f = \frac{I_C}{2 \cdot V_Z \cdot C_1}$$

As the input impedance of TR2 is very small $I_C$ is essentially determined by the resistor $R_4$.

Hence

$$f = \frac{V_C}{2 \cdot V_Z \cdot C_1 \cdot R_4}$$

The multivibrator frequency is thus a linear function of the input voltage $V_C$.

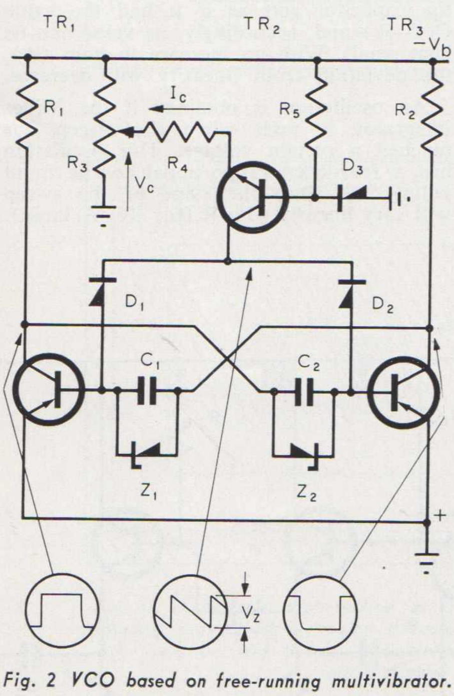


*Fig. 2 VCO based on free-running multivibrator.*

The linear range of this circuit configuration is limited to a lowest frequency of about 10c/s. The total range is very large — a frequency variation 1:200 being easily obtained. Apart from the application as a linear VCO this circuit is therefore ideal for a simple, wide range pulse generator.■

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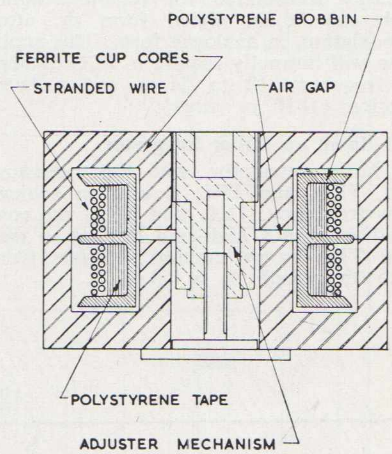
# HIGH FREQUENCY INDUCTORS

← page 19

It is apparent in (2) that as '$\mu_e$' is usually low ($\leqslant$40) and 'N' is small (<30), the strand diameter 'd' is the controlling factor and it is very beneficial to use the smallest gauge of wire available, even to the extent of using 50 s.w.g. In some instances it is also beneficial to use a 'Litzendraht' configuration of stranded wire. In reducing the strand diameter, it must be appreciated that the DC resistance (1) will be increased, and sufficient strands must be used to keep this part of the coil loss within reasonable limits.

It has been found that the leakage flux emanating from the gap in the centre of the pot core causes an increase in the eddy-current loss in the winding. This can be counteracted by keeping the winding as far from the gap as possible as shown in the figure. A suitable material for positioning the wire in the bobbin is polystyrene or melinex tape. The dielectric loss can be minimised by keeping the total self capacitance '$C_s$' (3) and its dielectric loss, small. The turns of the winding should be widely spaced, possibly by using multi-section polystyrene bobbins, and also positioned away from the centre of the pot core as shown in the figure.




The core losses also reduce the 'Q' factor, but for the Blue Range of Vinkors these are very low as they are carefully manufactured from a high quality grade of nickel-zinc ferrite (B10). The range comprises four sizes from 10 to 18mm in diameter and effective permeabilities of 16, 25 and 40.■

# VINKOR MANUAL

This Manual has been revised and enlarged and shows at a glance the extensive range of Mullard Vinkor Adjustable Pot Cores. This publication not only enables designers to select the Vinkor most suitable for a specific application, but shows all the essential technical characteristics set out in a convenient form. A lift-out broadsheet is included for quick design reference.

The Vinkor Manual is available from Mullard Offices and Distributors throughout the Commonwealth, priced at 5/3d plus 8d postage.

# DRIVE CIRCUITS FOR CRITICAL APPLICATIONS OF THE Z504S COUNTER TUBE

*Particularly economical circuits may be designed for the Z504S counter tube by using integrated pulse drive where the full performance of the tube is not required. However, there are situations where the tube must be driven at the upper limit of its frequency range, possibly with long standby periods, and it is imperative that the maximum possible life be obtained from the tube.*

As a result of a number of life test experiments, it has been found that if rectangular pulses are used for stepping the tubes, and certain voltage and current conditions are specified, that the life expectation is only limited by obscuration of the viewing dome (due to sputter deposition on the inner surface of the tube) at some 15,000 hours. Where visual indication is not required, a life in excess of 25,000 hours may be expected.

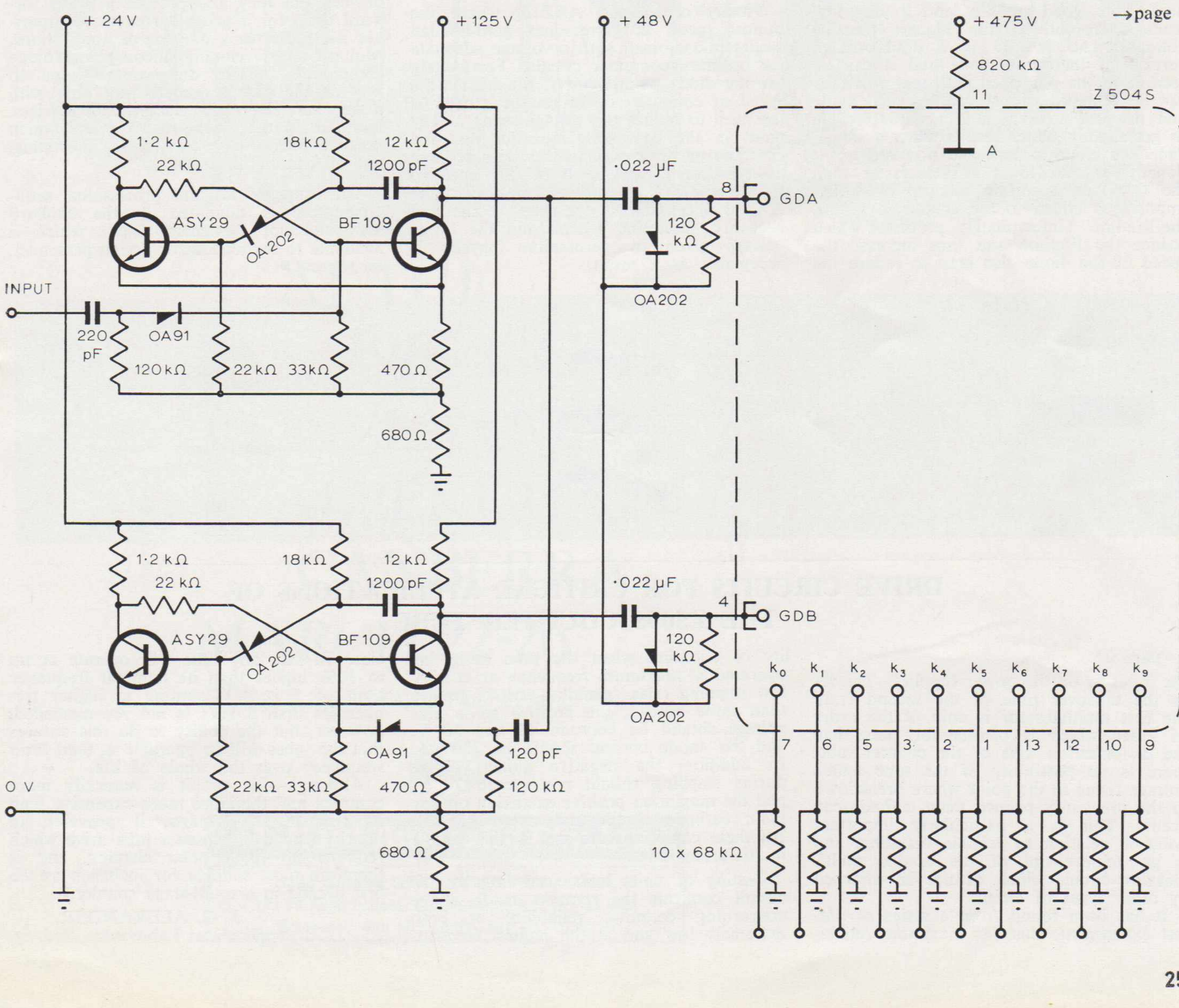
A circuit which meets all the requirements set by the tube when it is operated under stringent conditions is shown in the diagram, Fig. 1. It consists of two identical monostable multivibrators using N-P-N transistors. Each multivibrator uses one low voltage and one high voltage transistor. The high voltage transistors are powered from a + 125 V HT line, and their collectors are RC coupled to the respective counter tube guides via a DC restorer circuit which ensures that the guides return to their bias voltage of + 48 V during their rest periods.

The input signal triggers the first multivibrator, which has an input sensitivity of 5.5 V. Almost any signal which has a peak value greater than 5.5 V will trigger the circuit. The multivibrator when triggered produces a negative going output pulse which is coupled to Guide A on the counter tube. The positive (trailing) edge of this pulse is used to trigger the second multivibrator, this action taking approximately two microseconds (2 $\mu$sec) to complete. The second multivibrator then produces its negative output pulse which is coupled to Guide B on the counter tube.

There are a number of reasons for using monostable multivibrators in this drive circuit. Firstly an output pulse is produced which is accurately defined in voltage, width (time) and rise time, these terms being quite independent of the input signal. The shape of the pulse has an important bearing on the life to be expected from the tube, the life figures quoted above being obtained with rectangular pulse drive as produced by this circuit.

Secondly, the circuit can be driven by quite low input energies (5.5 V into approximately 25 k$\Omega$) but is nevertheless quite stable and unaffected by noise at

→page 26



# A SPECIAL PLACE FOR GOLD-BONDED DIODES IN THE SEMICONDUCTOR DIODE RANGE

When point contact diodes were the only germanium diodes available, the engineer had little difficulty in deciding what to use. There were then only two general types—high voltage and low voltage, the latter being rather faster than the former. Now, however, the user must choose between germanium point contact and germanium gold-bonded diodes and must also consider silicon alloy and silicon planar diodes, these followed later by micro-dot diodes and a complete range of similar new constructions.

**Gold-Bonded Diodes**

In the manufacture of gold-bonded diodes, the three main variables are: junction area, the resistivity of the original germanium wafer and the lifetime of the carriers in the material which affects the recovery time of the finished diode. Examination of the Mullard range of devices shows that the OA5 provides the highest hold-off voltage but is rather slow; the single-ended construction of the OA5 enables it to be etched after the junction is made and this gives a diode with good stability and a very low leakage current, at the highest possible voltage. The AAZ15 is a double-ended version of the same diode and it can be seen from the published data that whatever can be achieved with a double-ended diode and no post etching, it is always possible to get slightly better results when a single-ended envelope is used and post etching is done. The AAZ15 is a little faster than the OA5, as a certain amount of 'killer' impurity is added to the crystal to reduce the lifetime. Unfortunately, processes which reduce the lifetime and thus increase the speed of the diode also tend to reduce the maximum usable voltage. For even higher speed, the AAZ17 is a faster diode at an even lower voltage, supplied in double-ended construction. This diode achieves an acceptable compromise of voltage and speed.

**Fast Computer Diodes**

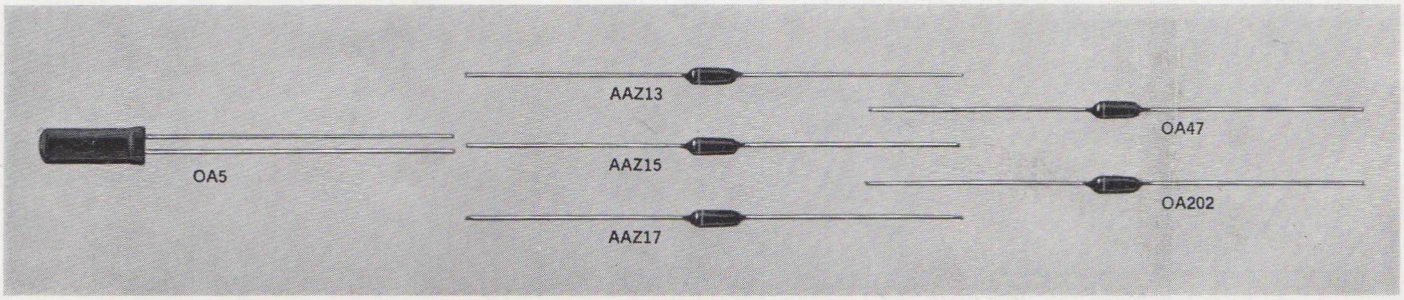
The OA47 has been available from Mullard for some considerable time and, at 25V and 280 pico-coulombs, is a good deal faster than the AAZ17, but of course has a lower maximum usable voltage. Most designers need a faster diode and 25V was quite adequate for transistor computers; but it was with the discovery that the OA47 at 25V was not fully satisfying the users' requirements that the AAZ17 was recently introduced. Some designers required a higher voltage diode—not because they wished to use the voltage but because they felt that the higher voltage gave them a greater safety factor.

