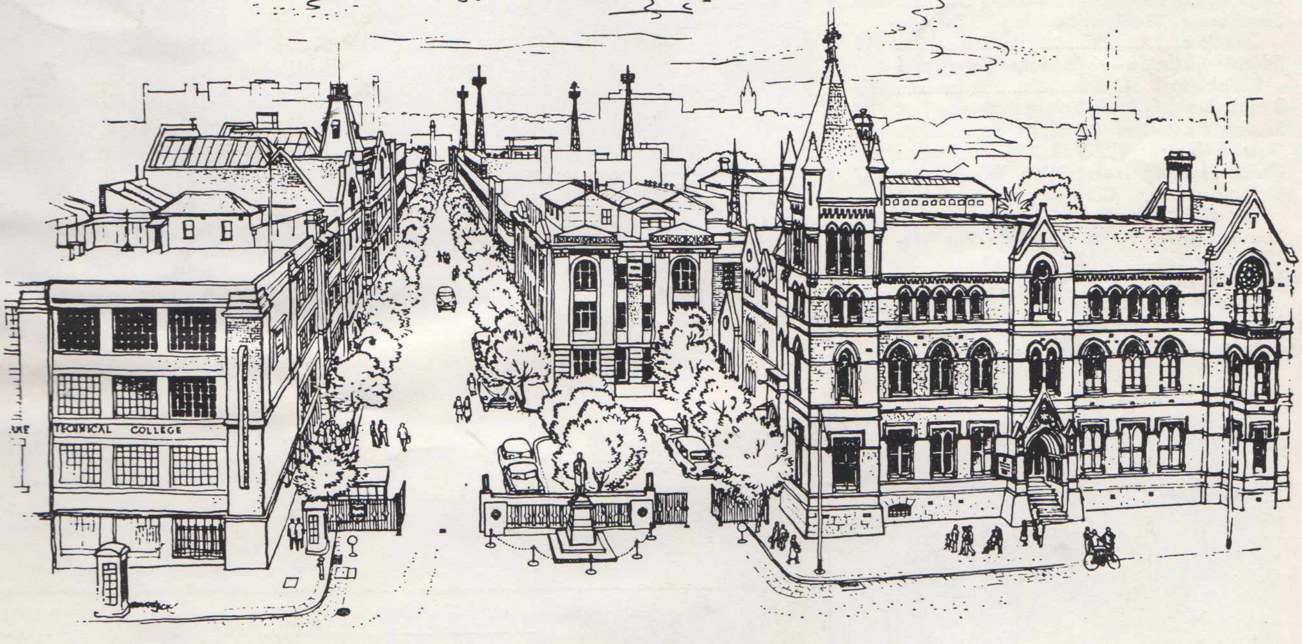
Mullard Outlook

AUSTRALIAN EDITION



1963 I.R.E. CONVENTION ISSUE

VOL. 6, No. 2,
MARCH-APRIL, 1963

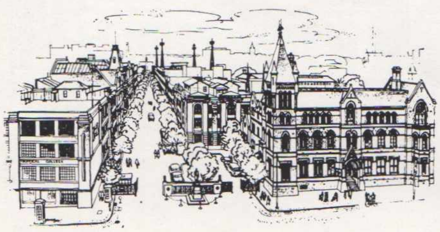
MULLARD-AUSTRALIA PTY. LTD.

VOL. 6 — No. 2 MARCH-APRIL, 1963

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

Editor:
JOERN BORK

TABLE OF CONTENTS



1963 I.R.E. CONVENTION ISSUE

The Royal Melbourne Institute of Technology, venue of this year's I.R.E. Convention.

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

ENGINEERS GALORE AND THE MARQUIS OF QUEENSBERRY

Engineering emphasis is the theme of this Outlook issue with the Institution of Radio Engineers, Aust. Convention being held within the precincts of the Royal Melbourne Institute of Technology, May 20th-May 24th, a bi-annual event for an exchange of views, ideas and theories, with all the objects and ideals of a learned institution and rapidly developing profession.

We accept that a gathering of engineers in an enquiring and informative mood, perhaps uninhibited, is an ideal atmosphere in which to show new developments and products, that each may stimulate the other. We therefore believe it is our duty to support the Convention with our engineering display and look forward to seeing our many friends in Melbourne at that time; particularly in no lesser place than the Royal Melbourne Institute of Technology—"Perita Manus Mens Exculta" —a skilled hand and a cultivated mind.

Apart from its fine academic background and emphasis on technology, we were interested to learn that the Convention Exhibition Centre in the Storey Hall, previously the Hibernian Hall, was where Dame Nellie Melba made her debut as a concert singer and in the same Hall, in 1888, none other than the Marquis of Queensberry, author of the well-known Queensberry Rules for Boxing, himself refereed a fight; the Noble Marquis arrived at the Hall late, found it overflowing and doors locked, was helped by a bystander through an open window in Bowen Lane, the bystander being rewarded with a sovereign. When inside, His Lordship found his gold watch and chain were missing! Within the institute area, we understand that Edward (Ned) Kelly was tried, hanged and buried.

