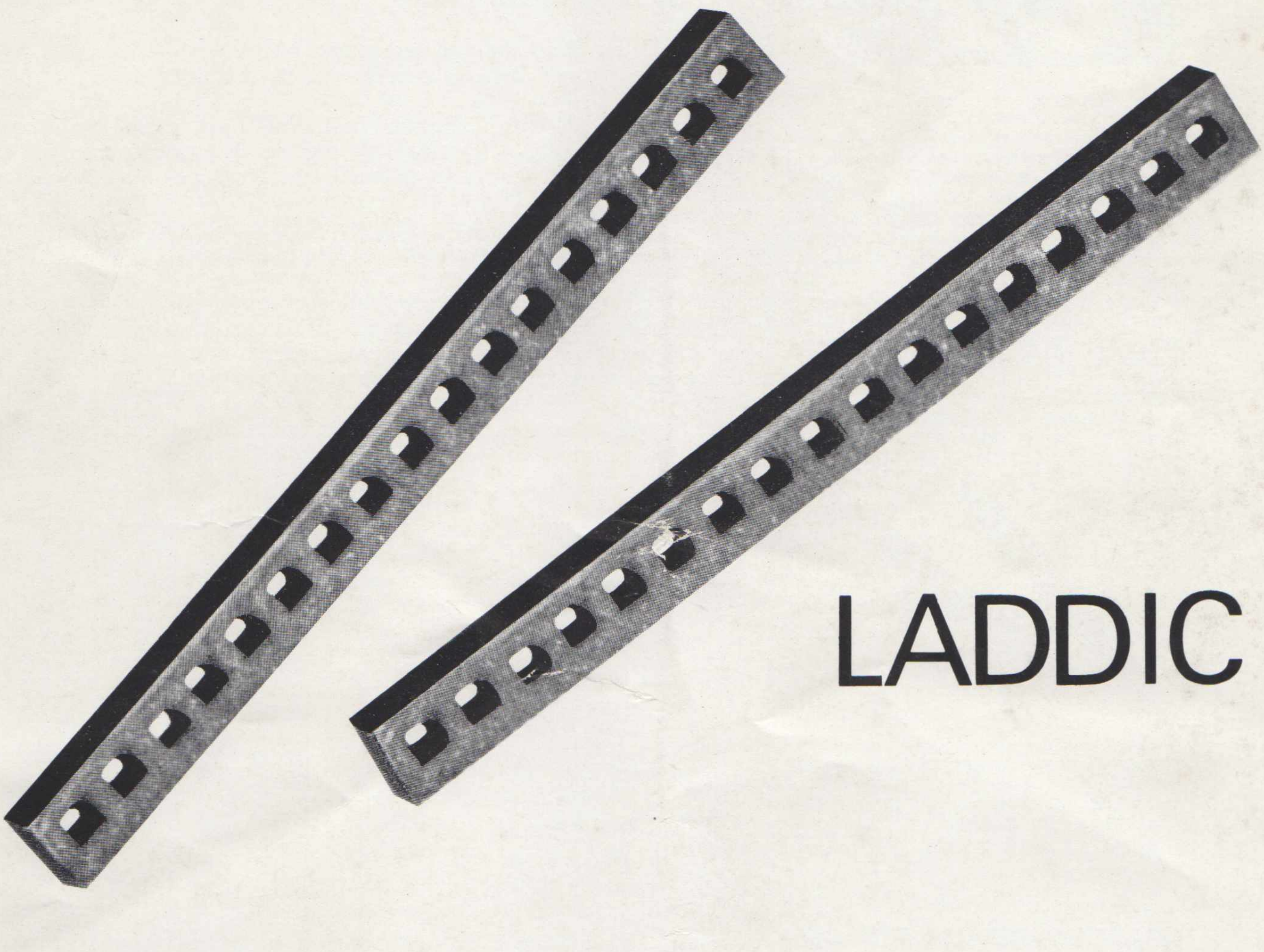
Mullard Outlook

AUSTRALIAN EDITION



LADDIC

NOVEMBER-DECEMBER, 1964
VOL. 7, No. 6

MULLARD-AUSTRALIA PTY. LTD.

VOL. 7 — No. 6 NOVEMBER-DECEMBER, 1964

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

Editor:
JOERN BORK

TABLE OF CONTENTS

Inherently robust and reliable, Laddic magnetic logic elements are suitable for many types of electronic control systems, including those designed to 'fail-safe'. Laddic is discussed on pages 74 and 75.

MULLARD-AUSTRALIA PTY. LTD.

35-43 CLARENCE STREET, SYDNEY
Phone: 29 2006

123-129 VICTORIA PDE., COLLINGWOOD, N.5
VICTORIA
Phone: 41 6644

Associated with
MULLARD LTD., LONDON

"At Christmas play and make good cheer
For Christmas comes but once a year."

THOMAS TUSSER.

The Australian Christmas scene is far removed from the traditional concept of yule log fires, frost and snow and robin redbreasts — nevertheless, the traditional roast turkey is still enjoyed *al fresco* on the beaches and picnic grounds.

How does the festive season affect you — the busy retailer who is enjoying, we hope, a steady volume of business?

You will leave others to bask in the sun while you're out on those last-minute calls — installing TV sets, servicing radio receivers and getting family record players in trim for the festivities.

But after you have shut up shop on Christmas Eve, there will be time for rest and relaxation — you will certainly have earned it.

There is a unanimous feeling of optimism within the Industry for the year ahead — and much to justify this. Enterprise and initiative will all play their part in the months ahead — the festival of Christmas is a good time to restore your energies for 1965.

We wish you a Merry Christmas and a Prosperous New Year!

EDITOR.

MULLARD DISTRIBUTORS

New South Wales

Martin de Launay Pty. Ltd.
Cnr. Druitt & Clarence Sts., Sydney. Phone: 29 5834

Cnr. King & Darby Streets, Newcastle. Phone: 2 4741

270 Keira St., Wollongong. Phone: 2 6020

21 Bayliss Street, Wagga. Phone: 4644

Victoria

Carnegie (Australia) Pty. Ltd.
Vere Street, Richmond. Phone: 42 2781

Queensland

C. A. Pearce & Co. Pty. Ltd.
33 Bowen Street, Brisbane. Phone: 2 3201

Western Australia

Tedco Pty. Ltd.
7 Fitzgerald Street, Perth. Phone: 28 4921

South Australia

Agents:
Woollard & Crabbe Limited
180 Wright Street West, Adelaide. Phone: 51 4713

Tasmania

Medhursts Wholesale Ltd.
163 Collins Street, Hobart. Phone: 2 2911

136 Wellington Street, Launceston. Phone: 2 2091

106 Wilson Street, Burnie. Phone: 1919

and all leading wholesalers throughout the Commonwealth

VIEWPOINT WITH MULLARD

"THE TELLY MAN"

"It's the small, everyday things that count. Like wiping your shoes on the mat, moving breakables out of your working area and spreading your dropcloth* on the carpet.

"Most technicians *do* watch these points, of course. But some don't and the customer is quick to notice . . . and remember."

These are the words of a typical specialist in radio and TV service in the home. We were talking in the spacious workshop adjoining his busy electrical store.

They don't see a lot of him here at base. For although he takes in a fair amount of work from retailers around, his reputation has been mainly built upon his ability to give skilled and reliable service without moving the set from the living room. How much this means to the customers can be judged from the success he enjoys.

Even the kids know him as "The Telly Man".

Scientific Approach

He set up in business on his own after a long experience of the radio trade which started with an apprenticeship and culminated in his appointment as manager to one of the larger service organisations.

He approached his solo venture scientifically — as he approaches most things — by carrying out a survey in the area in order to ensure that there was a real need and a market for the kind of "home service" he offered.

"It's the small, everyday things that count. Like wiping your shoes on the mat . . ."

Today he's handling some 3,000 calls a year. With his present resources that marks the peak. "When I'm ready to expand," he said, "I shall do it in my own way. It will mean a complete duplication of my present set-up. You see, I'm only interested in working with technicians who look at their job as I do. I'm only interested in sheer professionalism with no skimping."

A Sixth Sense

"The Telly Man's" service van is a familiar sight in the neighbourhood. It's equipped with a complete set of test gear, all the tools he's ever likely to need and, of course, Mullard valves, picture tubes and other components. ("I seldom use any other", he says. "You don't get so many call-backs with Mullard valves.")

"Through long experience," he told us, "I can usually pin down the area of a fault in about 15 minutes, employing a diagnostic technique rather like the doctor's. Have you ever watched how he asks you questions first, then studies you while you answer and finally examines you?

"Of course, the doctor has an advantage over me. His patients can talk back. Mine can't!"

For all that, as we saw, our friend relies heavily on the customer's evidence for his diagnosis. When did it happen? Has it happened before? How old is the set? When was it last serviced? These are the sort of questions he poses — with the polish of a barrister — before he touches a switch.