Finally comes the AAZ13, where the utmost speed in germanium gold-bonded diodes is combined with a voltage adequate for ordinary computer circuits. Fortunately for the diode manufacturer, the higher the speed of computer operation the greater is the need to reduce the voltage swing at any point in the computer, in order that the charging of stray capacitances does not use up too much energy. It is this need to reduce the voltage swing that makes the AAZ13 acceptable where speed is the first necessity, price the second and the lower voltage given by germanium physics is acceptable as a result.

**Future Developments**

Turning now to future developments, if silicon alloy diodes are plotted on a similar curve to those of germanium, it will again be seen that it is possible to achieve the various compromises of voltage and speed, but that in general, for a given speed, the voltage is higher than in germanium gold-bonded diodes. The OA202 represents one point on this curve, and is a typical alloy, low-to-medium speed, silicon diode in the double-ended miniature envelope. With silicon planar diffused junctions it seems to be possible to achieve a different order of compromise of speed and voltage and to provide diodes with higher speeds for a given voltage or higher voltages for a given speed, than can be found with silicon alloy diodes, which are themselves a considerable improvement in this characteristic on germanium gold-bonded diodes. It must, of course, be remembered that germanium gold-bonded diodes are not only much cheaper than the present silicon alloy devices, but they also provide a lower forward drop for a given current—an important parameter in a number of applications. Mullard are developing silicon planar diode techniques and in due course hope to provide the circuit designer not only with newer and improved performance devices, but also to mass produce at prices which prevent them being treated as interesting, but uneconomical, "specials".

The range of Mullard professional semiconductors is tabulated in the Mullard Semiconductor Designers' Guide which is available to semiconductor design personnel, on request.■




# DRIVE CIRCUITS FOR CRITICAL APPLICATIONS OF THE Z504S COUNTER TUBE

←page 25

the input or on the power supplies. Thirdly as the takeover time of the second from the first multivibrator is only of the order of 2$\mu$sec which is considerably less than the de-ionisation time of the counter tube, there is no possibility of the tube anode voltage rising to the point where breakdown to the previously primed main cathode can occur. This is a particularly important point in relation to reliable counting, and is an achievement of the double multivibrator circuit which cannot be matched by other types of circuit.

It has been found from a series of life-test experiments that for maximum reliability of counting when the tube is to be operated at maximum frequency or at very low stepping rates (standby periods greater than three hours), the positive guide bias voltage should lie between 45 and 50 V, and the anode current should be 330 $\mu$A. In addition, the negative guide voltage during stepping should not exceed 70 V, and the maximum positive excursion on any main cathode should not exceed 35 V. All these conditions are met in this double multivibrator circuit.

Testing of units built according to this circuit confirms the remarks made above concerning counting reliability at both extremely low and at the highest frequencies. In fact any tube will operate at up to 50% higher than its nominal frequency limit of 5 kc/s. Counting at higher frequencies than 5 kc/s is not recommended, however, but the ability to do this ensures that the tubes will be operable at their limit frequency over the whole of life.

Although this circuit is naturally more complex and therefore more expensive than its simpler counterparts, it provides an utterly dependable counter tube drive which requires no input pulse shaping, and is therefore most suitable for inclusion as the first decade in a multi-stage counter.■

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# LADDIC IN COMPLETELY FAIL-SAFE FIRE ALARM

*With the increasing world usage of nuclear reactors, growing complexity of railway systems, and expansion of other fields where malfunction can cause disaster, the necessity of switching systems which fail-to-safety has become more and more vital.*

In a normal switching system, the output can have two states, either on or off. If there is an internal failure, be it power supply, component or connection, then the output may remain in either the on or off state, depending on the type of failure.

A fail-safe system is a system specifically designed for use in critical situations, so that any conceivable internal failure will result in an "unsafe" condition. Where possible this failure to safety is extended to include any input devices associated with the system.

In designing a system which will fail safe, it is nevertheless important to ensure that unnecessary shut downs are reduced to a minimum by selection of the most reliable components. In most situations the consequences of a shut-down due to a failure in the control system can be quite serious and costly.

With these requirements as a guide, the Laddic[1] has been developed as a device particularly suited to performing logical functions in fail-safe systems. The Laddic is a multi-aperture ferrite core with a square hysteresis loop. Its shape has been evolved to provide maximum resistance to breakages, and to provide failure to safety in the event of a breakage. The resulting core bears resemblance to a ladder, from which its name was derived (LADDer logIC).

Considerable experience has been accumulated over the last decade with the use of square hysteresis loop ferrite cores in computer memories, and no measurable deterioration has been observed in their characteristics in a period of over ten years. With this evidence of reliability in its favour, a magnetic device such as the Laddic is clearly well suited to any application where ultimate dependability is essential.

Operation of the Laddic[2] can be explained with reference to the diagram, Fig. 2. A current pulse is passed through the set winding, saturating the core. The direction of flux is upwards on rungs A, C, E, G and J, and downwards on rungs B, D, F, H and K. The set and drive pulse generators are interlaced, both being locked to the mains for convenience, thus a drive pulse follows the set pulse. This drive pulse reverses the flux in rungs A and D as this represents the shortest path length, and no other rungs are affected.

If, however, the winding on rung D (called a HOLD winding for reasons which will be apparent below) is energised at the same time as the drive pulse, then this HOLD winding will prevent the flux from being switched in rung D. The switching mmF will then select the next shortest path, which will be rung F if its HOLD winding is not energised. The flux cannot switch in rungs B, C, E, G or J because they have already been saturated in the same direction by the set pulse.

If all HOLD windings D, F, and H, are energised, then the flux path will be via rungs A and K and the flux in rung K will be switched. When the next set pulse arrives, the flux in rung K will be switched again, producing an output in the winding on K. A small output voltage is produced due to the drive pulse, but as the set winding has a total of five turns, the output due to this pulse is considerably larger.

To produce an output, then, all the HOLD windings must be energised and both drive and set pulse generators must be operative. Thus the Laddic produces the AND function. If a number of windings are placed on any HOLD rung, an OR function is obtained as energising any one of such windings will prevent flux switching in that rung.

It is clear that if any of the HOLD drive or set signals fail, then no output will be produced. The core can only break in such a manner as to isolate the drive and output windings, thus no output will be produced in this event. To enable a fault condition to be detected, the Laddic is switched continually by the drive and set windings. Thus a safe system is represented by a train of output pulses and any failure will stop these pulses.

An example of a problem for which Laddic can provide a useful solution is a fire alarm warning system. For general installations, it is sufficient to provide protection against lamp failure where light is used for smoke detection. A typical circuit using Norbits is shown in Fig. 1.

Any light falling on the photocell (arrowed) is indicative of the presence of smoke. To provide discrimination against stray dust particles, a neon lamp is operated by the photocell when the smoke has reached the intensity at which it is desired to operate an alarm. A second photocell is operated by the neon lamp and acts as

→page 28

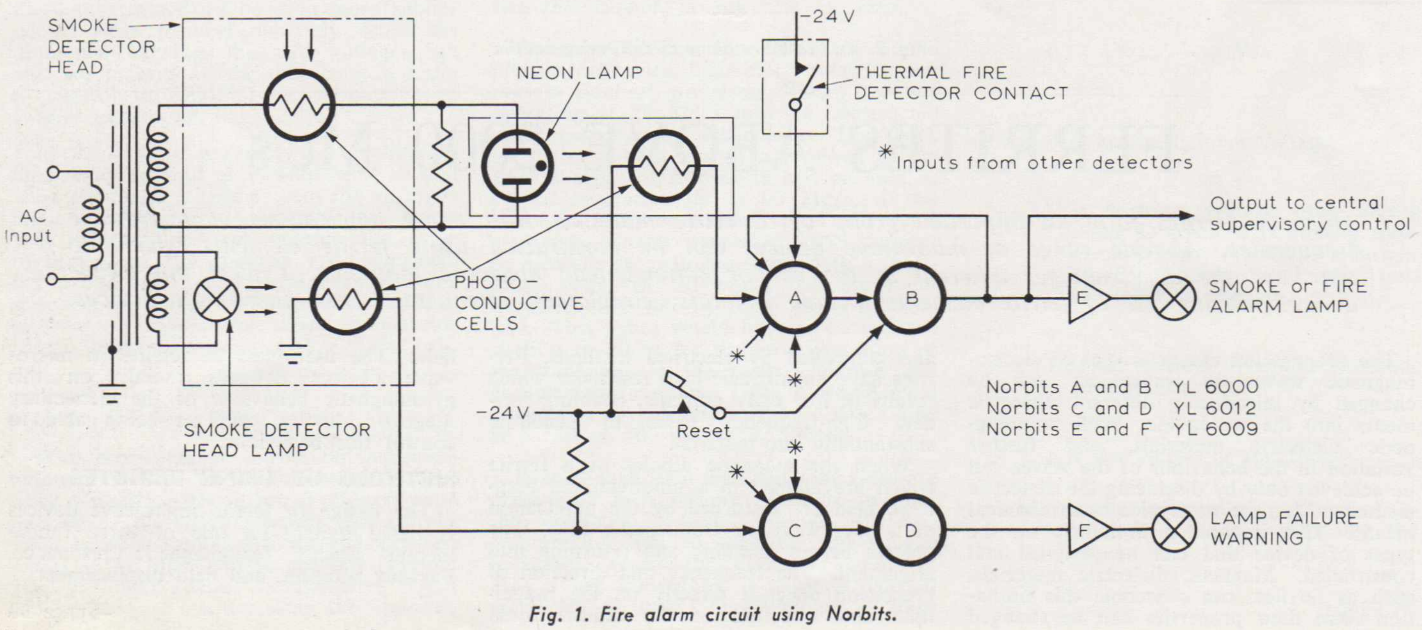


*Fig. 1. Fire alarm circuit using Norbits.*

## LADDIC IN COMPLETELY FAIL-SAFE FIRE ALARM

← page 27

input to a Norbit memory circuit. This memory circuit lights a lamp and provides an output to a remote supervisory point. The memory circuit is incorporated so that a fleeting alarm, such as a wisp of smoke, causes the lamp and alarm signal to come on and remain on even if the smoke subsequently disappears. The circuit can only be restored to normal by pressing the 'reset' button. A lamp test circuit is included by the use of a third photoconductive cell. Provided the lamp is on, the Norbit circuit remains off. If the lamp fails, however, the Norbit circuit energises a warning lamp.

In some critical situations, however, the protection against failure afforded by the Norbit system is insufficient and complete failure to safety is indispensable. In this case the Laddic can be used very effectively as shown in Fig. 2.

For fire and smoke detection, two pick-up heads are used. The fire head consists simply of a pair of normally closed contacts operated by a bi-metal strip. Excessive temperature rise or open circuit connection to the head will cause the amplifier to de-energise the HOLD winding on rung H, thus producing an alarm condition by inhibiting the output pulses.

The lamp detector circuit is incorporated in the circuit to guard against failure of the smoke detector lamp or power supply. Any such failure will de-energise HOLD winding on rung F and stop the output pulses.

The smoke detector head consists of a diffusion chamber containing a lamp and photo-conductive cell. The cell does not normally receive light, but in the presence of smoke, light is reflected from the smoke particles onto the photocell, causing it to operate and in turn de-energise HOLD winding D.

In this example, all the HOLD windings are DC energised, but they may be pulse controlled in synchronism with the drive pulses, in more sophisticated systems.

When no alarm is initiated and the system is functioning correctly, a train of pulses appears at the output winding of the Laddic, which is coupled into an output amplifier. This amplifier uses transistors in a transformer - coupled circuit to drive the output load. Due to the presence of the transformer, any open circuit or short circuit failure in the output amplifier will remove the final output signal, thus the amplifier can be considered fail-safe in itself.

It can be seen that each part of the circuit is inherently safe, so that the system described here provides a completely fail-safe fire alarm system. The same principles can be used to extend the application to the control of large and complex systems whose reliability is vital.■

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1. Gianola and Crowley: The Laddic — A Magnetic Device For Performing Logic. **Bell System Technical Journal,** January, 1959.
2. Hardy: Fail-Safe Logic Using Multi-Aperture Ferrite Cores. **British Communications and Electronics,** March, 1962.