Enterprising days in an enterprising city and with Melbourne as host city, we are confident the I.R.E. 1963 Convention will be an abounding success.

Up there Casaly!*

M.A.B.

* For Melbourne readers.

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
144 Keira St., Wollongong. Phone: B 6020
21 Bayliss Street, Wagga. Phone: 5 2265

Victoria
Howard Electrical & Radio Company.
Vere Street, Richmond. Phone: JB 4716

South Australia
Agents:
Woollard & Crabbe Limited
180 Wright Street West, Adelaide. Phone: LA 4713

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

Queensland
C. A. Pearce & Co. Pty. Ltd.
33 Bowen Street, Brisbane. Phone: 2 3201
13 Palmer St., Townsville. Phone: 5864

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 Applications Engineering Laboratory and the Technical Service Department

The Beginning

In the United Kingdom 35 years ago our Parent Company, in developing and introducing new valve types, was confronted with the problem of how to ensure these would be applied to the fullest advantage, for the corollary of technical advantage was commercial advantage, and valve development progress was rapidly outstripping end product development. A complete technical service, in some cases from drawing-board to end product was required, a down-to-earth, commonsense approach which in turn has influenced valve design parameters and so the Mullard Applications Engineering Service began and has since been emulated in many industries in many countries. But behind it the applications engineers were gathering information and experience and becoming specialists in a new engineering category.

In latter days a practical example of this approach has been our highly successful range of AM and TV receiving valves—a complete and positive range engineered for the specific purpose, rather than a hotchpotch carried over and garnered from previous AM, FM and telecommunications types.

The extent and scope of this Mullard service today covers valves, semiconductors, special electron tubes, ferrite materials, composite components, in fact all of the goods we offer and the more complex these become, the greater the applications effort and the greater the effort in compiling and issuing complete and accurate technical data.

The Applications Laboratory

Mr. H. S. Watson is leader of our applications team and responsible for the Laboratory located at our Head Office, 35-43 Clarence Street, Sydney. In addition to the main laboratory area, facilities include a technical library, model shop and customer conference room. It provides a precise, confidential and sophisticated service, geared to the tempo of electronic development. The logical choice of valve and semiconductor types and the trusted engineering liaison with the equipment manufacturers ensures their correct use with the ultimate in long life and reliability.

Work Study and Consulting Service

The Laboratory, in addition to its prime function of applications engineering, also undertakes specific consulting commissions, industry systems suggestions, feasibility studies, suggested production techniques and estimated cost analysis, in all, electronics in the widest sense.

Main section of Mullard Applications Laboratory, Sydney

The Technical Service Department

Readers may wonder from where emanates the never ending stream of Mullard publications and why, bearing in mind there are, we understand, 286 electronics periodicals in English if one was to subscribe to them all! Bearing in mind also that few engineers and technicians are serviced with all of our Mullard handbooks, brochures and reports, in view of the wide field covered.

In this field again, our Parent Company, Mullard Limited, London, by design and necessity have set the standard and the pace and many of the engineers of today have tucked away some of our earlier journals, well-thumbed and with sentiment reminiscent of student and apprenticeship days.

Mr. B. P. A. Beresford administers this department, which also caters for the day to day technical enquiries that do not require a specific applications project. As can be expected, it includes a central valve and semiconductor type and information library, in many languages—perhaps the most comprehensive specialised valve, electron tube and semiconductor reference library in Australia.

Our Specialised Business

These items are the very heart of almost every electronic device and as this is our business—our specialised business—we are pleased to provide the dual facilities of applications engineering and technical data service.

A client application engineering conference

Mullard

To assist you at the Convention — The Mullard-Australia Team!