It may sound a lengthy process, but in fact he claims to give the fastest service so far as actual execution of the work is concerned.

About 95 per cent of his work is on a cash basis. "It means less accounting and fewer bad debts. My bad debt figure for last year was negligible."

12-Hour Day

With a "beat" which has a radius in excess of 12 miles, his working day is seldom shorter than twelve hours — and that doesn't include the hour spent on paper work in the morning. Like many who serve the public his phone calls are diverted from office to home out of normal business hours.

Extra Little Something

What tips does our friend offer to the service man whose work frequently takes him into customers' homes?

"It's largely a matter of common sense and common politeness. But there are one or two points which might seem trivial, but are in fact important. *Never* turn the set upside down without making sure that the chassis bolts haven't been loosened. *Always* remove at least one of the legs of a set when working on the base — it will give you more room. *Do* insure yourself against every possible kind of accidental damage. *Always* finish by dusting the set and giving the screen a polish. It adds that extra little something to a good service job, well done."

*These are obtainable from your local Mullard distributor.

MULLARD-AUSTRALIA PERSONALITIES

Mr. J. A. Kitchen

John Kitchen has been with the Sales Division of this Company for over seven years and, during that time, he has become well-known to the purchasing officers of our customers in the TV and Radio Manufacturing Industry.

Prior to joining the Mullard organisation, John travelled for twelve months throughout England and the Continent before making Sydney his permanent home. He was born and educated in England and originally came to this country twelve years ago. John enjoys the theatre and his weekends are spent surfing and playing an occasional game of golf.

Binders for Outlook

In response to a number of enquiries from regular Outlook readers, we have prepared a plastic binder designed to contain up to three volumes of Outlook. The method of securing the copies into the binder is unique and new to Australia, and individual copies may be removed and re-inserted with ease. A self-adhesive, interchangeable strip from which appropriate portions may be detached for titling purposes, will be supplied with each order, thus making the binder suitable for past and future issues.

The anticipated charge for these binders will be approximately £1-5-0d. If you are interested in this offer, please complete and return the enclosed pre-paid postcard, which is designed to assist us in estimating the total requirement. Please do *not* include any remittance.

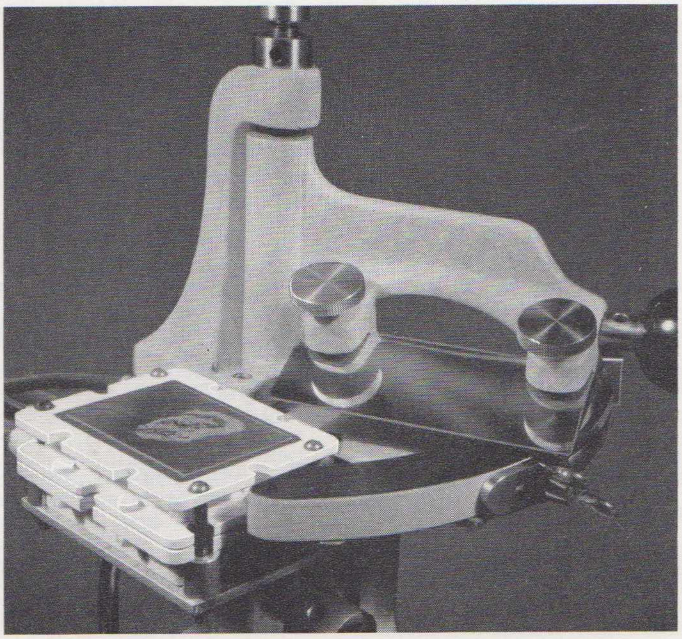
MULLARD PELTIER BATTERY AIDS RESEARCH

The Mullard Peltier battery PT47/5, designed for specialised cooling applications, has in this instance been adapted for use in the field of histology (the study of the minute structure of anatomical cells).

The battery has been attached to a freeze microtome which is used to prepare biological tissue for examination under a microscope, by firstly freezing the tissue and then cutting sections of desired thicknesses (5-50μ). The conventional freezing stage of this microtome was replaced by the Peltier battery. Normally, the freezing is effected by allowing liquid CO_2 to become gas under the metal stage which carries the tissue. The CO_2 freezing stage has two major disadvantages in that the temperature of the frozen tissue varies greatly and the cylinder containing CO_2 may have to be replaced during an experiment.

The Peltier battery PT47/5 contains 47 series-connected semiconductor thermo-elements*. The upper side of the unit is provided with a flat copper plate for "cold exchange" and works on the principle that, when a direct current is passed through a junction of semiconductor materials (bismuth telluride) a temperature gradient is established across the junction. Since the Peltier battery works on the cooling-gradient-principle, water is circulated through the battery so that the final temperature of the biological tissue is indirectly determined by the water temperature. The major advantage of this system is that the freezing of biological tissue can be kept at a constant temperature during sectioning.

* See also Outlook, Volume 7, Number 2, page 20.



A combination of microtome and Mullard Peltier Battery PT47/5

This picture is reproduced by kind permission of the Neurophysiological Laboratory, Psychiatric Research Unit, Callan Park, Rozelle, N.S.W.

THREE PLANAR TRANSISTORS MEET GENERAL APPLICATION REQUIREMENTS

With the introduction of three new n-p-n planar transistors—the BFY50, BFY51 and BFY52—the task of transistor-selection for a wide range of professional equipment is simplified.

These devices are truly general purpose, and may be used in many different circuit configurations in a wide range of amplifying, oscillating and switching applications.

A significant feature of the transistors is the exceptionally low saturation voltage. The values of 200mV for the BFY50 and 350mV for the BFY51 and BFY52, measured at a collector current of 150mA, are typical of germanium rather than silicon devices. The current gain is maintained over four decades of current, and the value of f_T, greater than 50Mc/s, enables most general-purpose applications to be readily met.

Instead of being faced with a multiplicity of type numbers for particular applications, the designer will be able to use a few general-purpose types for all but the most specialised applications. Limited quantities are currently available.

Brief data for the BFY50, BFY51, and BFY52 are given below.

Abridged Preliminary Data

	BFY50	BFY51	BFY52	
$V_{CB\ (max)}$	+80	+60	+40	V
$V_{CE(cut\text{-}off)}$	+80	+60	—	V
$V_{CE(sat)}$ (I_C=150mA)	+200	+350	+350	mV
h_{FE} (I_C=150mA)	>30	>40	>60	
f_T (V_{CE}=+6V, I_C=50mA)	>60	>50	>50	Mc/s
$P_{tot\ (max)}$ (T_{amb}=25°C)	800	800	800	mW
$T_{j\ (max)}$	200	200	200	°C

Three Mullard silicon n-p-n transistors for industrial and professional applications. Types BFY50, BFY51 and BFY52 are manufactured to the internationally standard size in TO-5 encapsulation.

HIGH Q-FACTOR COILS

Recent advances in equipment design in the communications industry have produced more stringent requirements for inductive components. Inductors have had to be made smaller whilst retaining the same value of Q-factor, or have had to yield appreciably higher Q-factors for the same size. These requirements can be met by the Mullard range of Vinkor adjustable pot-core assemblies which enables compact inductors with high Q-factors to be manufactured for operation at frequencies from 400c/s to 15Mc/s.

Definition of Q-Factor

The Q-factor of a coil is defined by the equation:

$$Q = \frac{\omega L}{R}, \qquad \ldots (1)$$

where ωL is the inductive reactance of the coil, and R is the total effective resistance, that is, the sum of all the core and coil losses. To achieve the highest Q-factor for a particular coil, R must be kept to a minimum. In other words, the losses associated with the coil and core must be as low as possible.

With the modern generation of inductors based on Ferroxcube pot-core assemblies, the loss factor is normally expressed in a form independent of frequency and inductance value. Thus:

$$\tan\delta_{tot} = \frac{R}{\omega L}, = \frac{1}{Q}, \qquad \ldots (2)$$

from Eq. (1) $\tan\delta_{tot}$ is the sum of the loss factors associated with the Ferroxcube core and the winding.

Losses in the Winding

For the winding, the losses consist of the DC loss, the eddy-current loss, and the dielectric loss caused by the self-capacitance.

DC Loss

The DC loss factor for the coil, $\tan\delta_{dc}$, is given by:

$$\tan\delta_{dc} = \frac{R_{dc}}{\omega L}, \qquad \ldots (3)$$

where R_{dc} is the DC resistance of the coil. This resistance can be minimised by choosing the largest possible core with the highest suitable permeability, and using the largest gauge of wire that will fill the bobbin. The DC loss predominates at the lower frequencies, and falls in proportion as the frequency rises until it equals about half the total loss at the frequency where the Q-factor is a maximum.