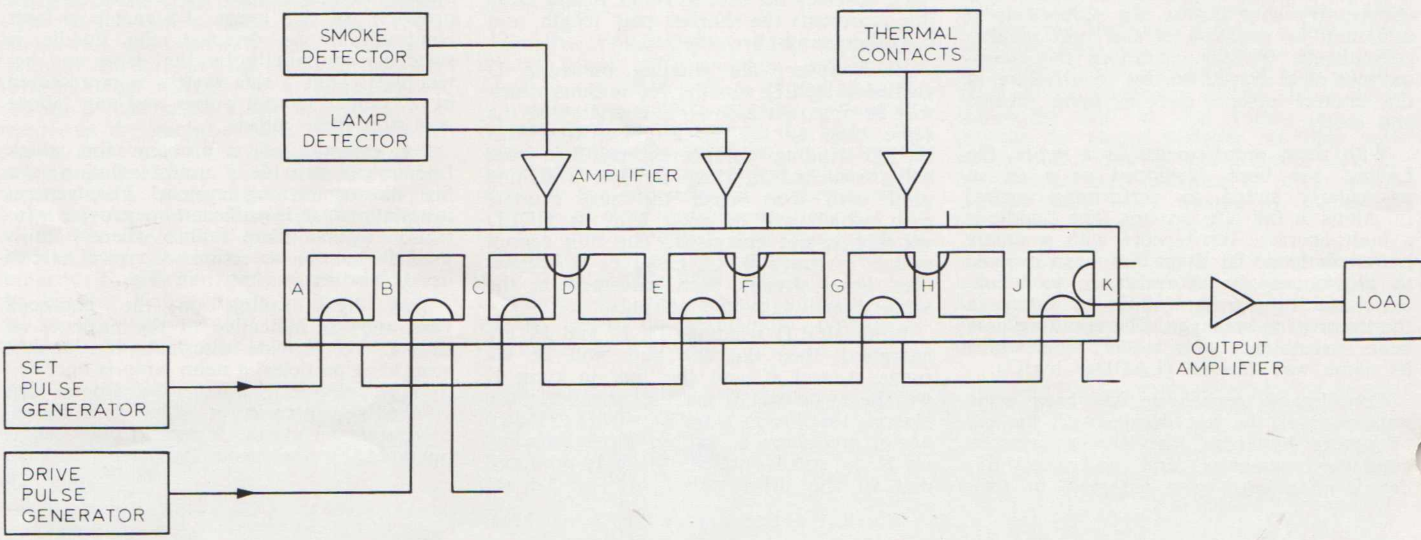


*Fig. 2. Fail-safe fire alarm circuit, using Laddic*

# FERRITES ABOVE 2000 Mc/s

*Magnetic ferrites form an important group of dielectric materials with many applications at microwave frequencies. A wide range of microwave devices can be constructed using ferrite elements which do not have the disadvantages inherent in the use of non-magnetic dielectric materials. Typical microwave devices making use of ferrite elements include isolators, circulators, attenuators, and phase-shift devices.*

The propagation characteristics of electromagnetic waves in waveguides can be changed by introducing different dielectric media into the waveguide. With non-magnetic dielectric materials, any further variation in the behaviour of the waves can be achieved only by displacing the dielectric element within the waveguide by mechanical means. This imposes a limitation on the types of device that can be designed and constructed. Magnetic dielectric materials, such as ferrites, can overcome this limitation since their properties can be changed and controlled by electrical methods. Ferrites have an intrinsic high resistivity which results in low eddy currents, enabling incident high-frequency fields to penetrate substantially into material.

When the magnetic dipoles of a ferrite which are aligned in a unidirectional magnetic field are disturbed by the application of a microwave electromagnetic field, they precess before damping and returning into alignment. The frequency and direction of precession depends directly on the magnitude and direction of the unidirectional field. The usefulness of ferrites in microwave devices depends mainly on this gyromagnetic behaviour of the elementary magnetic dipoles, and on being able to control this behaviour.

### METHODS OF USING FERRITES

The design of ferrite microwave devices is based mainly on one of three fundamental effects; ferromagnetic resonance, Faraday rotation, and field displacement.

→page 33

# ANALOGUE-TO-DIGITAL CONVERTERS USING SILICON PLANAR TRANSISTORS AND DIODES

There are two classical methods of deriving digital information from an analogue source. The more common method consists of a number of constant current generators which share a common resistor as load. The voltage which is generated across the resistor is then compared with the input analogue voltage. If the input is larger than the generated voltage, more current generators are switched into operation, but if the input is smaller than the generated voltage, some of the generators are removed from the circuit. When the input voltage is equal in amplitude to the generated voltage, the number of current generators in use is a digital measure of the analogue input.

The other method of performing an analogue-to-digital conversion is to generate a sawtooth voltage equal in amplitude to the analogue quantity being measured. A gate is then opened for the time taken for the sawtooth or ramp to reach this voltage and the number of pulses passing through this gate from an accurate pulse generator or clock during this time are counted, thus providing a digital quantity proportional to the original analogue voltage. If the slope of the ramp and the clock rate are correctly chosen, the constant of proportionality will be one and the digital output will equal the analogue input.

### Voltage Comparison Techniques

In both cases, the analogue input must be compared with an internally derived voltage and some action taken when the two are equal. The comparison process is usually effected by a chopper—either mechanical or electronic—and it is this process which defines the resolution capabilities of the converter. Mechanical choppers can resolve down to 2·5 $\mu$V while current electronic choppers may be used down to 2 $\mu$V. In the current generator converter the comparator must be even more sophisticated since it must not only detect the difference between the two voltages but also the polarity of the difference in order to control the addition or subtraction of current generators.

In the case of a ramp type converter a diode may be used as a comparator so that when the ramp voltage exceeds the analogue input, current flows through the diode. This system does not provide the same resolution as the method just described, but is far more economical even if some selection of diodes is required. If such a selection is made, the diode comparator should detect differences smaller than 1 mV.

### Selecting Sampling Rate

With any form of conversion, a certain amount of analogue information is lost since we sample the input at discrete time intervals. The system records the average input between sampling times and any fast pulses at the input will be virtually ignored. This drawback cannot be eliminated but it can be reduced by increasing the sampling rate. A few simple calculations show that for the ramp converter two equations must be satisfied. These equations are

$$\text{and } E = \frac{KC}{r} \qquad (1)$$

$$r = \frac{C}{N} \qquad (2)$$

where E is the maximum analogue voltage to be converted.

C is the clock rate in pulses per second.

N is the maximum count capacity of the counter following the converter.

r is the sampling rate.

K is the input voltage corresponding to one count into the output counter.

A little manoeuvring of these expressions shows that if the sampling rate is to be increased we must either reduce the capacity of the counter following the converter or increase the clock rate. Reducing the counter capacity means that the indication will be in coarser steps and the extra information gained by increasing the sampling rate cannot, in any case, be read.

Fortunately, since the advent of silicon planar transistors, high clock rates do not present much of a problem. Binary counters operating at 20 Mc/s may be constructed using conventional techniques. The clock oscillator is even less of a problem as the BSY38 series of transistors can be used as crystal oscillators up to 60 Mc/s. If we select 10 Mc/s as a clock speed and use a 4 decade counter (capacity 9,999 counts) to measure an analogue signal with a maximum value of 10 volts, the maximum sampling rate will be 1,000 times per second. This figure would have to be reduced slightly to allow for the switching time of the transistors in the ramp generator and ensuing logic circuits. Thus, it would seem reasonable to generate a ramp 1 msec long at a rate of 100 ramps per second.

Even though such high sampling rates are practicable it is usually desirable to filter the analogue quantity through an R-C combination and to measure the voltage appearing across the filter capacitor. If this is not done, and a constant current converter is used, the analogue voltage may "cross over" the internally generated voltage, with the result that the system will be continually chasing the input and may never catch up and so no output will be obtained. If the capacitor is not used with the ramp converter, there will be no alternate path for the current normally flowing through the coupling capacitor. This system will then also be unable to switch off.

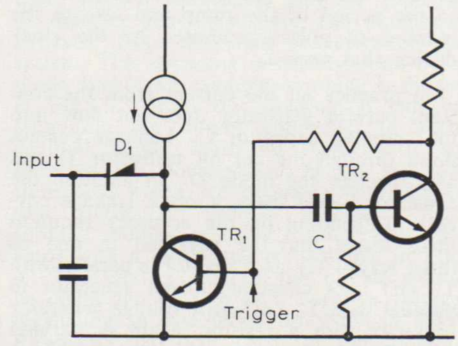


*Fig. 1. Skeleton diagram of a typical sawtooth generator.*

### Ramp Generators

There have been several articles on transistorised ramp generators in recent publications. A typical example is shown in outline in Fig. 1. This circuit is basically a monostable multivibrator in which $T_1$ is normally conducting and so absorbs all the current from the constant current generator. When a trigger pulse arrives at the input, $T_1$ turns off and the current from the generator passes through C into the base of transistor $T_2$ causing $T_2$ to turn on and so keep $T_1$ turned off. As the current passes through C the capacitor becomes charged and the circuit will revert to its stable state when

$$E + \Delta V = \frac{It}{C} + V_{be2} \qquad (3)$$

where E is analogue input voltage.

$\Delta V$ is forward voltage drop across diode D when it is passing sufficient current from the constant current generator to starve $T_2$ so causing it to turn on.

I is the current from the constant current generator.

t is the period of the ramp.

$V_{be2}$ is the base emitter voltage of $T_2$ when 1 mA are flowing into the base.

→page 30

# ANALOGUE-TO-DIGITAL CONVERTERS USING SILICON PLANAR TRANSISTORS AND DIODES

← page 29

If we arrange $\Delta V$ to be equal to $V_{be2}$ we arrive at the expression

$$E = \frac{I}{C} \times t$$

so that the analogue input is proportional to the period of the ramp and also to the number of pulses generated by the clock during that period.

In practice all the current from the constant current generator does not flow into the coupling capacitor C. Leakage current flows through the cut off transistor $T_1$ and also through the diode D. In addition the capacitor C will have a small leakage current. Depending on the accuracy required these losses may be regarded in one of three ways—(i) as a resistor in parallel with $T_1$, (ii) as a constant current generator in parallel with $T_1$, (iii) as a voltage generator in series with a resistor, again in parallel with $T_1$. No matter which approach is adopted, it should be borne in mind that the losses are temperature-dependent and that some form of temperature compensation will be necessary. It is usually convenient to apply this compensation to the constant current generator.

Instead of trying to equate the voltage drop across the diode and the transistor base emitter voltage as indicated above, we can arrange for both terms to be eliminated by incorporating two ramp generators in the analogue-to-digital converter. Both generators are identical, except that one has the analogue input as a reference and the other has a zero volt reference. In this case the analogue voltage is proportional to the time interval between the end of one ramp and the end of the other ramp. In addition to eliminating some of the errors in the converter, the double ramp enables input voltages of either polarity to be measured and also the polarity may be indicated by using a suitable indicator tube

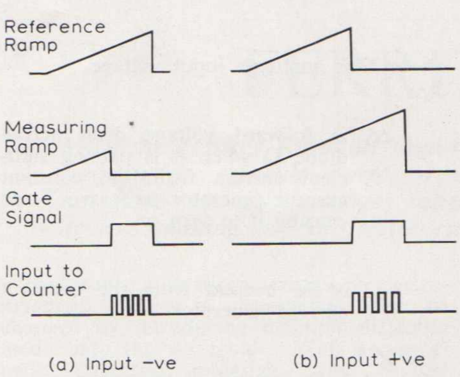


*Fig. 2. Timing sequence for double ramp type converter.*

in the counter. The waveforms which may be expected from this circuit for both polarities are shown in Fig. 2.

### Logic Circuits

If the double ramp generator is selected, a suitable logic network must be used to ensure that the gate opens for the correct period and at the right time. This network will have to be an exclusive "OR" gate and perform the Boolean function A B′ + A′ B. A suitable skeleton circuit is shown in Fig. 3. Transistors $T_4$ and $T_5$ provide definite voltage levels to the switching transistors $T_6$ and $T_7$ and also act as buffers to isolate the ramp generators from the switching circuit. As soon as the ramp ends, transistor $T_4$ or $T_5$ will go into conduction. Since both ramps start together,

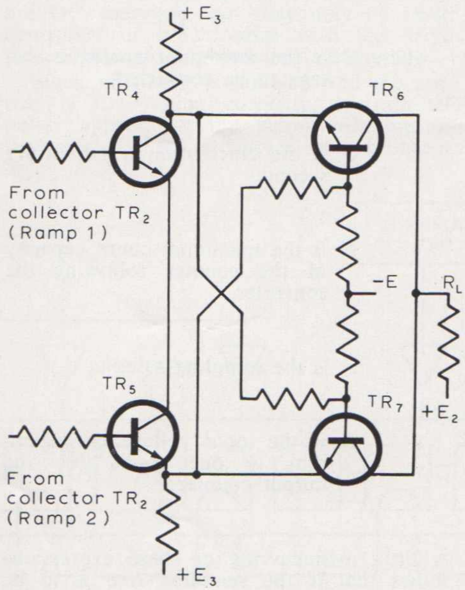


*Fig. 3. Exclusive "NOR" gate.*

the circuit may be proportioned to give the following truth table where the symbol 0 signifies non-conduction and the symbol 1 indicates conduction.

| $T_4$ | $T_5$ | $T_6$ | $T_7$ | Output Level |
|---|---|---|---|---|
| 0 | 0 | 0 | 0 | $E_2$ |
| 0 | 1 | 1 | 0 | $E_3$ |
| 1 | 0 | 0 | 1 | $E_3$ |
| 1 | 1 | 1 | 0 | $E_2$ |

The designer must be careful to avoid exceeding the reverse emitter-base voltage ratings on transistors $T_6$ and $T_7$. This may call for the collector supply to $T_4$ and $T_5$ to be reduced to a low level, say of the order of 3 V. This is made more acceptable by the low saturation voltage of the silicon planar transistor. These transistors have such a small desaturation time that it may not be necessary to clamp the collectors of the output pair. Since the propagation delay times of transistors $T_4$ and $T_5$ effectively cancel one another, it should also be unnecessary to clamp these. The output pulse across $R_L$ is then used to gate the counter or clock.

### Constant Current Generators

It will have become obvious by this time that the linearity and overall accuracy of any analogue-to-digital converter depends for the most part on the linearity and stability of the constant current generator employed. To a first approximation, a transistor with a large emitter resistor makes a good constant current source. However, to operate accurately over a wide range of temperatures, a more sophisticated approach is required. Such a circuit has been devised by G. Watson[1] and is shown in Fig. 4. Reference to the original article shows that drifts should not exceed ·005%/°C. Such a characteristic, if in the right direction, is useful for compensating the thermal changes in the ramp generator proper. A further advantage of this circuit is that it can be made to operate as a true constant current generator or as a voltage generator in series with either a positive or negative resistance. The negative resistance characteristic is most desirable since it enables further compensation to be made for the losses which occur in parallel with $T_1$.