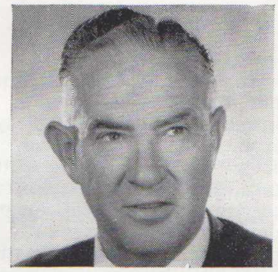
SALES ENGINEERING


L. A. WADE
Sales Manager


R. H. SMITH
Sales Engineer


W. R. PENBERTHY
Sales Engineer


W. J. WELLER
Sales Engineer


G. A. AUNGLE
Sales Engineer

TECHNICAL SERVICE DEPARTMENT


B. P. A. BERESFORD
Manager, Technical Services


J. BORK
Editor

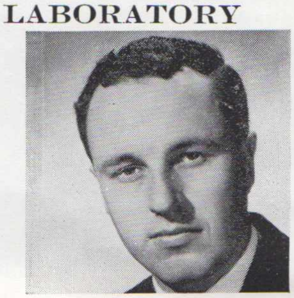

J. T. LAKE
Publications and Distribution Officer


H. S. KING
Manager, Valve & Tube Service Centres

APPLICATIONS LABORATORY

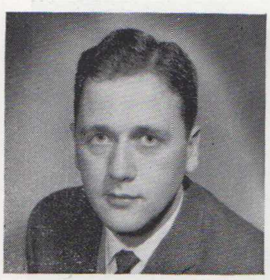

H. S. WATSON
Chief Engineer

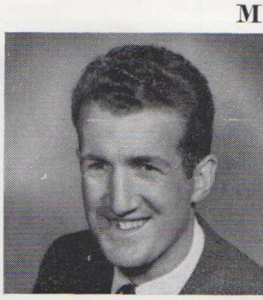

R. L. WEBB
Senior Applications Engineer

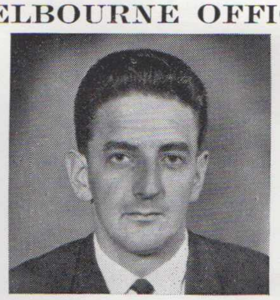

J. G. ALEXANDER
Applications Engineer

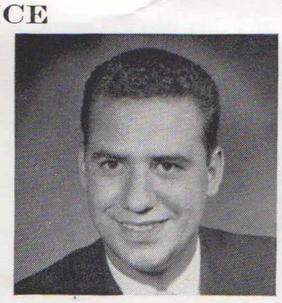

R. T. SMITH
Applications Engineer

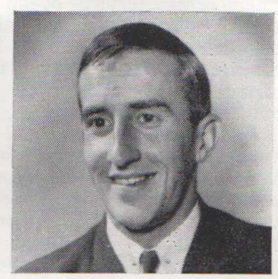
MELBOURNE OFFICE


J. E. PEARCE
Manager


G. W. GALE
Technical Commercial Manager


L. RONALDS
Sales Engineer


C. GRECH
Entertainment Valve Sales


A. FELTON
Entertainment Valve Sales

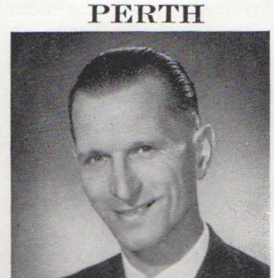
ADELAIDE


B. BRICE
Resident Representative

BRISBANE


R. E. JEFFERY
Resident Representative

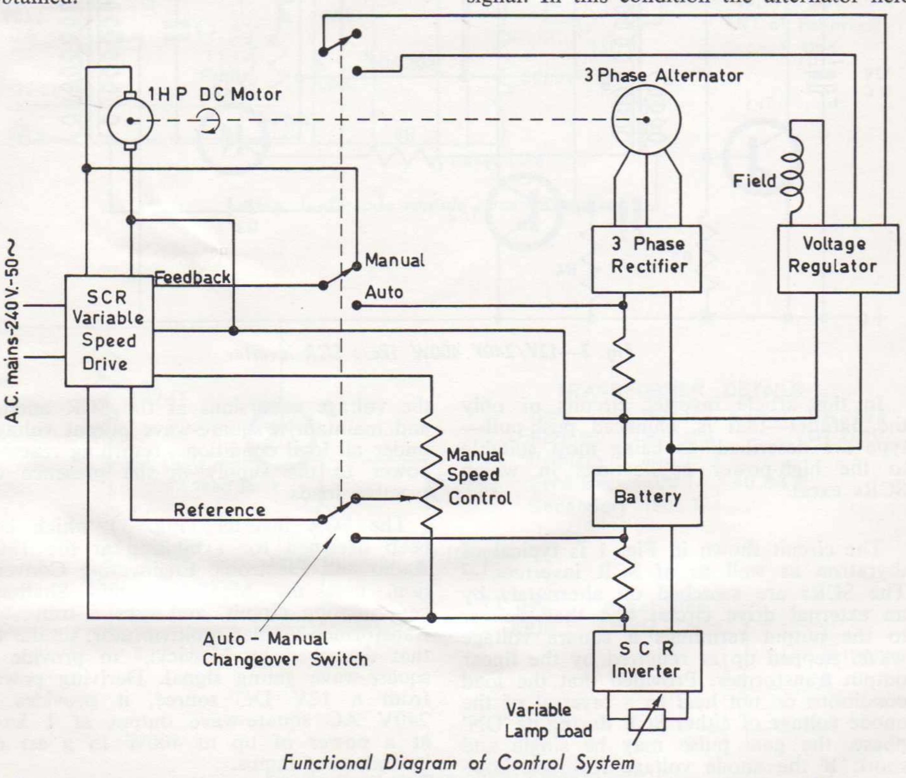
PERTH


L. E. NORRIS
Resident Representative

SOLID STATE POWER CONVERSION AND CONTROL

Focal point of the Mullard display at the I.R.E. Convention Exhibition is a comprehensive control system which illustrates a few of the many applications of semiconductor devices (transistors, silicon controlled rectifiers, diodes) in equipment designed to control quantities or processes.

The final quantity in this case is the state of charge of an accumulator subject to varying load. This load is the supply to a high-powered DC to AC inverter using silicon controlled rectifiers to light a series of lamps. The number of lamps connected at any one time may be varied by means of switches, so that a full range of conditions from no load to full load may be obtained.