Eddy-Current Loss

The principal eddy-current loss is caused by the proximity effect. The loss factor, $\tan\delta_{p.e.}$ is given by:,

$$\tan\delta_{p.e.} = \frac{k_e\ f\ d^4 N n}{\mu_e}, \qquad \ldots (4)$$

where k_e is the copper eddy-current constant (quoted in the introductory notes for Vinkors in Vol. 6 of the Mullard Technical Handbook), f is the frequency in c/s, d is the wire strand diameter in cm, N is the number of turns, n is the number of strands in the conductor (for solid wire n = 1), and μ_e is the effective permeability.

There may be an additional eddy-current loss caused by skin effect. This loss factor, $\tan\delta_{s.e.}$, is given by:

$$\tan\delta_{s.e.} = \frac{R_{s.e.}}{\omega L}, \qquad \ldots (5)$$

where $R_{s.e.}$ is the increase in the resistance of the winding caused by skin effect. Skin effect is significant in pot-core inductors only when they are wound with bunched conductors of large over-all diameter. The loss has the effect of lowering the peak of the Q-factor plotted against frequency curve.

The eddy-current losses become noticeable above 30kc/s, and can be minimised by using bunched conductors. The higher the operating frequency, the smaller should be the strand diameter.

Dielectric Loss Through Self-Capacitance

The dielectric loss factor $\tan\delta_{c.d.}$, caused by the self-capacitance of the winding, is given by:

$$\tan\delta_{c.d.} = LC_s\ \tan\delta_d \qquad \ldots (6)$$

where C_s is the total self-capacitance of the winding (normally between 5 and 50pF depending on the core size), and $\tan\delta_d$ is the loss factor of the associated dielectric. This loss can be minimised by using sectioned polystyrene bobbins and possibly by spacing off the winding from the centre of the core.

Losses in the Core

The losses in the core are made up from the residual and eddy-current loss, and the hysteresis loss.

The residual and eddy-current loss is controlled during the production process of Vinkors, and the value of this loss factor, $\tan\delta_{(r+e)}$, is shown on the individual Vinkor data sheets. A significant reduction in this loss has been achieved with the Violet range of Vinkors.

The hysteresis loss factor, $\tan\delta_h$, is given by:

$$\tan\delta_h = \frac{F_h\ I\ L^{\frac{1}{2}}}{2\pi}, \qquad \ldots (7)$$

where F_h is the hysteresis factor shown on the individual Vinkor data sheets, I is the r.m.s. current in amperes, and L is the inductance in henrys. Normally this current is negligibly small and the hysteresis loss may be neglected.

Evaluation of Q-Factor

When all the losses have been evaluated, the Q-factor for a particular inductance and frequency can be found from the equation:

$$Q = \frac{1}{\tan\delta_{tot}}, \qquad \ldots (8)$$

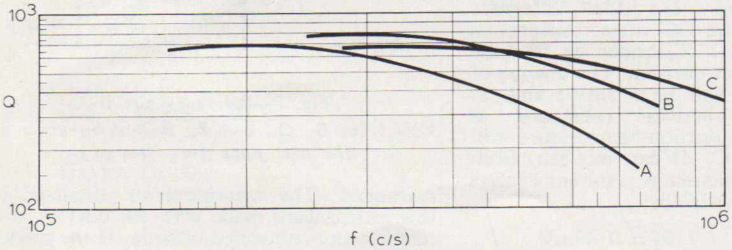
derived from Eq. (2).

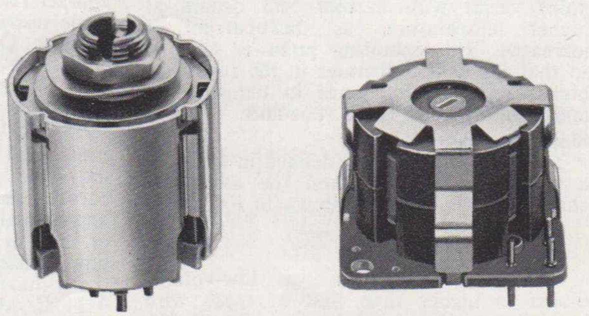
Other factors such as size and temperature coefficient influence the final design; these normally confirm the choice of a suitable Vinkor for the inductor before commencing the coil design.

Comprehensive design data and relevant information on Vinkor pot cores and Ferroxcube transformer cores may be found in Volume 6 of the Mullard Technical Handbook and in the Mullard publication "Vinkor Manual" available from Mullard offices and distributors throughout the Commonwealth, priced at 5/3d plus 8d postage.

LA2211

Curve	Coil former	No. of turns	Wire S.W.G.	Wire Insulation	Inductance
A	DT 2011	72	100/48	E.D.S.	574 μH
B	DT 2011	42	180/48	E.D.S.	195 μH
C	DT 2011	22	350/48	E.D.S.	54 μH





The core assembly LA2208 is shown on the left, mounted into the conventional Vinkor housing; and on the right, assembled into the low-cost clip and base available for insertion into printed wiring.

SIMPLE VOLTAGE REFERENCE CIRCUIT

A circuit consisting of three zener diodes and two resistors provides a reference voltage of $\simeq$ 5·3V stable within $\pm$0·3% over a temperature range 0 to 50°C, with mains voltage variations of up to $\pm$10%. Long-term drifts in the components may give a further voltage change of 0·012%.

The Mullard BZY78 is a silicon reference diode with a zener voltage of 5·1 to 5·6V. It has a voltage stability of $\pm$1% at a zener current of 11·5mA$\pm$10%, in ambient temperatures between −50 and +100°C. The investigation discussed in this article determined how the voltage stability could be improved by stabilising the zener current and restricting the temperature range.

Temperature Coefficient

There are two reverse breakdown mechanisms exhibited by semiconductor diodes. Low resistivity material, which has a high level of doping, gives very narrow p-n junctions which break down by internal field emission. The original theory of this form of breakdown is due to Zener, and for this reason all diodes with this characteristic are called zener diodes. In material with a lower level of doping, avalanche breakdown, due to ionisation by collision, will occur at a lower voltage than that required for zener breakdown. Thus, at a specific doping level there is a transition from true zener to avalanche breakdown. This point also depends upon the current density, the junction temperature, and the presence of any impurity in the junction.

The zener breakdown voltage is inversely proportional to the carrier concentration in the higher resistance side of the junction. This concentration increases with junction temperature, giving a negative temperature-coefficient of voltage. Conversely, the avalanche breakdown, which is the effect of ionisation of the valence electrons in the silicon, is critically dependent upon electron mobility, which decreases with a rise in temperature. As a higher voltage must then be developed to maintain the same rate of ionisation, the avalanche mechanism has a positive temperature-coefficient.

Breakdown, which depends upon the current and the junction temperature, may therefore be zener or avalanche or a combination of the two. The BZY78 reference diode operates in the transition region (at about 5·3V) and exhibits high stability with temperature changes.

The coefficient of breakdown voltage with temperature is, for the BZY78, negative at high temperatures (zener) and positive at low temperatures (avalanche). The inversion point varies with current and occurs at higher temperatures as the current is increased. The operating point is selected so that the device operates in the region of high stability with change in temperature, and also has a low slope resistance.

Slope Resistance

The slope resistance is defined in terms of a small change in I_Z and the associated change in V_Z. Thus, as shown in Fig. 1,

$$r_Z = \Delta V_Z / \Delta I_Z$$

The slope resistance of low-voltage diodes is higher than that of types with breakdown voltages of between 7 and 8V, in which region r_Z has a minimum value. Beyond this region the slope resistance increases at a greater rate than the nominal voltage.

The slope resistance also varies with zener current, decreasing exponentially as the current I_Z is increased. It is virtually independent of temperature.

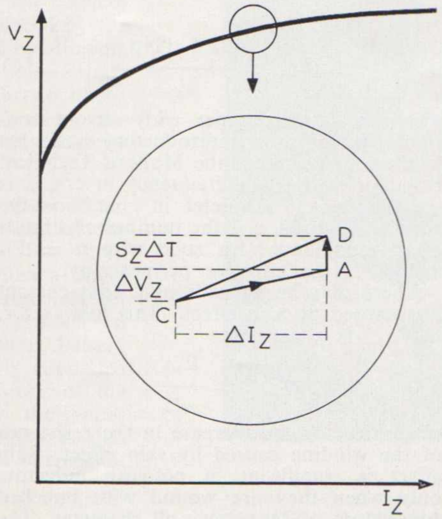


Fig. 1—Dynamic characteristic of zener diode, with small section of curve magnified.

The AC slope resistance will, in general, be different from the DC slope resistance at any point since a change in dissipation due to DC will also change the voltage, unless the temperature coefficient of voltage is zero at that point. In Fig. 1 a change of current ΔI_Z produces an instantaneous change of voltage ΔV_Z, and the operating point moves from C to A along the dynamic characteristic. The temperature of the device will then change by an amount ΔT because of the increased dissipation, and there is a further change of voltage $S_Z\Delta T$, where S_Z is the temperature-coefficient of voltage. When thermal stability is achieved, the operating point will have reached D, which is the point that would have been reached by increasing the current very slowly and moving along the line C-D (the static characteristic). Hence the small-signal AC slope resistance will be less than the corresponding DC value when the temperature-coefficient S_Z is positive, and greater than the DC value when the coefficient is negative.