The pulse which initiates the ramp is usually derived from an astable multivibrator. There are no critical factors in this section of the circuit as long as the minimum possible period of the multivibrator is greater than the sampling rate

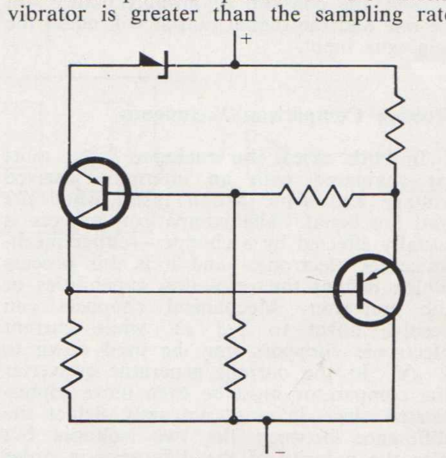


*Fig. 4. An improved constant current generator. Supply voltage may range from 10V to 25V.*

discussed above. This multivibrator also provides a pulse which resets the counter to zero in between samplings of the input voltage.■

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1. 'Two Transistors Equal One Constant Current Diode'; G. Watson, Electronics 6-7-62.

# DIRECT READ-OUT COUNTERS WITH SILICON MESA TRANSISTORS

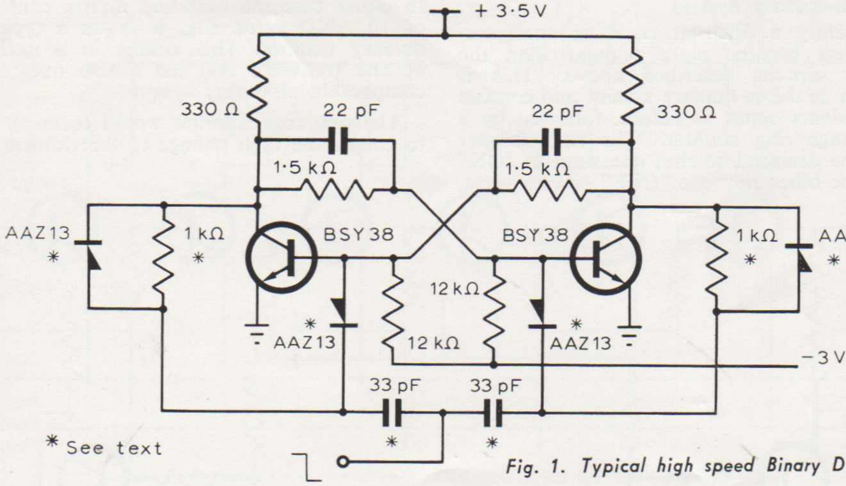


*Fig. 1. Typical high speed Binary Divider.*

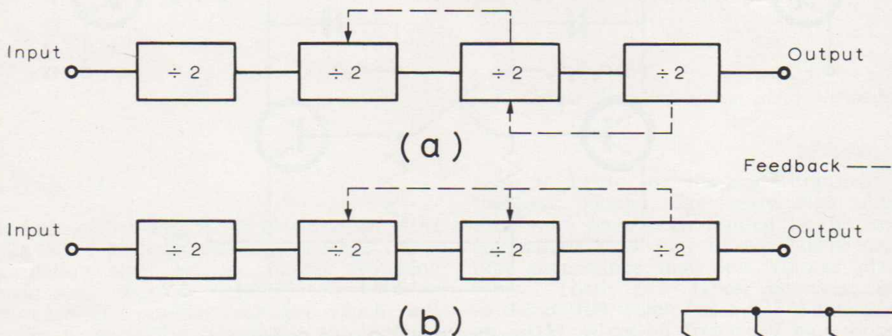


*Fig. 2. Examples of feedback paths which convert four Binary Dividers to a Decade Counter.*

One of the problems besetting the designer of counting systems has been to find a suitable method of indicating the state of the counter. A few years ago this problem was eased by the introduction of the numerical indicator tube now available in both end and side viewing styles. Unfortunately these indicators must be supplied with digital information having an amplitude of at least 60 V. The recently released series of high voltage silicon transistors enables the designer to discard the valve circuits which were previously needed and to create completely solid state counter circuits. It seems opportune at this time to review the various types of decade counting circuits available in order to establish the most economical method of using these new components.

**Binary Counters**

The most commonly used counting circuits to date have employed four stages of binary dividers. Normally cascaded, such an arrangement will provide one output pulse for every sixteen pulses fed into the input. In addition, the binary stages will establish a unique pattern of conducting and non-conducting states for each input pulse. The circuit diagram of a typical high speed binary stage is shown in Fig. 1. The components marked with an asterisk are those which convert a bistable circuit to a binary divider. The circuit as shown will operate at frequencies up to 20 Mc/s and is a fair example of the use of silicon planar transistors in high speed logic circuits. In order to use the scale of sixteen as a decimal counter, feedback is employed to advance the counter by six steps while ten pulses are applied to the input. Two of the many possible examples of feedback systems are shown in Fig. 2.

**Decoding Systems**

Irrespective of which feedback system is used, we are still faced with the problem of deriving ten signals for the numerical indicator from the eight collectors of the system. This decoding is usually done by a diode matrix, an example of which is shown in Fig. 3. From this diagram it can be seen that the diodes connected to each decimal output form an AND gate, so selecting the combination of on and off transistors which are necessary to establish that particular decimal number. The simplest method of using a numerical indicator with a diode matrix is shown in Fig. 4. The emitter side of the circuit would be connected to zero or a negative voltage depending on whether or not NPN or PNP transistors had been used for the counter. It should be remembered that this system causes a number to appear when the transistor is conducting so the input diodes must be connected to the collectors of stages which are cut off.

A more economical approach would be to couple the high voltage transistors to the transistors in the bistable circuits. This

→page 32

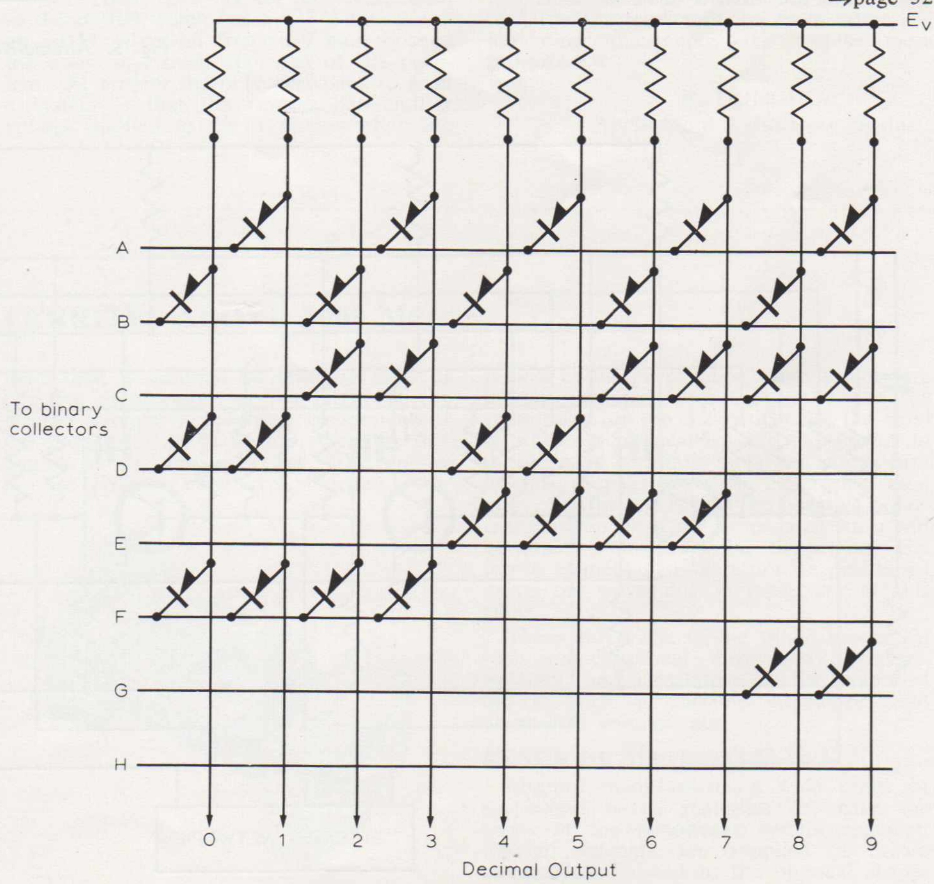


*Fig. 3. Binary to decimal decoding matrix.*

# DIRECT READ-OUT COUNTERS WITH SILICON MESA TRANSISTORS

←page 31

approach is shown in Fig. 5. The immediate saving is two transistors, but this must be balanced against the increased cost of using high voltage diodes in the decoding matrix. Fortunately there are no high speed requirements in this circuit after we have ensured that the binary transistor-collectors have not been loaded capacitively. Usually it is of little consequence if the indicator takes a few milliseconds to portray the answer, as long as the counter has performed its duties accurately.

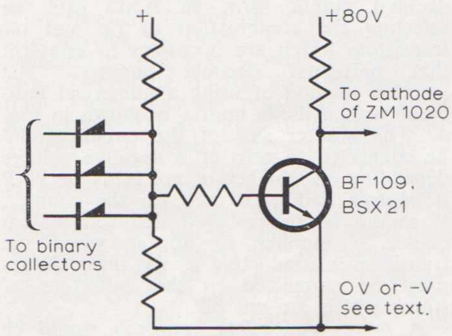


*Fig. 4. An example of coupling between matrix and numeral indicator tube.*

**The Bi-quinary System**

Recently a different counting configuration has become more popular than the binary circuits described above. This is known as the bi-quinary system and consists of a binary input as before, followed by a five stage ring counter. The ring counter may be designed so that one stage is "ON" and the other four are "OFF" or vice versa.

In either case the decoding matrix consists of 10 AND gates. Fig. 6 shows a typical quinary counter. This results in a saving of one transistor and ten diodes over the comparable all-binary system.

The optimum solution would seem to be to couple the high voltage to the transistors