Functional Diagram of Control System

In order to replace the power consumed by the load connected to it the battery is charged by a three-phase alternator feeding into a diode bridge rectifier circuit. The alternator in turn is driven by a one-horsepower DC motor supplied from a variable speed drive using silicon controlled rectifiers. The v.s. drive draws its current directly from the single phase mains supply.

Two modes of operation are possible, manual and automatic, changeover being effected by a switch. In the automatic condition, the motor speed is controlled by inverter supply current so as to maintain an input current to the battery equal to the load current on it.

The DC motor drives the alternator through a 3:1 step-up ratio, using pulleys and V-belts. The alternator itself is an automotive unit and is normally driven at 1·5 times engine speed. Thus the display current is switched to a constant (maximum) value, so that all control of charge installation simulates an engine speed range of 400 r.p.m., a typical idling speed, to 3,000 r.p.m., which represents 50-60 miles per hour in a vehicle. This enables a good indication to be made of the superior characteristics of an alternator, as the battery charge current is monitored and can be seen to be quite substantial even at low engine speeds. The actual charge current is set by a voltage regulator which maintains the battery voltage at a constant value by controlling the alternator field (rotor) current and hence its output.

When automatic control is selected, the DC motor speed is controlled by the v.s. drive in response to battery load current signal. In this condition the alternator field current is effected by automatic variation of motor speed. As the inverter load is changed by connecting more or less lamps to its output, the DC motor speed rises or falls to compensate.

Negative feedback is applied to the v.s. drive to stabilise its characteristics in two ways, depending on the type of control selected. In manual, a proportion of the armature voltage, which depends on the speed, is fed back to maintain the constant speed set by the hand control against changes in load. In automatic, the battery charge current signal is fed back to modify the motor speed until charge current and load current are equal.

With its facilities, therefore, the system represents a typical closed-loop feedback control system where the speed of a motor is varied automatically to provide close control of a manufacturing process.

J. G. Alexander
Applications Laboratory, Sydney

MULLARD MICROCIRCUITS

THE MULLARD TECHNIQUE OF MICROMINIATURISATION

Microminiaturisation means many different things to different people. In our view, apart from the obvious reduction in size, it implies the search for increased overall reliability, together with a reduction in cost. Both these characteristics become more necessary as electronic equipment becomes more complex.

There are three basic techniques for microminiaturisation, the solid state, the micromodule and the microcircuit. The solid circuit uses a piece of semiconductor material which is etched and doped to give the required electronic performance. The micromodule employs one wafer for each of the resistors of the circuit, others for the capacitors and so on. Interconnections are made by riser wires soldered to the edges of the wafers. This forms a stack which can then be encapsulated. The microcircuit uses film resistors and capacitors deposited onto the wafer together with their interconnections. Active components such as transistors and diodes are set into the wafer and there is, therefore, a considerable reduction in the number of joints required.

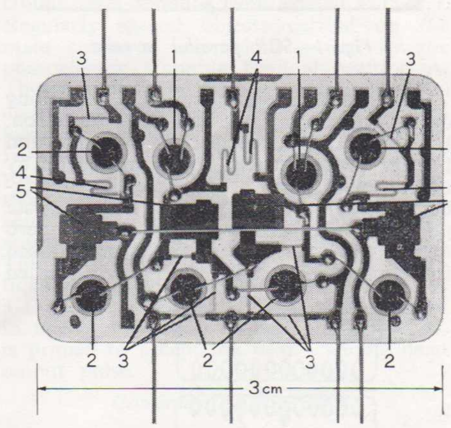


Fig. 1—Microcircuit shift register; (1) transistors; (2) diodes; (3) and (4) resistors; (5) capacitors

Mullard have selected the microcircuit approach because it offers the optimum compromise between the inflexibility of the solid state on the one hand and, on the other, the inherent unreliability of techniques using the interconnection of single components.

Three microcircuits are now available, the MC1101, a 'NAND' gate designed to perform the logical function 'AND NOT' for negative '1' inputs, the MC1102 a bistable circuit designed to act as a steered counter or shift register stage and the MC1103, a driver circuit, intended to drive counters or shift registers which present a capacitive load. Although only digital circuits have been developed, work on linear circuits is, in fact, now taking place.

HIGH POWER INVERTERS WITH SCRs

The SCR inverter provides AC power at high voltage from a DC source. The high voltage and high current ratings attainable in the SCR (e.g. 500A maximum recurrent peak current, 75-500V reverse voltage for BTY64 series) place the SCR inverter in a class beyond vibrator and transistor inverters, in which its main competitors are the genemotor and the thyratron inverter. While still quite expensive in comparison with these, this type of inverter presents the advantages of small size, long life, reliability, and increased efficiency at lower supply voltages. Typical applications are the inversion of a ship's DC supply for the operation of AC equipment, and the provision of fluorescent lighting in railway carriages and buses.