The BZY78 has a very low temperature-coefficient of voltage (see *Design*). It may be deduced that the difference between the AC and DC slope resistance is approximately 0·5Ω. Consider an increase of 1mA in the zener current. The change of power dissipation will be 5·3mW; and, as the maximum thermal resistance is 0·31°C/mW, the junction temperature will increase by 1·64°C. It can be seen, from *Design*, that the maximum resultant change of zener voltage will be

$$\pm(7{\cdot}5/25) \times 1{\cdot}64 = 0{\cdot}49\text{mV}$$

which represents a resistance change of 0·49Ω. This difference between the AC and DC slope resistance is small compared with 20Ω, which is the nominal dynamic resistance of the BZY78, and it will be neglected.

REFERENCE CIRCUIT

It will be seen that the principal causes of drift in the voltage of a zener diode are changes in reverse current and junction temperature.

Regulation

Each stage of the reference circuit (Fig. 2) may be considered to be a potential divider consisting of a resistor and diode in series, whose regulation is defined as the ratio of the change in output voltage to the associated small change of input voltage. Thus,

$$\text{Regulation} = \frac{\Delta V_{out}}{\Delta V_{in}}.$$

An expression is derived in the Appendix for the regulation in terms of the source and slope resistances in the circuit, and the overall regulation is the product of the regulations of the successive stages. The reference is designed to be operated from a mains-derived supply which may vary as much as $\pm$10%, therefore the circuit must have good regulation.

Temperature

If the reference is used over the range 0 to 50°C, and standardisation is carried out at room temperature, this interval may be expressed as 25°C$\pm$25°C. Hence, in the worst case, away from the condition of optimum stability, (the variation of voltage with temperature being approximately linear) this change may be represented by $\pm\Delta V_t$, where ΔV_t is the change in zener voltage over 25°C. At the condition of optimum temperature stability, where the variation is non-linear, ΔV_t will be less than the worst-case value.

The junction temperature will differ from ambient by an amount depending upon the power dissipation in the diode (which may be considered constant) and the thermal resistance. The measurements discussed in this article were made with the devices in a temperature-controlled oilbath. If the diodes

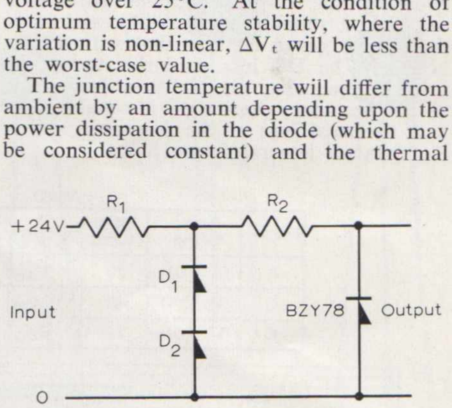


Fig. 2—Reference circuit
Values for D_1, D_2, and R_1, R_2 are discussed on the next page (see 'Design').

the next page (see 'Design')

This article is based on a report by J. F. Miles and A. J. Tumber, Mullard Semiconductor Measurement and Application Laboratory.

are mounted in air, thus increasing the thermal resistance, the temperature difference will be greater, giving a slightly different operating temperature range. For example, if the device is mounted by its leads in free air at a certain temperature, the junction temperature may be up to 10°C higher than when the device is immersed in oil at the same temperature.

Design

The overall regulation improves with the number of stages, but the temperature-coefficient of each stage, although subject to the regulation of succeeding stages, has an adverse effect on the stability. A three-stage reference circuit may be designed to only small advantage, therefore a two-stage circuit is considered in this article.

The first stage must provide an output voltage much greater than 5·3V with a low temperature-coefficient, and it must have good regulation. A suitable output voltage with a low temperature-coefficient may be achieved by using two diodes in series, each with a breakdown voltage of approximately 5V. A very low slope resistance and a high source resistance are necessary (see Appendix) to give good regulation; and so the operating current, and hence the supply voltage, should be high.

Zener diodes from the BZY88-C series were chosen for the first stage (D_1, D_2). The 5·1V diode type BZY88-C5V1 has a low temperature-coefficient, and two diodes in series give a typical combined voltage of $2 \times 5{\cdot}35 = 10{\cdot}7$V when run at 20mA, which is high compared with the voltage of the BZY78 (5·3V). At this current it is estimated that each BZY88-C5V1 has a worst-case temperature coefficient of +0·9mV/°C, and a slope resistance of 10·5Ω. With a 24V supply, which is high compared with 10·7V, and allowing for a zener current of 11·5mA into the BZY78, the values of R_1 and R_2 are 430Ω and 470Ω respectively. The regulation is 0·045 in the first stage and 0·041 in the second, as calculated in the Appendix.

The drift of the BZY78 reference diode was found to be less than ±7·5mV over the range 25° ±25°C under conditions of constant drive.

The stability of the reference also depends upon the resistors in the circuit. The variation in resistance value with operating conditions was studied with reference to the Welwyn C22 Panclimatic type, and it was found that the predominant factor was the temperature-coefficient. Self-heating and the voltage coefficient contribute further small variations, giving an estimated total of 0·63% for the resistors in the first stage, and 0.60% in the second.

Calculation

The maximum deviation in the output from each stage may be calculated as follows, using the regulation factors deduced in the Appendix.

First stage:

Supply
$\pm 10\% \times (24 \times 10^3) \times 0{\cdot}045 = \pm 108$mV

Resistor R_1
$\pm 0{\cdot}63\% \times (13{\cdot}3 \times 10^3) \times 0{\cdot}045 = \pm\ 3{\cdot}8$mV

Diodes D_1, D_2
$\pm 25 \times (2 \times 0{\cdot}9) = \pm\ 45$mV

Total first stage deviation = ±156·8mV

Second stage:

First stage drift
$\pm 156{\cdot}8 \times 0{\cdot}041 = \pm\ 6{\cdot}43$mV

Resistor R_2
$\pm 0{\cdot}60\% \times (5{\cdot}4 \times 10^3) \times 0{\cdot}041 = \pm\ 1{\cdot}33$mV

BZY78 drift = ± 7·5mV

Total deviation in output = ±15·26mV

This total deviation represents ±0·3% of the minimum nominal reference voltage.

Alternative Design

With a 5·6V diode type BZY88-C5V6 substituted for each BZY88-C5V1, the same overall performance will be obtained if the values of R_1 and R_2 are changed to 390Ω and 530Ω respectively. This design has the advantage of better regulation at the expense of temperature stability; so if it is necessary to use the reference over a wider range of input voltage, this second design is preferable.

Long-term Drift

The reference voltage will be subject to the long-term stability of the resistors and diodes. The resistance values may vary as much as a further 0·15%, and the diodes are subject to a maximum drift of 200μV. These drifts may be in addition to those already calculated, giving a further possible drift of 0·012% from the initial voltage.

APPENDIX TO "SIMPLE VOLTAGE REFERENCE CIRCUIT"

The stage regulation may be calculated with reference to the equivalent circuit, Fig. 3, where

$$V_{in} - V_1 = R_1(i_1 + i_2) \qquad \ldots(1)$$

and

$$V_1 - V_2 = i_2 R_2. \qquad \ldots(2)$$

From Eqs(1) and (2)

$$V_{in} = V_1 + i_1 R_1 + (V_1 - V_2)(R_1/R_2)$$

or

$$V_1[1 + (R_1/R_2)] = V_{in} - i_1 R_1 + V_2(R_1/R_2)$$

therefore

$$\frac{dV_1}{dV_{in}}\left(1 + \frac{R_1}{R_2}\right) = 1 - R_1\frac{di_1}{dV_{in}} + \frac{R_1}{R_2}\cdot\frac{dV_2}{dV_{in}} \qquad \ldots(3)$$

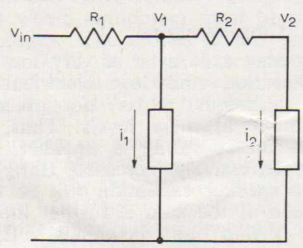


Fig. 3—Equivalent circuit.

Now

$$\frac{di_1}{dV_{in}} = \frac{di_1}{dV_1}\cdot\frac{dV_1}{dV_{in}} = \frac{1}{r_1}\cdot\frac{dV_1}{dV_{in}} \qquad \ldots(4)$$

where r_1 is the slope resistance of the first zener voltage at the current i_1.