→page 33

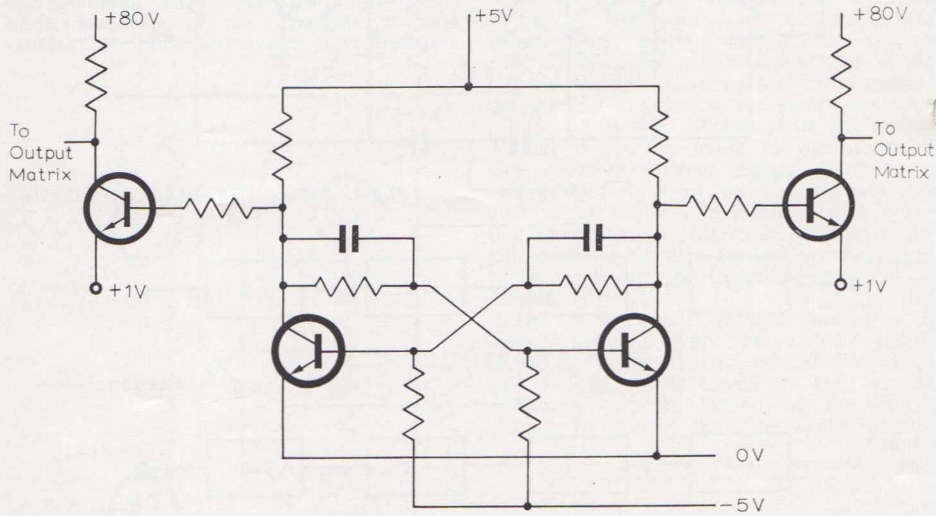


*Fig. 5. Alternative Binary Output System.*

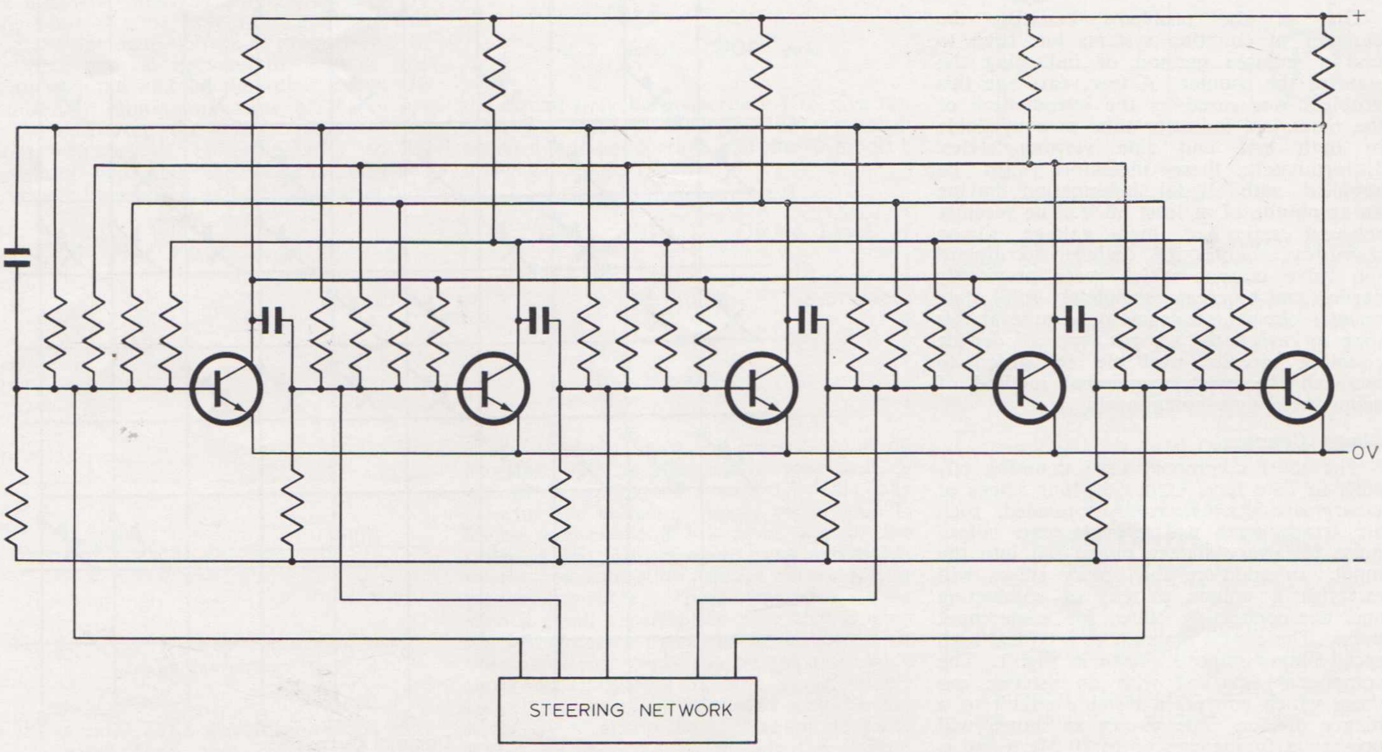


*Fig. 6. Normal configuration for Quinary Counter.*

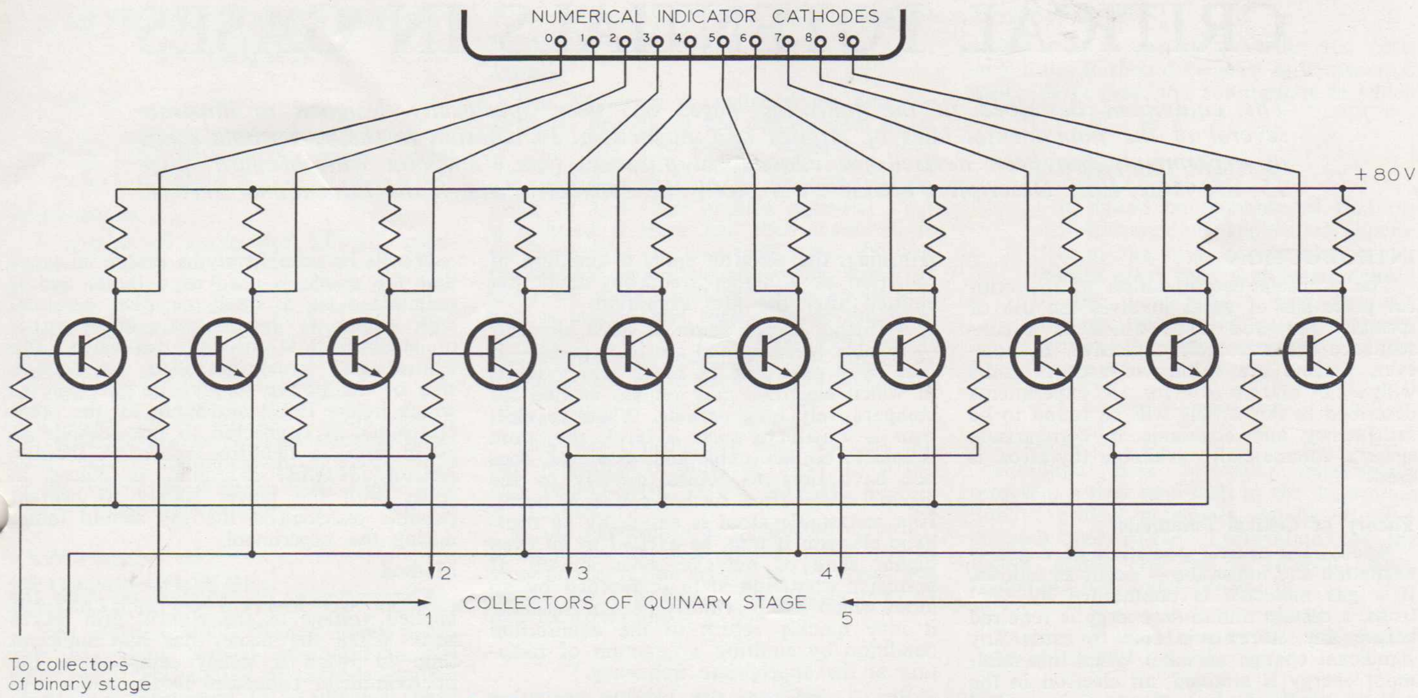


*Fig. 7. Output decoding system which eliminates diodes.*

← page 32

as indicated in Fig. 7. In this system, mesa transistors perform the high voltage transformation and also the output switching logic since there is only one combination of base and emitter voltages which will cause a particular transistor to conduct. When designing such a system care must be taken that the reverse base emitter voltage ratings of the transistors are not exceeded.

Of course the ultimate solution would seem to be to use the mesa transistor in the basic counter. Previously, high voltage transistors have been limited to the lower frequencies since they have a higher depletion capacitance than low voltage planar types. These new types, however, such as the BF109 which has a 135 V rating and an 80 Mc/s cut-off frequency have opened the door to a fresh appraisal of the problem. At present the only drawback to such a system is that the reverse base-emitter voltage applied to the transistor when the preceding stage fires, may be as great as the supply voltage, i.e., 80 V. This will certainly exceed the rating of the transistor.

The Mullard Applications Laboratory is currently investigating the problem of reverse base-emitter breakdown, with a view to extracting maximum use from numerical indicating tubes and high voltage mesa transistors.■

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# FERRITES ABOVE 2000 Mc/s

← page 28

**Ferromagnetic Resonance**

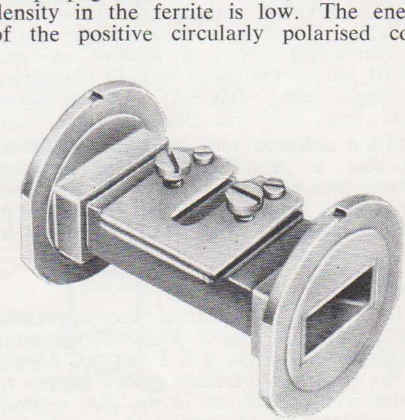
In a precessing dipole system, damping occurs and results in an absorption of energy from the microwave field. The absorption of energy reaches a peak value when the frequency of the field is equal to that of the precessing system. Ferromagnetic resonance occurs at this frequency.

**Faraday Rotation**

A linearly polarised wave which is transmitted through a ferrite-loaded wave-guide in the direction of the undirectional magnetic field experiences a rotation of the plane of polarisation as it travels along the waveguide. This effect, called the Faraday Rotation, is caused by the difference in the phase velocities of the two circularly polarised components of the transmitted wave.

**Field Displacement**

The field distribution in a waveguide can be altered by loading the waveguide with a ferrite element. If the unidirectional magnetic field is adjusted so that the ratio of effective permeability of the ferrite to the positive circularly polarised component of the propagated wave is zero, then the flux density in the ferrite is low. The energy of the positive circularly polarised component of the propagating wave is therefore largely excluded from the ferrite. At the same value of the magnetic field, the ratio of effective permeability of the ferrite to the negative circularly polarised component of the propagated wave is greater than unity, and the energy of the negative polarised component of the propagated wave will be largely contained within the ferrite. The ferrite element can therefore be positioned inside the waveguide to make use of this effect.



*X-band wave guide filter*

These three effects are made use of in such non-reciprocal devices as gyrators, isolators, and circulators, or in reciprocal devices such as variable attenuators and phase-shift components.

**RANGE OF MATERIALS**

Mullard manufactures a wide range of specialised ferrite materials to cover the needs of the microwave system engineer. Ferrite elements are supplied as tested components finished to the physical shapes required in the device.■

CRITICAL POTENTIALS IN GASES

# CRITICAL POTENTIALS IN GASES

*The equipment described in the following pages has been specifically designed to illustrate several of the fundamental laws of physics and magnetism. In addition to these, a whole range of experiments has been devised for educational purposes (see Viewpoint with Mullard, page 15, this issue. Ed.). Descriptive booklets will shortly be available through the Educational Service.*

## INTRODUCTION

The accurate determination of the critical potentials of gases involves the use of specially prepared tubes which as a consequence are extremely expensive. If, however, readers are willing to accept results with a fair margin of error, the experiments described in this article will be found to be satisfactory and economic in comparison, since a commercially available thyratron is used.

### Theory of Critical Potentials

Briefly, the critical potentials of a gas — excitation and ionisation — occur as follows. If a gas molecule is bombarded by electrons, a certain minimum energy is required before any electron is able to cause any significant change of state. When this minimum energy is attained, an electron in the valency shell of the gas atom is excited by the collision and jumps into an orbit of higher energy. Immediately afterwards a return to the equilibrium or ground state may occur, the excess energy being given off as a quantum of electromagnetic radiation. The frequency of this radiation is proportional to the energy given up in this transition and can be calculated from Planck's equation if desired. This minimum energy is the excitation energy which varies from gas to gas but which in the case of Xenon is of the order of 8·4 volts.

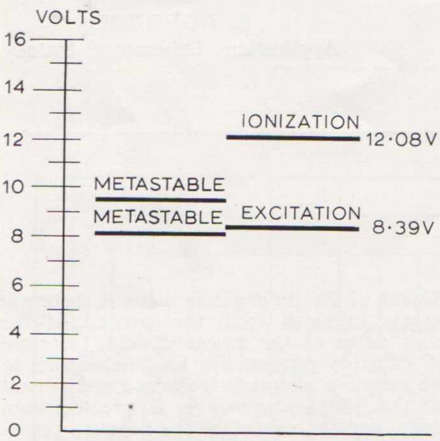



*Fig. 1*

If now the speed — and hence the energy — of the bombarding electron is increased further, a second critical state is reached where a valency electron is given sufficient energy to leave the atom completely, causing ionisation. This second critical potential is the ionisation potential which also varies from gas to gas but for Xenon is of the order of 12 volts.

A second excitation can occur at a potential higher than that of the first critical potential. Here an orbital electron of higher energy is lifted from the ground state to a level of higher energy. Upon returning, the electron emits a quantum of radiation at a higher frequency than that emitted after the first excitation.

A further complication in some elements — notably mercury and the inert gases — is that there exist one or more energy levels in which electrons may remain excited for comparatively long periods. When an electron is raised to such a level, the atom is said to be metastable and, as a rule, does not have time to return directly to the ground state before further collisions occur. If a metastable atom is struck by an energetic electron it may be excited to an even higher state or may be even ionised. If a higher excitation state is reached by an atom which has previously been metastable, it may quickly return to the equilibrium condition by emitting a quantum of radiation at the appropriate frequency.

Fig. 1 indicates the various excitation and ionisation potentials for Xenon, which is the gas chosen for the experiments described in this article.

## EXPERIMENT 1: DETERMINATION OF FIRST EXCITATION POTENTIAL OF XENON

In this experiment a modified Hertz method is used. It is assumed that cathode-emitted electrons, with just sufficient energy to cause excitation, come to rest after an excitation collision and are then unable to overcome a retarding potential existing between two electrodes placed in the electron stream. The Mullard EN91 thyratron, which is Xenon filled, can be used to demonstrate this effect.

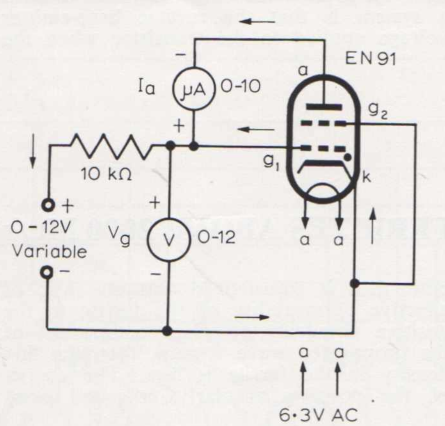



*Fig. 2*

### Apparatus

The circuit diagram of the apparatus used is shown in Fig. 2. Major components include the thyratron, a 6·3 volt AC supply for the heater, a variable power supply to accelerate the electrons in the gas, a limiting resistor, a voltmeter and a sensitive microammeter.

It will be seen from the circuit diagram that the anode is used as a target and is maintained at a small negative potential with respect to the control grid by virtue of electron flow through the valve. The control grid is connected to the positive side of the power supply and the grid $g_2$, which in the EN91 surrounds all the other electrodes, is connected to the cathode so as to form a field-free space. A limiting resistor, 10,000Ω in value, is placed in series with the power supply to prevent possible overload if the gas should ionise during the experiment.

### Method

The apparatus is switched on with the applied voltage to the control grid set to zero. When the heater has had sufficient time to reach a steady temperature the microammeter connected to the anode will indicate a steady current arising from thermionic emission at the cathode.