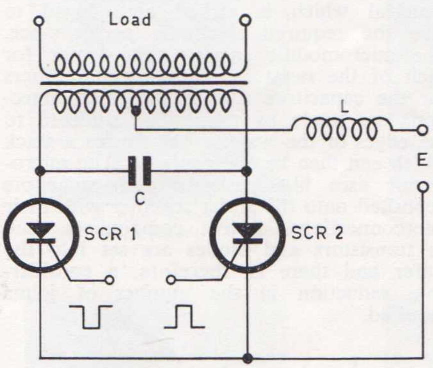


Fig. 1—SCR parallel inverter

The SCR is essentially a switching device, paralleling the gas-filled thyratron, but with lower conduction voltage and heavier current rating. It is switched on by a low-power pulse at the gate (input) circuit, and is switched off by a negative gate pulse accompanied by negative or zero anode voltage which must be maintained for some definite period (see article entitled

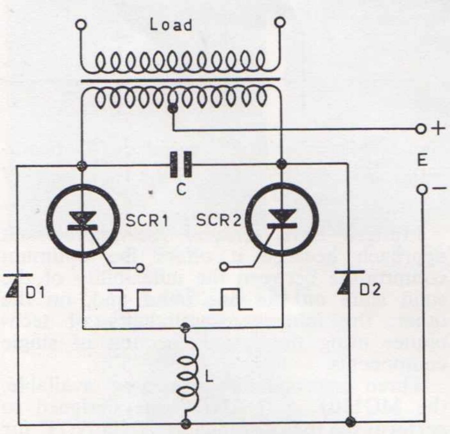


Fig. 2—SCR parallel inverter with McMurray and Shattuck commutating circuit

'Firing Requirements, Firing Circuits and Motor Control Applications' in this issue).

The necessity of applying 'OFF' signals to anode and gate simultaneously, and the absence of any linear mode of operation, complicate the design of SCR inverter circuits. Because of phasing difficulties, it is not generally practical to design them for self-excitation. Instead, an external oscillator (using transistors, for example) is employed to apply gating pulses to the SCRs, while the SCRs supply negative anode pulses to one another.

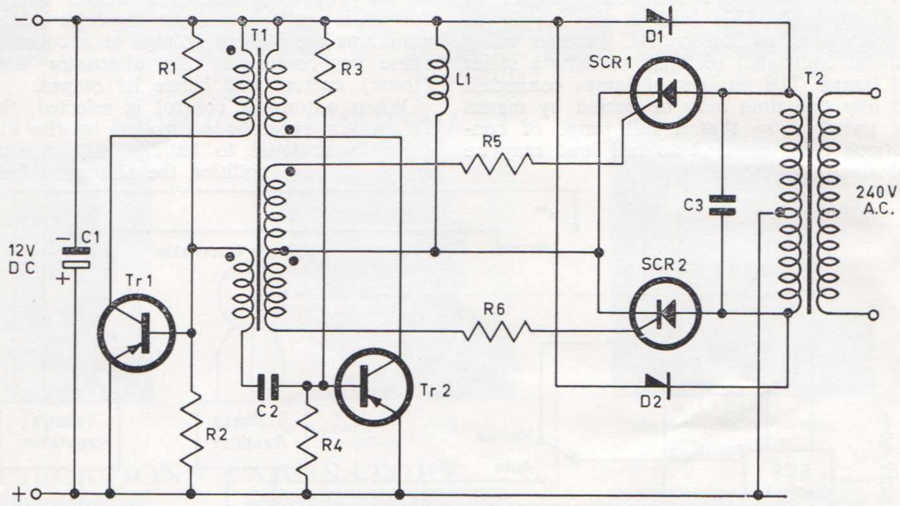


Fig. 3—12V/240V 400W 1kc/s SCR inverter

In this article inverter circuits of only the parallel—that is, balanced push-pull—type are described, as being most suitable to the high-power applications in which SCRs excel.

The circuit shown in Fig. 1 is typical of thyratron as well as of SCR inverters.[1, 2] The SCRs are switched on alternately by an external drive circuit and they deliver to the output terminals a square voltage wave, stepped up as required by the linear output transformer. Provided that the load conditions do not lead to a reversal of the anode voltage of either SCR during its 'ON' phase, the gate pulse may be single and short. If the anode voltage reverses, however, as may happen with an inductive load, a repeat gate pulse, or—preferably—a square-wave gating signal, must be applied. The commutating capacitor C is used to communicate a negative anode pulse to SCR 1 at the instant when an 'ON' gate pulse is applied to SCR 2 and an 'OFF' pulse to SCR 1. The inductor L in series with the DC supply is necessary to limit the discharge current flowing in C at the moment of commutation. (Otherwise, assuming a perfect transformer and zero-impedance supply, C would have to reverse its charge instantly.) Upon switching, the anode voltage of SCR 1 is driven from nearly zero voltage to $-2E$ volts, and recovers to $+2E$ volts in consequence of a damped oscillation executed by the series circuit C-load-L-supply SCR 2.

An improved commutating circuit (Fig. 2) has been devised by McMurray and Shattuck.[3] Auxiliary diodes are used to limit the voltage excursions at the SCR anodes and maintain a square-wave output voltage under all load conditions, returning reactive power to the supply in the presence of reactive loads.