Also, from Eq(2),

$$V_2 = V_1 - i_2 R_2$$

therefore

$$\frac{dV_2}{dV_{in}} = \frac{dV_1}{dV_{in}} - R_2\frac{di_2}{dV_{in}}$$

$$= \frac{dV_1}{dV_{in}} - R_2\frac{di_2}{dV_2}\cdot\frac{dV_2}{dV_{in}}$$

If r_2 is the slope resistance of the second zener voltage at the current i_2, we have

$$\frac{dV_2}{dV_{in}} = \frac{dV_1}{dV_{in}} - \frac{R_2}{r_2}\cdot\frac{dV_2}{dV_{in}}$$

therefore

$$\frac{dV_2}{dV_{in}} = \frac{1}{1 + (R_2/r_2)}\cdot\frac{dV_1}{dV_{in}}$$

$$= \frac{r_2}{r_2 + R_2}\cdot\frac{dV_1}{dV_{in}} \qquad \ldots(5)$$

From Eqs(3), (4), and (5)

$$\frac{dV_1}{dV_{in}}\left(1 + \frac{R_1}{R_2}\right)$$

$$= 1 - \frac{R_1}{r_1}\cdot\frac{dV_1}{dV_{in}} + \frac{R_1}{R_2}\cdot\frac{r_2}{r_2 + R_2}\cdot\frac{dV_1}{dV_{in}}$$

or

$$\frac{dV_1}{dV_{in}}\left(1 + \frac{R_1}{R_2} + \frac{R_1}{r_1} - \frac{R_1}{R_2}\cdot\frac{r_2}{r_2 + R_2}\right) = 1$$

Simplifying, we obtain

$$\frac{dV_1}{dV_{in}} = \frac{1}{1 + \dfrac{R_1}{r_1} + \dfrac{R_1}{r_2 + R_2}} \qquad \ldots(6)$$

The regulation of the second stage may be calculated with this equation by substituting R_2 for R_1, and r_2 for r_1, and by making R_2 infinite, giving

$$\frac{dV_2}{dV_1} = \frac{1}{1 + (R_2/r_2)} \qquad \ldots(7)$$

In this case the circuit values are as follows:

$R_1 = 430\Omega$ $\quad R_2 = 470\Omega$
$r_1 = 21\Omega$ $\quad r_2 = 20\Omega$

Then, from Eq(6),

$$\frac{dV_1}{dV_{in}} = 0{\cdot}045$$

and, from Eq(7),

$$\frac{dV_2}{dV_1} = 0{\cdot}041\,.$$

MICROPHONY IN TAPE RECORDERS

Howlback and clang microphony are the two most common types of interference experienced with a valve tape recorder. This article defines these terms and suggests design limits in the manufacture of valve tape recorders whereby these two types of microphony can be reduced to a satisfactory level.

INTRODUCTION

Microphony is occasionally experienced in a tape recorder where high gain AF valves are used in the first stages. There are two types of microphony: howlback and clang microphony. The two types of interference are defined and discussed in detail under separate headings.

HOWLBACK

Howlback, or feedback howl, is an oscillation which occurs in the following way. Any excitation force may cause electrode vibrations in a valve resulting in an initial microphony (Ref. 1). The initial microphony of a pre-amplifier valve is then amplified and converted by the speaker into vibrations. These vibrations are fed to the valve acoustically, mechanically, or by a combination of both, in such a manner as to sustain the original electrode oscillations which produced the initial microphony.

Requirements

As implied in the definition of howlback, this type of microphony occurs only if the valve is excited with a force of suitable phase and of sufficient amplitude. Therefore, even with a suitable phase relationship, howlback will not occur if the equivalent microphonic grid voltage of the pre-amplifier valve (resulting from a specified speaker power) is less than the signal input voltage to the same valve required to produce the specified speaker power. This means that, for a speaker power of 1W and with a safety factor of two, the equivalent microphonic grid voltage $V_{eq(1W)}$ resulting from a speaker power of 1W, should be less than half the signal input voltage $V_{in(1W)}$ required for a speaker power of 1W. That is,

$$V_{eq(1W)} \leqslant \tfrac{1}{2} V_{in(1W)}$$

or, since $V_{eq(1W)} = V_{eq(1g)}\ g_{1W}$ and the speaker power is proportional to the square of the output or input voltage (or the input voltage is proportional to the square root of the speaker power),

$$V_{eq(1g)} \cdot g_{1W} \leqslant \tfrac{1}{2} V_{in(0.05W)} / \sqrt{0 \cdot 05}\ ,$$

or

$$V_{eq(1g)} \cdot g_{1W} / V_{in(0.05W)} \leqslant 2 \cdot 2 \quad \ldots (1)$$

where

$V_{eq(1g)}$ = equivalent microphonic grid voltage of the pre-amplifier resulting from a peak acceleration of 1g,

g_{1W} = peak acceleration, in g, of the pre-amplifier valve resulting from a speaker power of 1W,

$V_{in(0.05W)}$ = signal input to the pre-amplifier valve required for a speaker power of 0·05W.

Design Limits

Valves which are intended for use in tape recorder input stages are designed with a $V_{eq(1g)}$ value such that they meet the requirement in Eq (1). This requirement applies in a tape recorder having a typical $g_{1W}/V_{in(0.05W)}$ ratio.

A typical ratio corresponds to the maximum possible values which occur in reasonably well-designed tape recorders. Normally, the valve published data give the recommended or minimum permissible value of $V_{in(0.05W)}$.

Maximum permissible average acceleration of the valve resulting from a speaker excitation of 50mW can be considered as 0·015g r.m.s. at frequencies higher than 500c/s, and 0·06g r.m.s. at frequencies lower than 500c/s. Considering that the maximum r.m.s. acceleration is equal to approximately three times the average r.m.s. acceleration (or $1/\sqrt{2}$ times the peak acceleration) and that the acceleration is proportional to the square root of the speaker power (Ref. 1), the maximum value of g_{1W} which should be used with the minimum $V_{in(0.05W)}$ is $3 \times 0 \cdot 015 \sqrt{2} / \sqrt{0 \cdot 05} = 0 \cdot 28\text{g}$ for frequencies above 500c/s and $3 \times 0 \cdot 06 \sqrt{2} / \sqrt{0 \cdot 05} = 1 \cdot 1\text{g}$ for frequencies lower than 500c/s.

Therefore, the maximum value of $g_{1W}/V_{in(0.05W)}$ is 0·28/recommended $V_{in(0.05W)}$ for frequencies above 500c/s and 1·1/recommended $V_{in(0.05W)}$ for frequencies lower than 500c/s. Consequently, howlback microphony can reasonably be expected to result from the unsuitable design of the equipment if the $g_{1W}/V_{in(0.05W)}$ ratio is greater than the above-mentioned values.

CLANG

Clang is an audible interference caused by any external forces such as speaker, motors or especially switch excitations.

General Considerations

Clang may exist even at very low values of acceleration and low electrical sensitivities, although it is not necessarily objectionable at these levels. Thus, in an amplifier incorporating a speaker, if the gain is progressively increased, clang resulting from speaker excitation will be noticed first. Then, if the gain is further increased, up to a certain limit, howlback will occur.

In principle, the relative value of the interference with respect to the useful signal level determines whether the interference is negligible or not. With motor or switch excitation, the useful signal can be zero for a given interference. Therefore, ideally, the quality of motors, switches, and their mountings in the equipment should be such that the interference resulting from the excitation of these sources, with a useful signal level of zero, is not objectionable.

Generally, the clang resulting from switch excitation is more objectionable than that resulting from speaker or motor excitation. Therefore, switch excitation clang is discussed in detail and some practical requirements given.

Switch Excitation

It is important to distinguish between clang and mechanical interference. Clang is the electronic interference which is produced in the pre-amplifier valve and, after being amplified by the following stages, is audible from the loudspeaker. Mechanical interference is the sound of the switch operation as heard directly by the ear.

In the playback position the requirements for clang can be derived as follows. If the duration of the clang t_e is less than the duration of the mechanical interference t_m, the equivalent microphonic grid voltage $V_{eq(SW)}$, due to switch excitation, should be less than the signal input voltage $V_{in(P)}$ required for a speaker power of P watts, where P is the speaker output which has the same interference effect on the ear as that of the mechanical interference. That is,

$V_{eq(SW)} \leqslant V_{in(P)}$, or since $V_{eq(SW)} = V_{eq(1g)}\ g_{SW}$ and the input voltage is proportional to the square root of the speaker power,

$$V_{eq(1g)} g_{SW} \leqslant V_{in(0.05W)} \sqrt{\frac{P}{0 \cdot 05}}\ ,$$

$$V_{eq(1g)} g_{SW} / V_{in(0.05W)} \sqrt{P} \leqslant 4 \cdot 5 \quad \ldots (2)$$

where g_{SW} = peak acceleration in g of the pre-amplifier valve resulting from the switching action.