The control grid voltage ($V_g$) is then increased in small steps up to about 11 volts and the corresponding values of anode current ($I_a$) noted.

A graph of control grid potential versus anode current is then plotted and will be of the form shown in Fig. 3.

The shape of this curve may be explained as follows. Even at grid potentials lower than expected for ionisation, a few electrons passing through the centre of the control grid cylinder have sufficient velocity to cause excitation and therefore lose their energy. They are thus unable to overcome the negative field surrounding the target. As the applied voltage is increased, more and more electrons lose their energy after excitation and the target current falls accordingly. A minimum value of target current is achieved when most of the electrons passing through the control grid are able to cause excitation. As the potential is increased beyond this critical value, electrons, after causing excitation, are able to

→page 35

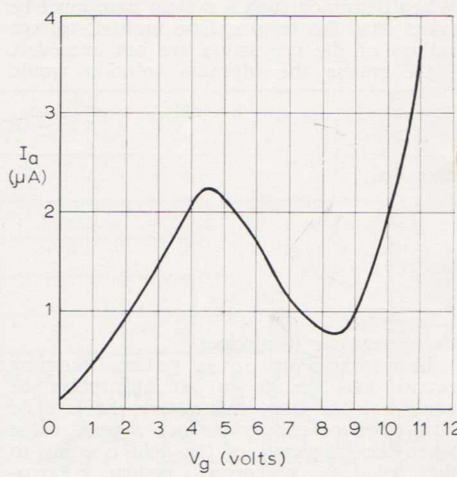



*Fig. 3*

## CRITICAL POTENTIALS IN GASES

←page 34

accelerate sufficiently to overcome again the retarding potential and the target current commences to rise.

**Observations**

A number of curves for different specimens of EN91 thyratron has been taken giving a value of first excitation potential at about 8·2 volts. It is presumed that contact potentials in the thyratron may affect the results obtained and this could explain the slight differences between one specimen of tube and another.

It might be of interest to note that if the readings are taken with extreme accuracy, there is an indication of sharper dips in the region of 8 to 9 volts. These may indicate metastable states.

**EXPERIMENT 2: DETERMINATION OF IONIZATION POTENTIAL OF XENON**

The Mullard EN91 Xenon filled thyratron has again proved successful in this experiment. The energy of cathode emitted electrons is increased to a sufficient level to allow ionisation and a small current resulting from positive gas molecules appears at the target electrode which is maintained at a large negative potential of about 50 volts.

**Apparatus**

The circuit diagram of the apparatus used is shown in Fig. 4. Major components

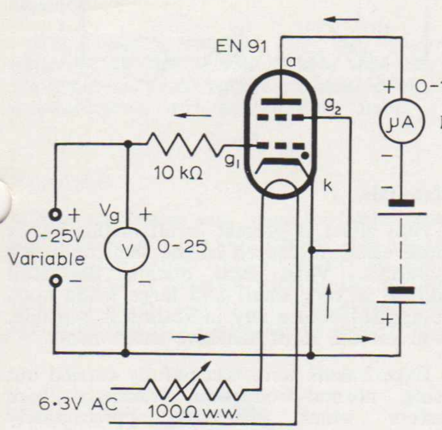



*Fig. 4*

include the thyratron, power supplies (one variable), a voltmeter and a sensitive microammeter.

It will be realised that electrons emitted from the cathode are accelerated towards the postive control grid. The anode of the thyratron is again used as a target and is maintained at a large negative potential with respect to the cathode. The grid $g_2$ is connected to the cathode and a 10,000Ω resistor is connected in series with the control grid supply voltage to limit the cathode-grid current after ionisation. Without this resistor, the current may exceed the maximum permissible value of 10mA resulting in destruction of the thyratron.

It has been found desirable to operate the heater of the thyratron at a greatly reduced voltage to reduce the space charge to a minimum and thus to increase the sensitivity of the experiment.

**Method**

The apparatus is switched on and the potential applied to the control grid is set to 15 volts. The heater voltage is then *slowly* adjusted for maximum reading of anode current ($I_a$) which should be of the order of 1$\mu$A. The applied potential ($V_g$) is reduced to zero and then increased in small steps up to about 20 volts, the corresponding anode current being noted. A graph of applied voltage versus anode current is plotted and should be of the form shown in Fig. 5. The exact potential at which ionisation takes place is found by extending the curve back to cut the $V_g$ axis.

**Observations**

This experiment has been carried out with a number of different EN91 thyratrons and results have varied between 12 and 12·5 volts.

If the experiment is conducted carefully, a slight discontinuity in the curve will be seen in the region of $V_g = 14$ volts. The reason for this is obscure but is possibly due to excitation.

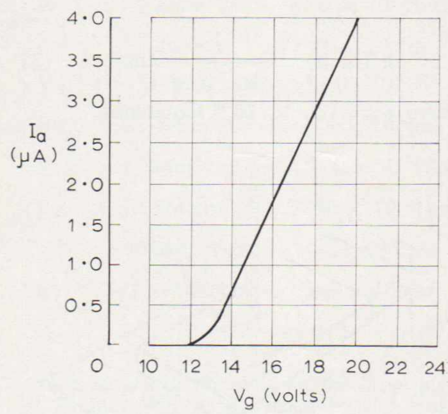



*Fig. 5*

**NOTES ON APPARATUS USED**

**Meters**

In both experiments the applied potential was measured on a 20,000Ω per volt instrument. For current measurements a 10$\mu$A meter is called for but if unobtainable, a transistor DC current amplifier can be used.

**Power Supplies**

In both experiments, the variable potential was derived from a power supply having a variable output up to 150 volts. The current drawn in both experiments is negligible since it is limited by the 10,000Ω series resistor. The same power supply unit can also provide the 6·3 vclt heater supplies.

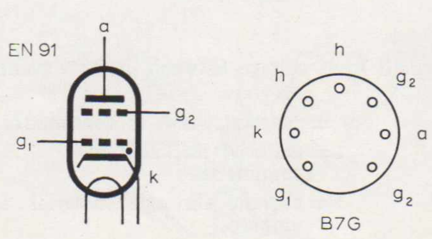



*Fig. 6*

**Thyratron EN91**

This tube is readily available and costs very little. It should be used in conjunction with a B7G base, the connections to which are shown in Fig. 6.

**IMPORTANT NOTICE**

It should be remembered that for any conventional application, operation of the thyratron at any heater voltage other than that recommended by the manufacturer may seriously damage the cathode and greatly shorten the life expectancy of the device. ■

**References**

Although these particular experiments represent a new approach to the determination of critical potentials, details of other methods and further explanations of the theory might also be of interest. Some references are given below.

*Physics of the Atom* by Wehr-Richards. Addison-Wesley Co. Inc.

*Electronics* by Parker. Arnold.

## MULLARD EDUCATIONAL SERVICE

←page 16

mium ratio difference and fraction, resonance activation, self-absorption techniques, experiments on foils, data sheets on foils and extensive references. In addition, there is a special educational foil kit at lower prices for training purposes, which also contains an Experiment manual. The educational foils are designed to be used in subcritical assemblies, neutron howitzers, critical assemblies, training reactors and particle accelerators.

**Geiger Counter Tubes**

We feel little reference is required to the wide and ever-increasing range of Mullard Geiger Counter Tubes that are available and have been in extensive use for many years in Australia's training establishments.

**The Future and the Scope**

Readers will see that already the Mullard Educational Service covers a fairly wide field and if it were practicable, the temptation is great to extend this on a grand scale. We therefore only cover our particular field, believing that language laboratories, high-speed readers and so on, are the province of the system supplier and the tutor, each in his own right.

We are mindful of the opportunity and the extreme satisfaction that can, and will accrue, from those who see the need of additional sophisticated and elegant youth training aids and who will create, produce and supply these items.■

# HALL EFFECT MEASUREMENTS

**Introduction**

The Hall Effect is named after E. H. Hall who discovered it in 1879 in the Rowland Laboratory of the Johns Hopkins University. He was experimenting with a strip of gold leaf and found that a magnetic field perpendicular to the strip caused a deviation of the charge carriers in the material. The effect is of great importance in the semiconductor field since it allows a measurement to be made of the charge carrier density and also allows determination of the polarity of the carriers i.e. electrons or holes.

**Theory**

If a suitable specimen of metal or n-type (excess negative charge carriers) semiconductor material has its end faces connected to a battery, electrons will flow from left to right (Fig. 1). A magnetic field (B) applied perpendicularly to the paper exerts a force on the electrons, causing a few of them initially to drift towards side Q of the material. This initial drift of electrons sets up a transverse electric field between faces R and Q opposing the drift of further electrons and the transverse current rapidly falls to zero.

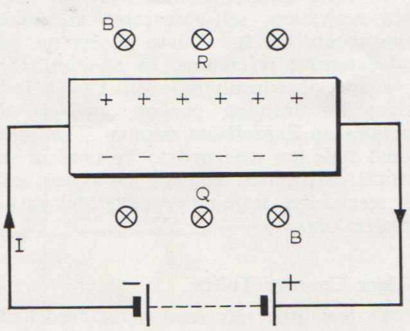



*Fig. 1*

The potential difference between faces Q and R can be measured with a high impedance voltmeter and is called the "Hall Voltage" ($V_H$). The polarity of $V_H$ is, in most cases, governed by the predominant carriers in the material i.e. electrons in n-type and holes in p-type. If holes are the predominant carriers they drift from right to left in Fig. 1 and on applying the magnetic field a few are caused initially to drift towards the side Q. The polarity of the voltage between Q and R is therefore different from the case of an n-type sample.

Measurement of the Hall voltage, field strength, current and material thickness enables the calculation of a quantity known as the "Hall Coefficient" ($R_H$).

$$R_H = \frac{V_H \times t \times 10^8}{I \times B}\ \text{cm}^3/\text{Coulomb} \qquad (1)$$

where $V_H$ = Hall voltage in volts

$t$ = material thickness in centimetres

$I$ = current through material in amperes

$B$ = flux density in gauss

The electron concentration $n$ (or in the case of a p-type semiconductor, the hole concentration $p$) is given by the formula:

$$n \text{ (or } p) = \frac{1}{R_H \times e}\ \text{cm}^{-3} \qquad (2)$$

where $e = 1{\cdot}6 \times 10^{-19}$ Coulombs.

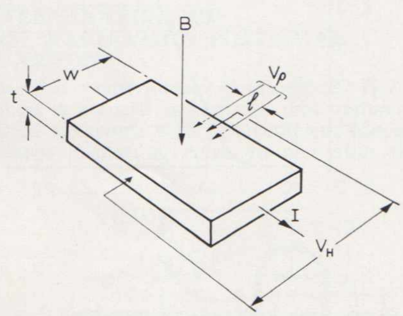



*Fig. 2*

One of the most commonly measured electrical properties of a semiconductor is its resistivity ($\rho$). This can easily be obtained by measuring the voltage $V_\rho$ between two points, using a high impedance voltmeter, as shown in Fig. 2. The value of $\rho$ can be calculated using the equation:

$$\rho = \frac{V_\rho \times w \times t}{I \times l}\ \Omega\text{cm} \qquad (3)$$

where $V_\rho$ = voltage between the two points in volts

$w$ = material width in centimetres

$t$ = material thickness in centimetres

$I$ = current through material in amperes

$l$ = distance between the two points in centimetres.

The results obtained in equations (1) and (2) can be used to calculate the carrier mobility ($\mu$). This is the drift velocity of the charge carrier acquired per unit electric field and is given by:

$$\mu = \frac{R_H}{\rho}\ \text{cm}^2/\text{volt sec.} \qquad (4)$$

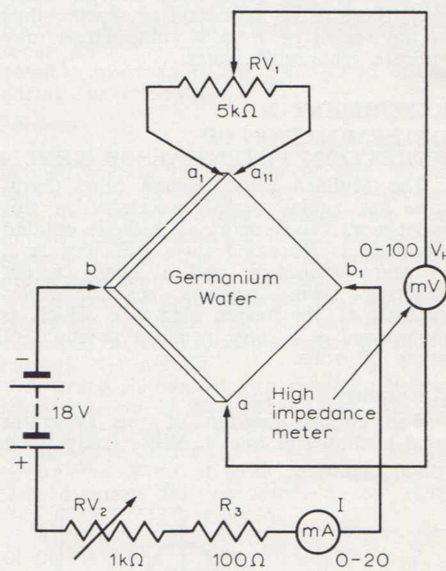



*Fig. 3*

**Materials**

Hall effect is present in all metals but is more easily measured in the semiconductor materials. With most metals, the Hall voltage is very small and large fields must be applied before any indication is possible, even on the most sensitive instruments.

Experiments were successfully carried out using normal-production transistor base wafers which measure approximately 5 × 5 × 0·4 mm. In practice, it was found convenient to make contacts to the corners of the wafer as shown in Figs. 3 and 4. Fig. 4 shows the completed prototype viewed from both sides. A suitable magnetic field may be provided by the use of permanent magnets normally available in the Science classroom.

**Practical Measurements**

The circuit of Fig. 3 is set up and RV2 is varied to set the current at a convenient value. This value will depend upon the size of the material, the size of the field and the sensitivity of the instrument measuring $V_H$. R3 is a fixed limiting resistor necessary to prevent excessive current and subsequent damage to the sample. For the wafer quoted, the current should not exceed 12 mA.