The SCR inverter (Fig. 3) which has been designed for exhibition at the 1963 Radio and Electronic Engineering Convention, uses the McMurray and Shattuck commutating circuit, and uses a transistor transformer-coupled multivibrator, similar to that described by Nowicki,[2] to provide a square-wave gating signal. Deriving power from a 12V DC source, it provides a 240V AC square-wave output at 1 kc/s at a power of up to 400W to a set of incandescent lamps.

R. L. Webb
Applications Laboratory, Sydney

References

1. Mullard Overseas Limited. *'Thyratrons for Industrial Control.'* Manual TP318B. London, W.C.1, England, 1957.
2. J. R. Nowicki. 'SCR Inverter for Fluorescent Lamps.' *Mullard Technical Communications*, Vol. 7, No. 61, November, 1962.
3. McMurray, W. and Shattuck, D. P. 'A Silicon Controlled Rectifier Inverter with Improved Commutation.' *AIEE Transactions (Communications and Electronics)*, No. 57, November, 1961.

IF YOU CHANGE YOUR ADDRESS:

1. Notify Mullard-Australia Pty. Ltd. immediately.
2. If possible, let us know in advance. Thirty days' notice will enable our mailing system to operate more efficiently.
3. Notify us in writing and include the address label from one of the recent issues showing your old address and code line.

THANK YOU.

SEQUENCE CONTROL USING DECADE COUNTERS

An ever-widening field of applications is being found for electronic decade counting. This field includes batch counters, business machines, computers, electronic telephone exchanges, sequence control and industrial machine control.

Complete counters can be constructed by interconnecting a number of individual modules each designed to perform a specific task. The number of modules used and the interconnection arrangement can be varied to suit the particular application. The sorting and batch counting control used in the display was constructed by this method. Only four basic modules were necessary to complete the full complex system. The four module functions are:

1. **Decade Module:**
A scale of ten counter, based on the well known Z504S cold cathode stepping tube. The drive circuit consists of a transistor blocking oscillator which provides a paired-pulse drive waveform capable of stepping the tube at a maximum count rate of 3,000 counts/sec. An output corresponding to any count position is provided. The circuit diagram is shown in Fig. 1.

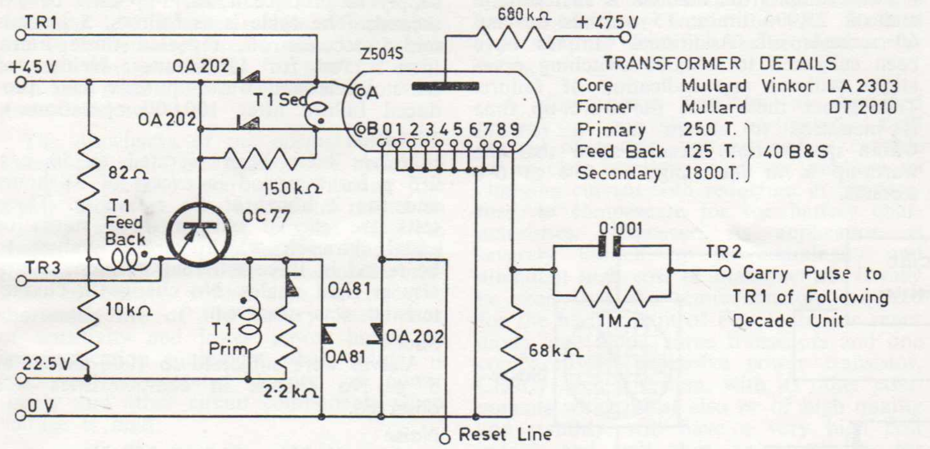


Fig. 1—Decade module circuit diagram

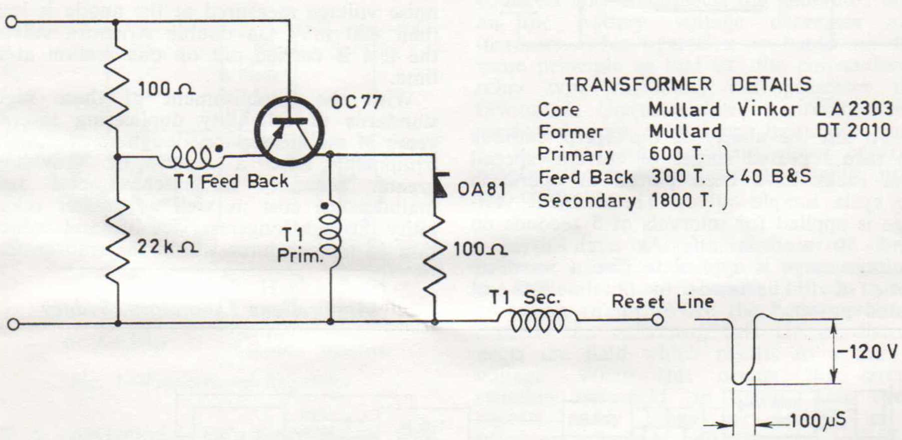


Fig. 2—Reset module circuit diagram

2. **PE Input Module:**
Converts the output of a semiconductor light sensitive device into a pulse suitable for driving a decade module.
3. **Reset Module:**
Provides a negative pulse of sufficient amplitude to reset all decade modules to zero. The circuit diagram is shown in Fig. 2.
4. **Output Module:**
Basically an electro-magnetic relay made more sensitive by the addition of an amplifying stage. It is designed to operate directly from the output of one decade module or from multiple decade modules via a coincidence circuit.