If the duration of the clang is longer than that of the mechanical interference, experiment indicates that the ratio of the signal input voltage required for the maximum available speaker power $V_{in(Po)max}$ to the equivalent microphonic grid voltage $V_{eq(SW)}$ should be greater than 24dB (16 times). That is,

$$V_{eq(SW)} / V_{in(Po)max} < 1/16$$

or

$$\frac{V_{eq(1g)} \cdot g_{SW}}{V_{in(0.05W)} \sqrt{\dfrac{P_o max}{0 \cdot 05}}} < 1/16$$

or

$$V_{eq(1g)} \cdot g_{SW} / V_{in(0.05W)} \sqrt{P_{o(max)}} < 0 \cdot 3 \quad \ldots (3)$$

Poor mechanical design of the equipment can produce a ringing effect and make the duration of the clang longer than that of the mechanical interference. Therefore, the problem of ringing effects should be taken into account at the design stage of the tape recorder. Thus the duration of clang would not exceed the duration of the mechanical interference and consequently, the rather rigid requirements of Eq (3) need not be considered.

Design limits for clang microphony can be derived simply, as discussed in the results given for howlback by using Eq (2) and (3) with the following typical values.

PERFORMANCE OF THREE TYPICAL TAPE RECORDERS

Measurements have been made on three different types of commercially available tape recorder.

Vibration

The acceleration of the pre-amplifier valve resulting from a speaker power of 50mW was measured, in three mutually

TWO NEW LOW-COST SILICON P-N-P TRANSISTORS

Mullard have introduced two new high-voltage, medium-gain silicon transistors, the BCY54 and OC207, which have been developed from the highly successful BCY38 and OC204 series. They are particularly suitable for pulse and audio applications, compact DC converters, servo process control, power switching and relay drivers.

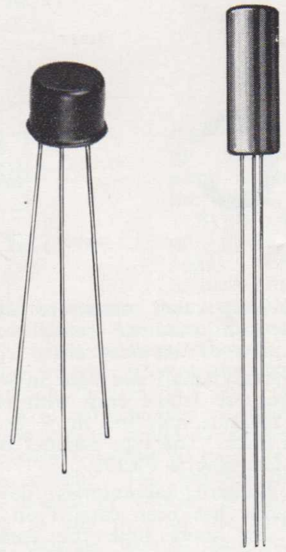
Having a typical gain of 50 and a collector voltage of 50V, the BCY54 and OC207 will find special application in equipment where the higher voltage rating of the OC205 and BCY39 is not required, yet where a higher gain would be an advantage.

Uprated BCY38 Series

The BCY54, in TO-5 encapsulation, joins the recently uprated BCY38 series which now has a power dissipation of 400mW in free air at 25°C and 500mW at a case temperature of 88°C. Sample quantities will shortly be available.

Abridged Data
for BCY54 and OC207

$V_{CB\,(max)}$ (I_E=0mA)	−50V
$V_{CE\,(max)}$ (cut-off)	−50V
h_{FE} (I_C=30mA, V_{CE}=−2V)	50 typical
h_{FE} (I_C=150mA, V_{CE}=−1V)	25 typical 12 minimum
f_T (V_{CE}=−6V, I_C=1mA)	2·0Mc/s typical



THYRISTOR FOR COMPUTER PRINT-OUT EQUIPMENT

The BTX18 has been specially introduced for high-speed applications. It is designed for use in fast line-and-time print-out devices for computers.

Fast Print-out Thyristor

A device with a very high gain and fast operating time is required between the logic circuits of a computer and the print-out equipment. The BTX18 can provide sufficient current to operate the print-hammer solenoid directly from the gating pulse produced by a transistor or defluxing core. The maximum value of gate current required to fire the thyristor is 15mA with a pulse length of 6μs.

The BTX18 is in TO-5 encapsulation for ease and economy of mounting. Because of the short duty cycle encountered in this application (about 5%) the device will handle peak currents of 3·5A without exceeding the dissipation limit. The rapid rise in current necessary to operate the print-hammer solenoid quickly requires a high driving voltage and the BTX18 has a 150V rating to enable it to be used with wide voltage margins.

Although specifically designed for print-out applications, the BTX18 can also be used for other applications such as DC inverters and for firing other thyristors where a high current is required for a short time.

The BTX18 is available in sample quantities.

NEW TYPE NUMBERING SYSTEM FOR THYRISTORS

A new type numbering system has been adopted for the Mullard high-voltage thyristor range and will gradually be introduced for the remaining ranges.

Thyristors with the same current rating are given a type number consisting of the letters B (denoting silicon), T (denoting thyristor), and a letter and two figures as a serial number. Individual types within the series are identified by a group of figures which represent the maximum repetitive peak reverse voltage of the device. A final letter R indicates a 'reverse-polarity' device, that is, the anode is connected to the stud, e.g. "BTY87-400R".

ZENER DIODES FOR SHUNT STABILISER CIRCUITS

The BZY91 series of zener diodes has been introduced for use in shunt stabiliser circuits, and is particularly useful in applications where equipment must be protected from surges. The diodes in this series will dissipate 75W at a stud temperature of 65°C and the surge ratings are 4·4kW for 100μsec and 170W for 100msec.

The range of zener voltages is from 10 to 75V with nominal zener voltage tolerance of ±5%. The diodes use stud mounting and are in DO-5 encapsulation. Reverse polarity versions will be available later.

MULLARD PUBLICATIONS

A revised, up-to-date edition of the Mullard Film and Filmstrip Catalogue is available from Mullard offices on receipt of a stamped, self-addressed, foolscap envelope endorsed "Films".

Also available from Mullard offices and distributors throughout the Commonwealth are reprints of the two popular Mullard publications "Stereo Sound Systems" and "Vinkor Manual" priced as follows:—
"Stereo Sound System": 6/3 plus 8d. postage
"Vinkor Manual": 5/3 plus 8d. postage

Continued from page 72

perpendicular directions, by a Brüel and Kjaer accelerometer, type 4329, mounted on the valve socket. The peak accelerations caused by a 0·05W speaker power in the frequency spectra above 500c/s and below 500c/s were less than 0·05g and 0·15g respectively. The peak accelerations g_{1W} corresponding to a speaker power of 1W in the same frequency spectra were less than 0·2 and 0·7g respectively. The peak-to-peak acceleration (that is $2g_{SW}$) resulting from the operation of switches was between 0·5 and 2g, with durations of 15 to 60ms and frequencies up to 500c/s.

Electrical Sensitivity

In the playback position, the value of the input voltage $V_{in(0\text{-}05W)}$, in the frequency range 50 to 10 000c/s, varied with frequency and the individual equipment from 0·1 to 0·8mV. The maximum available speaker power $P_{o(max)}$ varied between 1 and 3W.

Howlback, Clang and Mechanical Interference

Howlback and clang tests were carried out in the playback position with the volume control at maximum. No interference could be detected by ear.

Mechanical interference and clang resulting from switch operations were measured with a Peekel Sound Spectrometer, type GR3, and an oscilloscope. The two sources of interference were isolated with a sound-proof box, and the following results obtained:

Duration of mechanical interference t_m = 50 to 200ms.

Duration of clang t_c = 20 to 30ms.

The relative amplitude of clang with respect to the amplitude of the mechanical interference, as sensed by the spectrometer = 0·2 to 0·7.

The speaker power required to produce the same interference effect on the spectrometer as that of the mechanical interference of the switching action = 0·9 to 2·5W.

Thus, in the three tape recorders examined, it was found that clang and howlback microphony were within the limits given in this article.

REFERENCES

1. Dagpunar, S. S. 'Valve Microphony'. **Mullard Technical Communications,** Vol. 7, No. 61, Nov. 1962, pp. 2 to 17.

S. S. Dagpunar
Mullard (U.K.) Applications Research Laboratory.

LADDIC

LADDer logIC

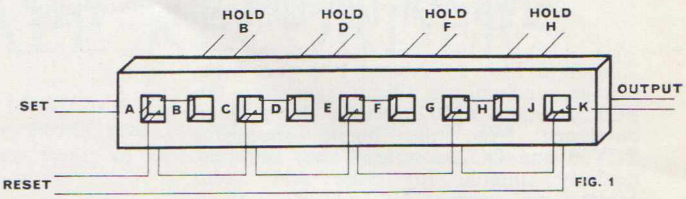


FIG. 1

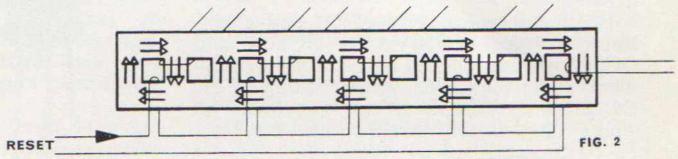


FIG. 2

Since electronic controls first became practicable, reliability has been a major factor in determining how readily — or even whether—industry has accepted them as aids to productivity. It is evident by the number of industrial applications utilising electronics, that designers of components and systems are attaining the required standards.