→page 37

# HALL EFFECT MEASUREMENTS

← page 36

If two contacts are used for the voltage measurement and they are not exactly opposite each other, a residual voltage will be indicated on the voltmeter, even before a field is applied. To eliminate this voltage, three contacts (a, a′ a″) are used with a potentiometer RV1 and the null point is obtained before applying the field. The field is then applied and the voltage measured.

The field direction and current direction can be reversed in turn, four values of $V_H$ obtained and a mean value calculated. The various measurements taken can then be substituted in equation (1) to obtain $R_H$ and, in turn, $R_H$ used to calculated n.

The resistivity of the sample can also be measured (with the field removed) by passing a known current through the sample and measuring the potential drop ($V_\rho$) across the contacts a′ and a″. RV1 must be disconnected when this measurement is taken. Use formula (3) to calculate $\rho$ and substitute this value for x in equation (4) to obtain the carrier mobility ($\mu$).

When making the above calculations some difficulty can arise unless care is taken with the units mentioned earlier. Many authors quote widely differing quantities and systems and in the section marked "Theory" all calculations are made in terms of volts, amperes, centimetres and gauss.

### Results

The results obtained in the original experiments are given below and serve merely as a guide to the magnitude of the measurements and calculations made.

### Apparatus

The specimens are usually fairly small and fragile and for this reason it is recommended that some form of holder be made. It is difficult to connect wires to the surfaces of the material and thus it is suggested that spring-loaded point contacts be employed.

In the prototype holder, 8 mm watch-strap bars were used and held in position by solder tags which also served as terminals. An exploded view of the holder is shown in Fig. 5 and the only component requiring any degree of accuracy of manufacture is the Perspex piece containing the bars and wafer. The centre hole can be drilled and then filed out with a $\frac{1}{8}$ inch square file until it is 5 mm square. The holes for the watch-strap bars are drilled into the sides of the Perspex using a size 52 drill. The piece of packing material is necessary to raise the germanium wafer so that contact is made with the bars.

| | C.G.S. UNITS | M.K.S. UNITS |
|---|---|---|
| B | 1000 Gauss | $0{\cdot}1 Wb/m^2$ |
| I | 3mA | $3 \times 10^{-3} A$ |
| t | 0·038mm | $3{\cdot}8 \times 10^{-5} m$ |
| W | 6mm | $6 \times 10^{-3} m$ |
| $l$ | 2mm | $2 \times 10^{-3} m$ |
| $V_H$ | 9·3mV | $9{\cdot}3 \times 10^{-3} V$ |
| $R_H$ | $11{,}780 cm^3/C$ | $1{\cdot}178 \times 10^{-2} m^3/C$ |
| $V_\rho$ | 94mV | $9{\cdot}4 \times 10^{-2} V$ |
| $\rho$ | 3.56Ωcm | $3{\cdot}56 \times 10^{-2} \Omega m$ |
| $\mu$ | $3{,}300 cm^2/Vsec$ | $0{\cdot}33 m^2/Vsec$ |
| n | $5{\cdot}3 \times 10^{14}/cm^3$ | $5{\cdot}3 \times 10^{8}/m^3$ |

The length of the slots on the large piece of Perspex will be governed by the length of the bars obtained. These bars are stocked by most jewellers and cost a few shillings each. The balance potentiometer RV1 can be made an integral part of the holder by using one of the small carbon trimming potentiometers now available on the market. These are typically less than 1 inch square and can be bolted to the larger Perspex piece.

NOTE: Mullard-Australia Pty. Ltd. cannot make this equipment available either in completed form or as a kit of parts. This leaflet has been produced simply as a guide as to what can be achieved by the school itself.

This article is intended only as a guide; other experiments which though possible, have not been attempted are variation of resistivity—and hence carrier mobility—with temperature; the effect on Hall voltage of various temperatures, and the measurement of magnetic fields once the initial results are obtained for a given sample. ■

### Further Reading

The Hall Effect and Related Phenomena by E. H. Putley (Butterworth).

Hall Effect in Semiconductor Compounds by M. J. O. Strutt (Electronic and Radio Engineer 1959, pp. 2-10).

A Simple Laboratory Method for the Measurement of Some Properties of a Semiconductor, by B. Stuttard (International Journal of Electrical Engineering Education, Vol. 1, No. 1, June 1963).

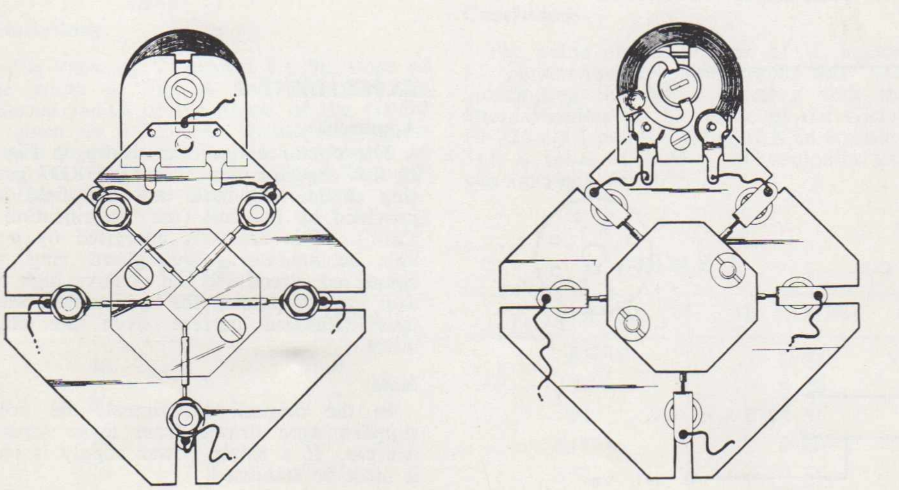


*Fig. 4*

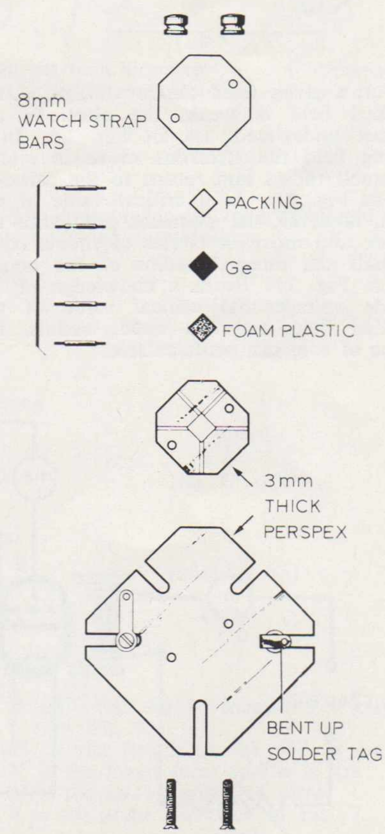



*Fig. 5*

# DETERMINATION OF $\frac{e}{m}$ USING THE MAGNETRON EFFECT

This experiment provides a simple and inexpensive method of determining the charge-to-mass ratio of the electron.

By the application of a magnetic field parallel to the straight filament of a diode having a co-axial cylindrical anode, the electrons are deflected into curved paths which approximate to circles.

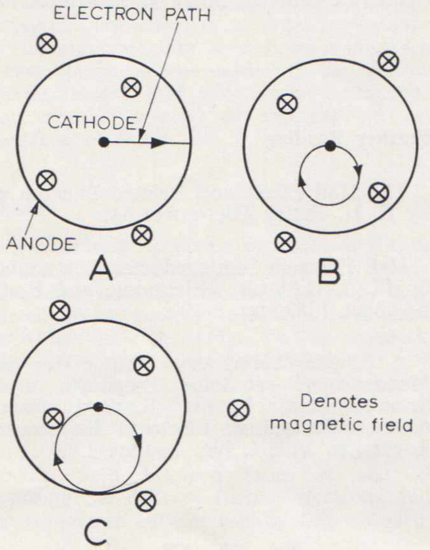



*Fig. 1*

For a given fixed anode voltage, if the applied field is weak, the electrons are almost undeviated. (A in Fig. 1). In a strong field the electrons move in circles of small radius and return to the cathode. (B in Fig. 1). At a critical value of the field, however, the electrons just graze the anode and move in circles of radius equal to half the internal radius of the anode. (C in Fig. 1). From a knowledge of the anode voltage, the critical value of the applied field and the anode radius, the value of e/m can be calculated.

**Theory**

Most of the potential difference between the anode and the cathode acts across the first few millimetres of electron path and the electrons gain their velocity in this distance. Also, the task of moving electrons from the cathode to the anode requires all the kinetic energy they possess and the centrifugal force on the electrons just balances the force due to the field. We can therefore write:

$$\tfrac{1}{2}mv^2 = eV_a \qquad (1)$$

$$\frac{mv^2}{r} = Bev \qquad (2)$$

Combining equations (1) and (2) we have

$$\frac{e}{m} = \frac{2V_a \times 10^8}{B^2r^2} \text{ coulombs/g} \qquad (3)$$

where

B is the applied flux density in gauss

$V_a$ is the anode voltage

v is the electron velocity in cms./sec

r is the radius of the circle of motion of electrons measured in centimetres.

e and m are the charge and mass respectively of the electron

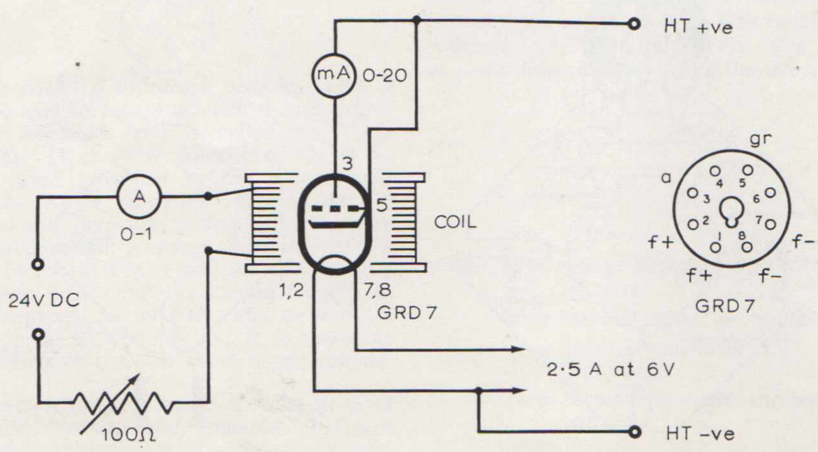



*Fig. 2*

If R, the anode radius, is introduced instead of r and the field is provided by a coil having a constant K, where K is defined by

$$\left(B^2 = \frac{Ic^2}{K}\right)$$

then equation (3) becomes

$$\frac{e}{m} = \frac{8V_aK \times 10^8}{I_c^2R^2} \qquad (4)$$

where $I_c$ is the current flowing through the coil and the constant, K, can be calculated from the dimensions of the coil.

It can be seen from equation (4) that in addition to the method outlined in the introduction, e/m can be determined by keeping $I_c$ constant and varying the anode voltage $V_a$ until the required conditions are fulfilled. This does not, of course, serve as a check on the accuracy of the first method as the same types of error are present in both cases.

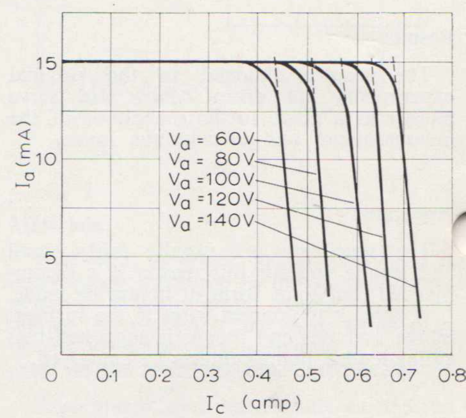



*Fig. 3*

**EXPERIMENT 1**

**Apparatus**

The circuit diagram is shown in Fig. 2. In this experiment a Ferranti GRD7 guard ring diode was used and the field was provided by the coil (see "Specification of Coil"). This coil was energised by a 24 volt accumulator. The guard ring was connected directly to the positive high tension line, by-passing the milliameter, which thus measured current from the anode alone.

**Note**

In the original experiment, the power supplies were drawn from three separate sources. If a single power supply is used, it must be stabilised.

→page 39

## DETERMINATION OF $\frac{e}{m}$ USING THE MAGNETRON EFFECT

← page 38

### Method

The anode voltage, $V_a$, was kept constant and the coil current $I_c$ was increased by means of a rheostat until there was a sudden fall in the anode current, $I_a$, indicating that the field was sufficiently large to prevent electrons from reaching the anode. This process was repeated for several values of $V_a$.

### Observations

A graph was drawn of $I_a$ versus $I_c$ for each value of $V_a$ (Fig. 3) and the value of $I_c$ at which $I_a$ falls off in each case was determined as the point of intersection of tangents drawn to the curve above and below the "knee". The values of $I_c$ obtained in this way were tabulated against the corresponding values of $V_a$ (Table 1) and a graph was then drawn of $V_a$ against $I_c^2$ (Fig. 4).

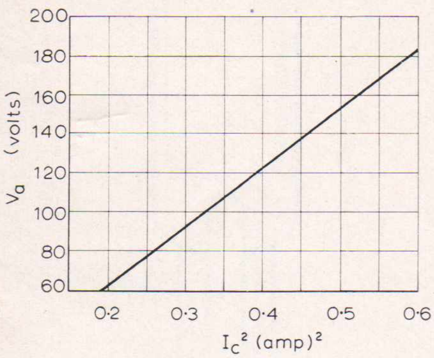



*Fig. 4*

### Conclusions

The value of $V_a$ against $I_c^2$, the slope of the graph in Fig. 4, was 300 and the internal radius of the anode of the GRD7 is given as 0·325 cm. Substituting these results together with the coil constant K in equation (4), a value of $1.73 \times 10^7$ coulombs/gm. was obtained for e/m.