SORTING AND BATCH COUNTING DEMONSTRATION

The display model simulates a main conveyor branching into four separate conveyors. Coloured blocks represent the conveyed objects which are automatically sorted, according to colour, by the electronic control unit. The complete electronic-mechanical display serves to illustrate some of the problems encountered in industrial conveyor systems.

Fig. 3 shows diagrammatically the colour sorting conveyor system and Fig. 4 a block diagram of the electronic control system.

Assume initially that all decade modules are in the zero position. The output of the sequence decade module (abbreviated S.D.M.) is then applied to only the first colour reset module (abbreviated C.R.M.). Regularly spaced objects carried on the main conveyor are colour sensed by the photoelectric scanning head at position A. The output from the scanning head is applied to all colour reset modules and also, after a suitable time delay, to S.D.M. However only the first C.R.M. is initiated which, in turn, resets the transfer decade module (abbreviated T.D.M.) from the zero position to the first position of the appropriate colour (red, yellow, or blue). This satisfies the 'NOT GATE' pass conditions, hence synchronising pulses reach the T.D.M. drive circuit. The S.D.M. now steps to the next position so that No. 2 C.R.M. is primed to accept the next scanning head output pulse.

Continued on page 20.

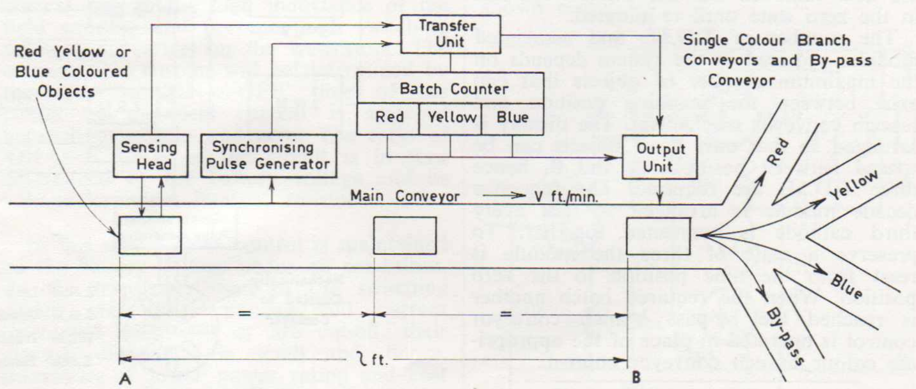


Fig. 3—Colour sorting conveyor system

'HARP' CATHODE—WARM-UP TIME AND RELIABILITY

The quick-heating valve display model on the Mullard stand at the 1963 I.R.E. Convention Exhibition is a working model of one of the many quality control tests given to the 'Harp' cathode used in quick-heating valves.

Fig. 1 illustrates in block form the test set-up. When the time switch closes, a voltage surge derived from the 12 V supply is applied to the heater via R1. When C1 is charged via R2 the relay RL closes changing the heater supply to a filament winding on the transistor DC to DC converter.

The surge voltage and the relay delay time are set to provide a warm-up time of 0·1 seconds. For simplicity in this model RF excitation has not been used but DC voltages are applied to the anode and screen. Grid bias is set to provide normal average anode current. An oscilloscope coupled to the anode circuit displays the warm-up time to rated-power output. Note that the anode current has reached 70% of maximum in 0·1 seconds, a rate of warm-up which is universally acceptable.

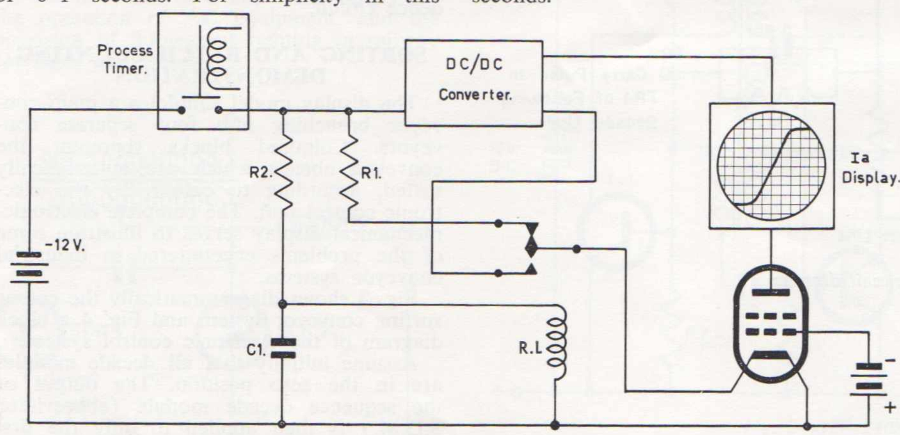


FIG. 1.