Insistent demands for greater reliability have forced designers away from relays and other devices with moving parts. Instead they have turned to the inherently reliable solid state devices as a means of performing switching and logical operations in control systems. In doing so they have gained an additional advantage, for with solid state or static switching, it is easier to design systems that 'fail-safe'— an essential requirement in applications like nuclear power stations and railway signalling installations.

In 1958 a new type of solid state device was added to the system designer's armoury by the Bell Laboratories. Known as Laddic, it makes use of the square-loop magnetic properties of ferric materials (such as Ferroxcube) to provide, in an almost indestructible form, a logic element of great intrinsic reliability able to operate in systems with fail-safe characteristics.

Laddic is pressed from Ferroxcube into the shape of a ladder (hence LADDer-logIC). Its only other parts are simple, often single-turn windings round certain rungs and parts of the side rails.

It is extremely small and light in weight a general-purpose laddic core with eleven rungs may measure only 0·8 in × 0·13 in × 0·08 in thick. The type number of the Mullard laddic core is FX2717.

In the Mullard laboratories development of laddic has been carried on since 1959, and for some time the company has manufactured cores so that designers might experiment with the new device. During this period when practical manufacturing techniques were being worked out, applications research was begun to discover the best ways of applying laddic to the solution of practical control problems. Its use in reactor safety systems was the subject of one study, undertaken recently.

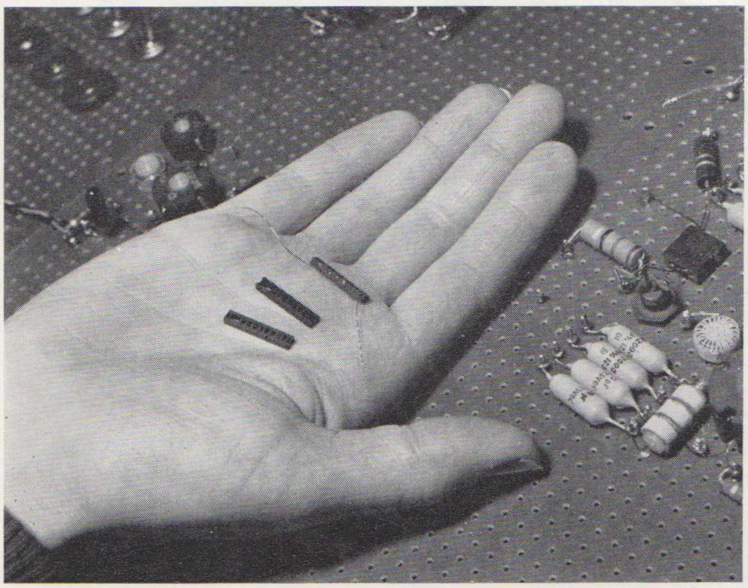
A laddic core may be wound in many different configurations according to the job it is to do. One of the simplest arrangements, that to perform the logic function of AND, is also the simplest by which to explain how laddic works.

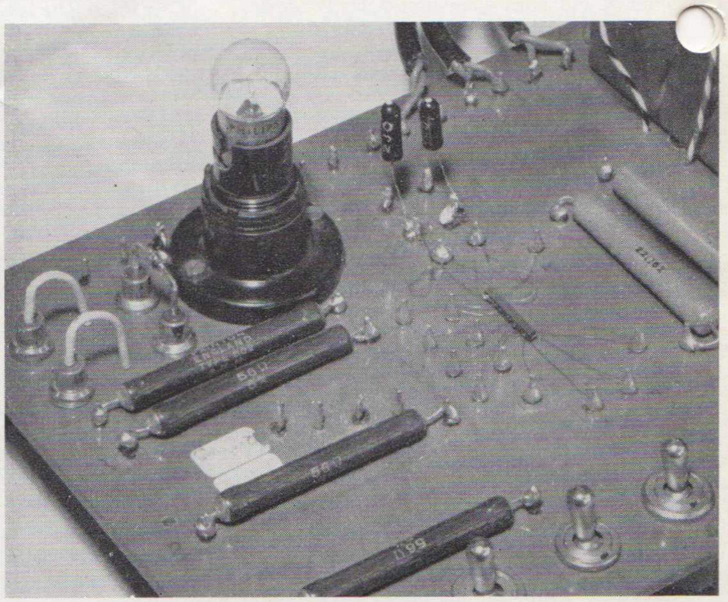
As Fig. 1 shows the laddic core has a 'set' winding round rung A, a 'reset' winding, which threads every alternate aperture, 'hold' winding round rungs B, D, F and H, and an output winding round rung K.

A pulse of current through the reset winding magnetically saturates the whole core, setting up a flux pattern as shown by the arrows in Fig. 2. A second pulse, of opposite polarity to the reset pulse, is then applied to the set winding on rung A. The set pulse tries to reverse the flux through the shortest possible path, which is *down* rung A, along the side, *up* rung B and along the other side back to rung A (see Fig. 3). The remainder of the ladder is unaffected—there is no reversal at the right-hand end and therefore no voltage is developed in the output winding.

But if the hold winding on rung B (Fig. 4) is energised by a control or hold circuit in a direction that opposes the flux change on that rung, the flux reversal must seek the next shortest path. Rung C has already been saturated in the 'reverse' direction by the initial reset pulse, so the flux reversal must occur in rung D. If this, too, has been energised by its hold winding that path is also closed. Rung E, like rung C, is already saturated in the direction of the change, so that path too is impassable.

Now assume that the hold winding on rung F is not energised, so leaving an open path for the flux reversal along the opposite sides of the ladder and back to rung A. Obviously, rung K and the output winding on it are unaffected and there can be no output signal.





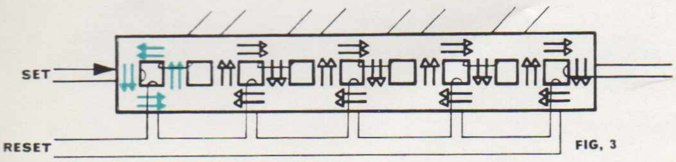


FIG. 3

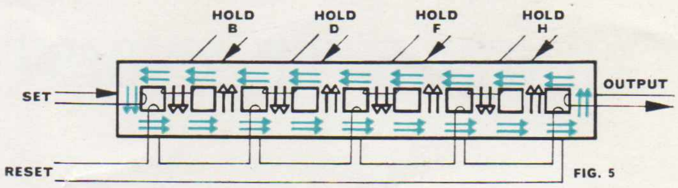


FIG. 5

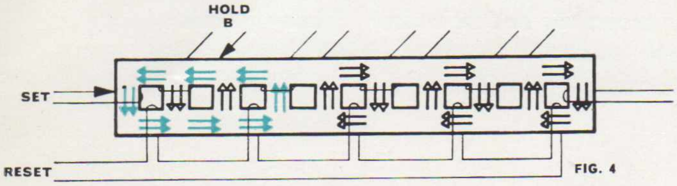


FIG. 4

Thus only if *all* the hold windings are energised will a current pulse applied to the drive winding produce an output signal (Fig. 5), and the laddic is therefore able to perform the logic function of AND. In practice the set and reset pulses are applied alternately so that the core is saturated, ready for operation, after y drive pulse.

Other logical functions can of course be produced. For example, a hold rung may be energised by any of several windings placed on it, in which case the laddic can be made to operate as an OR gate.

The set and reset pulses may be repeated at whatever frequency the system requires—they are equivalent to the clock pulses of a computer. The laddic core itself can be switched in about three microseconds, but the maximum operating speed of a laddic system is at present limited by the associated circuitry to about 10 kc/s. There is no lower limit to the operating frequency and it is practicable, and economical, to use the supply mains frequency to control the clock.

Designs for pulse generators covering most frequencies likely to be needed are being developed in the Mullard Applications Laboratories. Provided they are maintained long enough to switch the core, the hold and set pulses need not be accurately defined. Laddic's toleration in this respect would be useful in industrial control systems, enabling the hold windings to be energised by transducers at some distance from the laddic.

At its present stage of development it seems certain that laddic's first appeal will be for systems that must be, above all else, reliable. An example is the safety system of a nuclear reactor. Here, the position of the neutron-absorbing control rods would govern the hold circuits. As soon as a rod occupied a wrong position the hold circuit would be broken and the laddic would decline to pass the set pulses to the output. The absence of a signal which could, of course, be caused by a failure in any part of the safety circuit, would indicate trouble and be made to initiate the necessary alarm action.

In railway signalling systems the hold circuits could be operated by the signals, by the points, or by electrical impulses representing traffic conditions. Here again, a no-signal condition would immediately cause the system to fail-safe.