**TABLE 1**

| $V_a$ (VOLTS) | $I_c$ (AMPS) |
|---|---|
| 60 | 0·44 |
| 80 | 0·51 |
| 100 | 0·57 |
| 120 | 0·63 |
| 140 | 0·68 |

## EXPERIMENT 2

### Apparatus

The circuit used for this experiment was exactly the same as that used for Experiment 1.

### Method

The current through the coil, $I_c$, was kept constant and the anode voltage $V_a$ was decreased from a large value until there was a fall in the anode current $I_a$. This process was repeated for several fixed values of $I_c$.

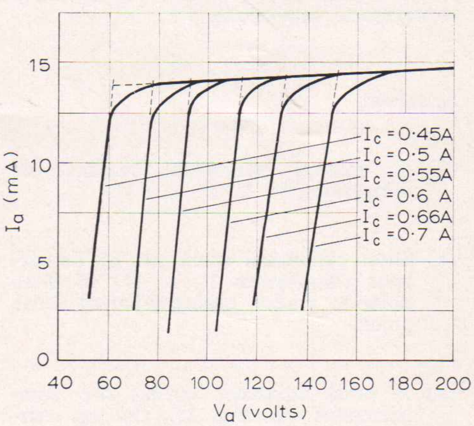



*Fig. 5*

### Observations

A graph was drawn of $V_a$ against $I_a$ (Fig. 5) and the critical values of $V_a$ were determined by the intersection of tangents drawn to the curve on either side of the "knee". These values of $V_a$ were tabulated against the corresponding values of $I_c$ (Table 2) and a graph was drawn of $V_a$ against $I_c^2$ (Fig. 6).

### Conclusions

The value of the relation of $V_a$ against $I_c^2$ obtained from this graph was 304. Substituting this value together with the internal radius of the anode of the GRD7 (0·325 cm.) and the value of K in equation (4), a value of $1 \cdot 76 \times 10^7$ coulombs/gm. was obtained for e/m.

**TABLE 2**

| $V_a$ (VOLTS) | $I_c$ (AMPS) |
|---|---|
| 62 | 0·45 |
| 78 | 0·50 |
| 94 | 0·55 |
| 114 | 0·60 |
| 132 | 0·66 |
| 152 | 0·70 |

### Specification of Coil

The dimensions of the coil used in the prototype experiments 1 and 2 are shown in Fig. 7. It comprised 2,100 turns of 23 s.w.g. enamel wire wound in 22 layers giving a resistance of approximately 25 ohms.

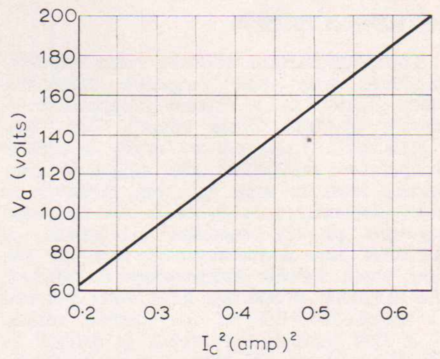



*Fig. 6*

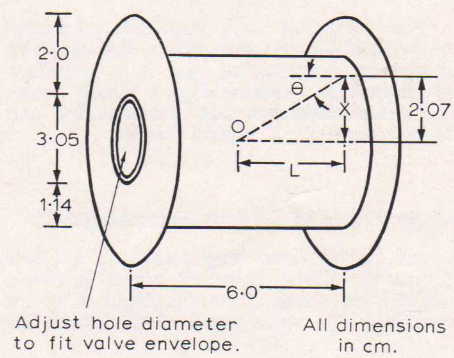



*Fig. 7*

The coil constant, K, was calculated from the following relationship:

$$B = \frac{4\pi N I_c \cos\theta}{10 \times 2L}$$

hence

$$B^2 = \frac{4\pi^2 N^2 I_c^2}{100\,(L^2 + X^2)}$$

and

$$K = \frac{100(L^2 + X^2)}{4\,\pi^2 N^2}$$

where

B is the flux density measured in gauss at $\theta$ in Fig. 7
2L is the length of the coil in cm
X is the mean turns radius in cm
N is the total number of turns
$\theta$ is the angle indicated in Fig. 7

and for the coil specified above, the value of K was taken as $7 \cdot 63 \times 10^{-6}$. ■

# DETERMINATION OF THE CURIE POINT OF FERROXCUBE

**INTRODUCTION**

The phenomenon whereby ferro-magnetic materials lose their magnetic properties when heated to a certain temperature is known as "the Curie effect". The term "Curie point" appears to imply a precise temperature at which the magnetic properties cease to exist but this, however, is not what occurs in practice. As the temperature of the specimen is raised, it becomes less ferro-magnetic and by the time some specific temperature is reached, the magnetic properties have been reduced to about 5%-10% of the normal values. It is this temperature which is defined as the Curie point. Of course, some magnetic properties do exist at temperatures exceeding the Curie temperature, but for practical applications these properties are negligible.

The values of such magnetic properties as permeability do not reduce linearly with increasing temperature and it is therefore of interest to examine Fig. 1 which shows the relationship between permeability and temperature of a typical ferrite.

**Curie Points of Various Ferro-Magnetics**

All known ferro-magnetics have a Curie point: the Curie point of cobalt is about 1200°C but those of certain alloys of nickel and iron are much lower. This factor can be employed in the production of special alloys for use as magnetic shunts, negative coefficient compensators on permanent magnets and also for applications in instruments, where it is used to counteract the errors inherent in the behaviour of springs with changing temperature. The Curie points of the samples of Ferroxcube selected for this demonstration lie in the region of 100°C to 250°C and the values are relatively easy to determine.

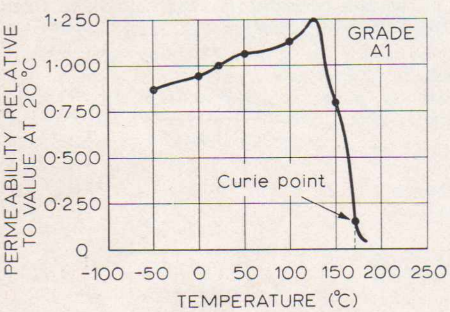



*Fig. 1*

**Ferroxcube**

Ferroxcube is a ferro-magnetic material developed principally for use as a magnetic core material in inductors, transformers and similar wound components for applications at and above audio frequencies. In such cases, the use of conventional permeable core materials results in high losses mainly due to eddy currents. However, ferrites such as Ferroxcube have a very high resistivity, so that eddy currents are generally of negligible importance.

**Apparatus**

(1) A Ticonal permanent magnet, of dimensions $\frac{3}{8}'' \times \frac{3}{8}'' \times 1\frac{1}{2}''$.

(2) Short cylindrical specimens of Ferroxcube (dimensions $1'' \times \frac{1}{4}''$) of three different grades, having different Curie points.

(3) A brass container having two compartments (see Fig. 2). The top compartment has solid walls and accommodates the permanent magnet. The lower compartment is intended to accommodate the Ferroxcube, and is slotted to ensure good circulation of oil and to permit the specimen to be observed.

(4) Various pieces of apparatus such as a thermometer, retort stand, etc.

**Method**

The permanent magnet is inserted into the top compartment of the container and the Ferroxcube inserted into the lower compartment, the retaining pin being inserted.

The container is then suspended vertically in an oil-bath, taking care that the whole of the Ferroxcube is immersed. It is, in fact, preferable that the whole of the container be immersed. The temperatures specified for the experiment will not affect the permanent magnet.

After placing a thermometer in the liquid, close to the specimen, the oil bath is heated while the oil is continuously stirred to ensure a uniform temperature.

When the Curie point is reached, the Ferroxcube will fall away from the permanent magnet. The temperature at which this occurs should be noted. After heating the oil through a further 10°C or so, it is then allowed to cool until the Curie point is again reached. At this temperature, the Ferroxcube will again be attracted to the magnet.

The mean of the two temperatures recorded may be taken as the Curie point of the specimen under test. The experiment may be repeated with each of the three specimens of Ferroxcube.

**Observations**

The observed values of the Curie point can be checked against the values given in Fig. 3.■

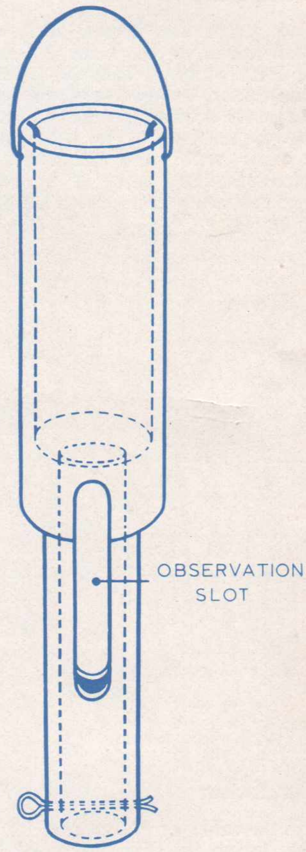



*Fig. 2*

| Grade of Ferroxcube | Approx. Curie Point |
|---|---|
| A4 | 150°C |
| B1 | 200°C |
| B2 | 250°C |

*Fig. 3 Table showing Curie points of various grades of Ferroxcube*



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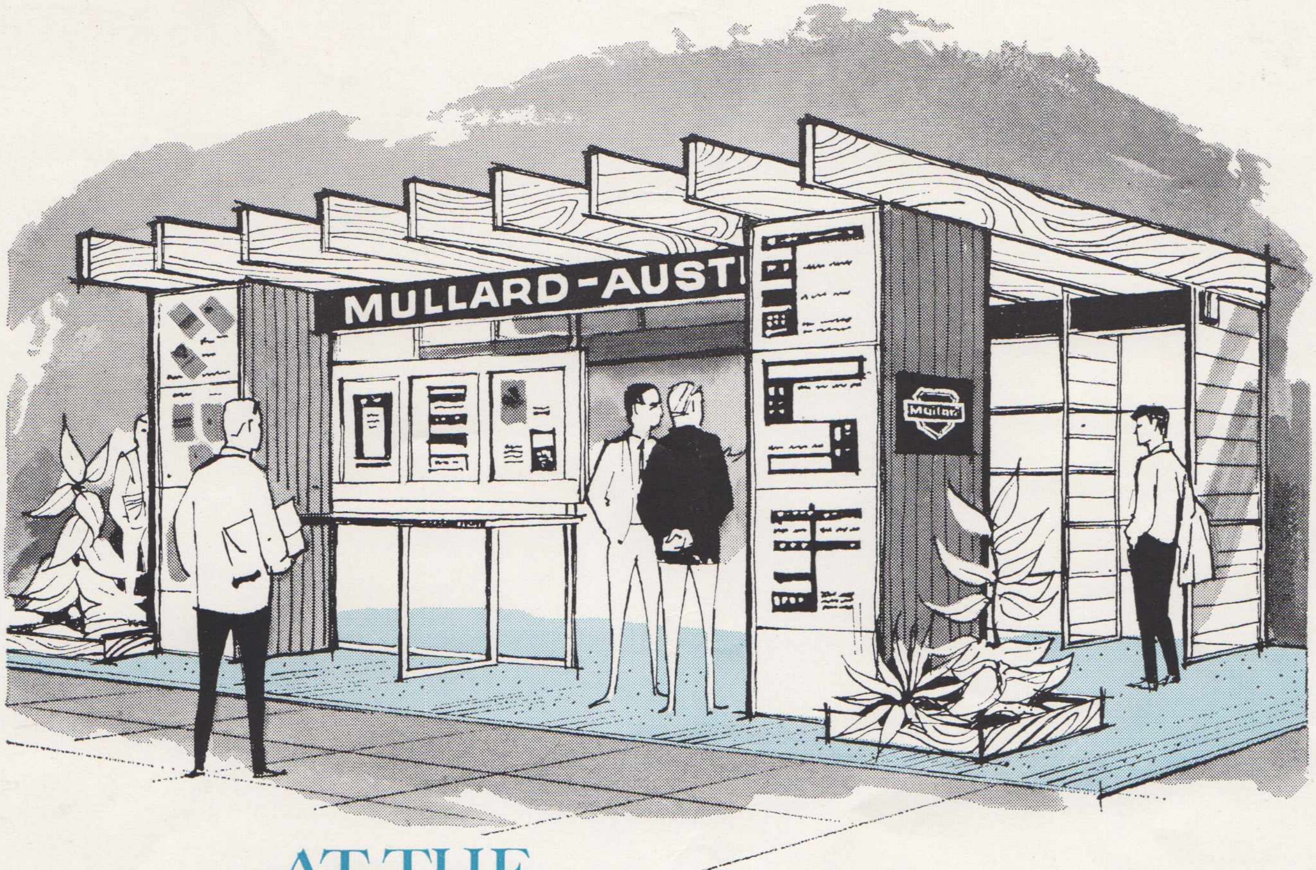


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STAND Nos. 27 and 28

**Mullard-Australia Pty. Ltd.**

35-43 CLARENCE STREET, SYDNEY, N.S.W., 29 2006.
123 VICTORIA PARADE, COLLINGWOOD, N.5, VIC., 41 6644
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M153