Those are obvious situations; there are many others where laddic systems could increase safety or reduce operating costs. The chemical industry has many dangerous processes which must be precisely controlled such as the manufacture of nitroglycerine, in which the reaction, if too fast, produces excessive heat and becomes uncontrollable unless the temperature is lowered immediately the danger point is reached.

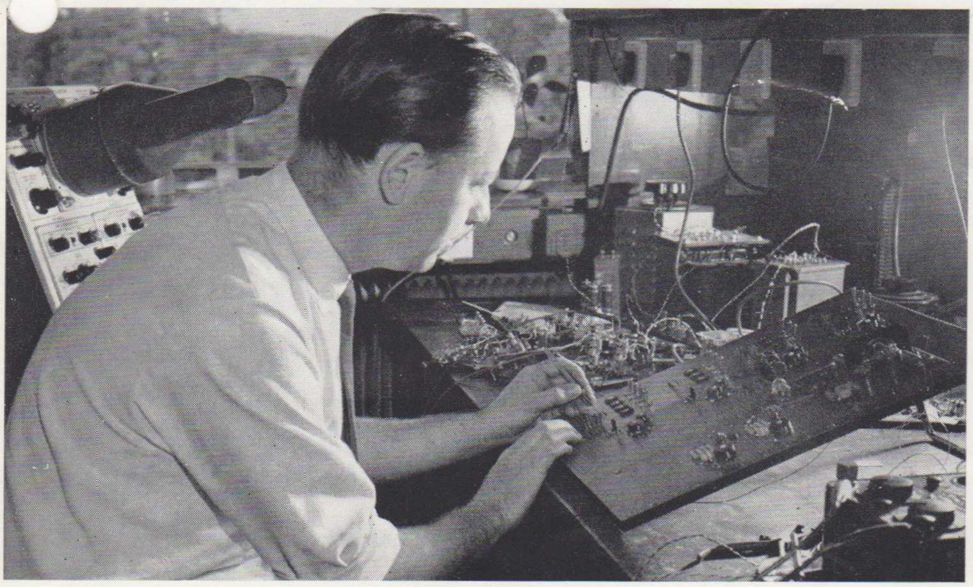
In all but the most rudimentary systems a number of laddics would have to be connected in cascade, and methods of coupling them using various transistor and thyristor circuits are being devised. Magnetic coupling elements, although they introduce a complication, are a possible alternative and one now under development is expected to give satisfactory results.

This article has described only the bare bones of laddic. There are, as already stated, many elaborations by which the device could be called-on to provide almost any mode of logic control. By variations in the winding configuration, by a multiplicity of windings, and by different methods of interconnecting a number of laddics, an extremely wide field of operation can be covered.

In critical control situations—for instance, in nuclear engineering—it is usual to use three transducers to monitor the same parameter, and the decision to 'go' or 'no-go' is a majority verdict. Thus, the failure of one transducer need not mean the unnecessary, and expensive, shut-down of a large plant. Provided two of the transducers show that conditions are normal the laddic elements, to which all three are connected, can be arranged to pass the set pulses to the output.

Reliable magnetic logic systems could be designed for air traffic control and for aircraft blind-landing aids. Where high operating speeds are not essential laddic can also take its place in industry, and many concerned with its future development believe that it will eventually be used extensively in railway signalling.

The Mullard contribution to this highly reliable approach to electronic control is to help plot the future for a control device that, in its sphere, may yet rank in usefulness with the transistor.



Far left:
Laddic Ferroxcube logic elements in an experimental model of a safety control system for nuclear reactors.

Centre:
An experimental 50 c/s industrial logic system using a laddic. The output from the laddic triggers silicon controlled rectifiers which in their turn switch large amounts of 50c/s power.

Left:
Examples of laddic Ferroxcube elements.

SIMPLE TRANSISTOR MEASUREMENTS

This article is the second in a series being published in Outlook and containing suggestions and instructions on a number of experiments in which the properties and behaviour of alloy junction transistors are examined. They are intended to give the student a better understanding of the basic operation of the transistor and to illustrate its practical application.

MODES OF OPERATION

There are three basic methods of connecting the transistor in any circuit—in grounded or common base, in common emitter or in common collector configuration. These are indicated conventionally in Figs. 5, 6 and 7.

Common Base

Because junction transistors have a slight current loss between emitter and collector (represented by the base current) the whole of the power gain in this arrangement is obtained as a result of the low input impedance and high output impedance.

A transistor in common base connection is shown in Fig. 5. If I_e is the emitter current, I_c the collector current and I_b the base current then

$$I_b = I_e - I_c$$

For small signal inputs the current gain (α) may be taken as

$$\frac{I_c}{I_e}$$

$$\therefore I_b = (1-\alpha)I_e$$

Taking into account the leakage current (I_{co}) which flows from collector to base

$$I_b = (1-\alpha)I_e - I_{co}$$

and

$$I_c = \alpha I_e + I_{co}$$

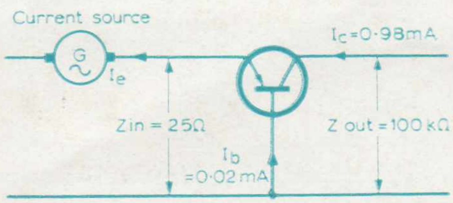


Fig. 5

Thus the collector current depends upon the leakage current and the current gain. If the base is disconnected $I_c = I_e$ and

$$I_c = \frac{1}{1-\alpha} I_{co} = I'_{co}$$

An input change of 1mA through 25Ω in the emitter circuit produces a change of 0·98mA through 100,000Ω in the collector circuit. The current gain (α) is thus 0·98 and the **theoretical** power gain is calculated as follows:

$$\text{*Total Power in Collector Circuit} = I^2R = \left(\frac{0\cdot98}{1000}\right)^2 \times 100{,}000\Omega = 96\text{mW}$$

$$\text{Input Power} = I^2R = \left(\frac{1}{1000}\right)^2 \times 25\Omega = 25\text{mW}$$

$$\text{Power Gain} = \frac{\text{Power in Collector Circuit}}{\text{Input Power}} = \frac{96 \times 10^{-3}}{25 \times 10^{-6}} = 3{,}840$$

*This power will be divided between the collector diode and the load, and the useful output power will depend upon load matching.

Common Emitter

In the common emitter arrangement it is the base current and not the emitter current which is the controlling element. If the base current is increased or reduced by an external circuit, the output current is also increased or reduced.

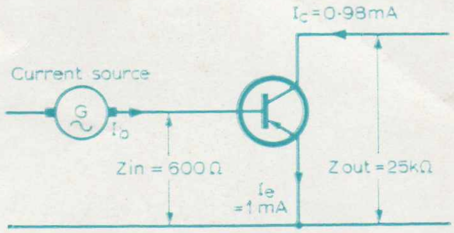


Fig. 6

A transistor in common emitter connection is shown in Fig. 6. For small signals the current gain denoted by α' is taken as

$$\alpha' = \frac{I_c}{I_b}$$

It can also be shown that

$$I_c = \alpha' I_b + I'_{co} \text{ and } \alpha' = \frac{\alpha}{1-\alpha}$$

The characteristic of the common emitter circuit is high current gain, medium input impedance and medium output impedance. A higher leakage current is also apparent.

In this example, using the same transistor as employed in Fig. 5, a change of 0·02mA in base current produces a change of 0·98mA in collector current, so that the current gain is 49.

In this case the power gain depends upon two components. There is in the first place the current gain and there is also a power gain due to the ratio between the output and input impedances. This power gain is of the order of:

$$\frac{25{,}000}{600} = 41\cdot7$$

Common Collector

The arrangement shown in Fig. 7 is similar to the common emitter circuit but now the input is effectively between base and collector and the output is effectively taken between emitter and collector. The general characteristics are similar to those of a cathode follower. With a load in the

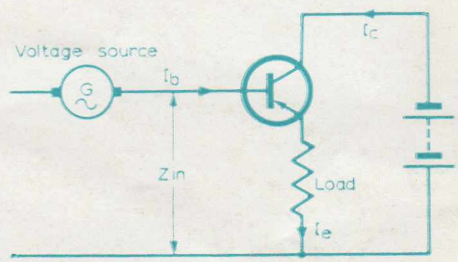


Fig. 7

emitter circuit, the input impedance is approximately equal to the product of the load resistance and the current gain α' which, as previously indicated, is in this case 49. The voltage gain of the circuit is less than unity.

(*Continued in Next Issue*)

PRINTED IN AUSTRALIA BY STAR PRINTERY PTY. LTD.

MULLARD OUTLOOK

TABLE OF CONTENTS—VOL. 7, 1964

When locating products and their type numbers, readers should also refer to the Product Type Number Index shown overleaf.

Continued

MULLARD PRODUCT TYPE NUMBER INDEX

The tabulation below, in alphabetical and numerical order, is of Mullard product type numbers which are frequently referred to and which have appeared in Outlook, Volume 7.