Filament

In addition to the normal rigorous control procedure every conceivable life test has been applied to the quick-heating valves to prove the reliability of the 'Harp' cathode.

For example, the filament is switched on and off 25,000 times; 15 seconds on and 60 seconds off. Additional samples have been subjected to 100,000 switching operations without any indication of failure. Throughout these tests the warm-up time is measured to ensure that it remains within specification. Results show that the warm-up is in the range of 0·4 to 0·8 seconds.

To test the ability of the 'Harp' cathode to take repeated surges of current, special test racks have been placed in operation to cycle sample valves. The filament voltage is applied for intervals of 5 seconds on and 30 seconds off. At each cycle a voltage surge is applied to give a warm-up time of 0·1 seconds to obtain 70% of rated power.

In another test rack an even more vigorous test is given in that anode and screen voltages are applied and RF power generated. The filament voltage is surged from a battery, then the normal filament voltage is applied from a transistor power supply to produce a warm-up time of 0·1 seconds. The cycle is as follows; 5 seconds on, 5 seconds off, repeated three times, then a rest for $1\frac{1}{2}$ minutes. Neither of these accelerated warm-up tests have produced failure after 100,000 operations.

Vibration

Valves have been vibrated at 50 c/s, 2·5 g for a period in excess of $5\frac{5}{8}$ hours and for 2 hours at 50 c/s, 5 g. These tests are carried out in three stages of equal duration which permits valves to be tested in three different positions mutually at right angles. No change in characteristics was observed.

Shock

Valves were subjected to 1000 shocks at 5 g. No change in characteristics was observed.

Noise

$$\text{At } V_a = V_g2 = 250 \text{ V}$$
$$I_a = 2 \times 10 \text{ mA}$$
$$R_a = 2 \text{ k}\Omega$$
$$f = 50 \text{ c/s}$$
$$g = 2{\cdot}5$$

noise voltage measured at the anode is less than 800 mV. On double structure valves the test is carried out on one section at a time.

With the establishment of these high standards of reliability duplicating several years of mobile service, designers of mobile equipment have a means of providing greater economy in purchase cost and maintenance cost as well as greater reliability through longer service life and reduction of temperatures within the transmitter.

H. S. Watson
Applications Laboratory, Sydney

Continued from page 19

Number 1 T.D.M. will continue to step, independent of the condition of any other module, until the output position is reached. The resulting output pulse initiates both the appropriate batch counter and branch conveyor control, as well as resetting the T.D.M. to zero. The synchronising pulses are now removed and the T.D.M. remains in the zero state until re-initiated.

The number of T.D.Ms and associated modules required in the system depends on the maximum number of objects that can exist between the scanning position and branch conveyor mechanism. The display is arranged so that only three objects can be spaced between positions A and B, hence three T.D.Ms are required. The sequence decade module is arranged so that every third cathode is connected together. To preserve a scale of three the module is reset from the nine position to the zero position. When the required batch number is reached the by-pass branch conveyor control is operated in place of the appropriate colour branch conveyor control.

R. T. Smith
Applications Laboratory, Sydney

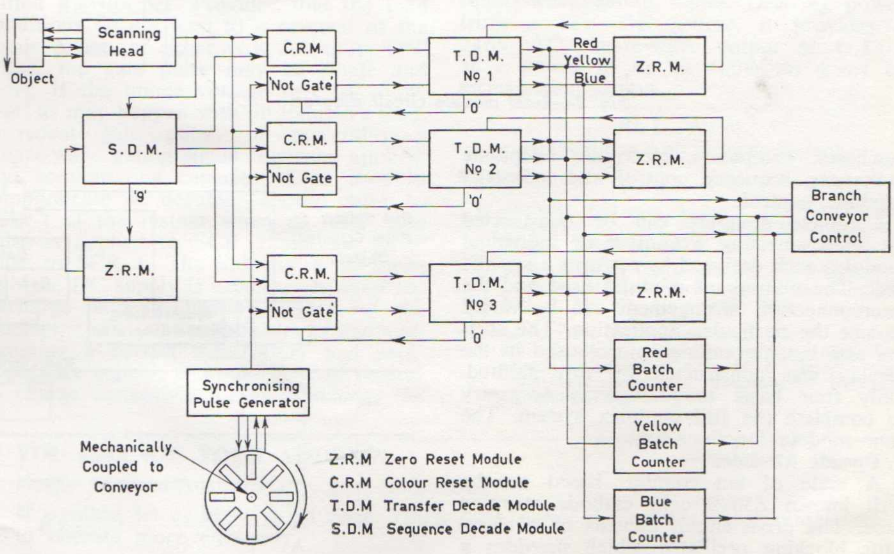


Fig. 4—Colour sorting and batch counting control system

REGULATION OF LOW VOLTAGE GENERATORS

A need has long been felt for the replacement of the traditional type voltage regulator used for the control of battery voltage under charge by a unit which will provide closer regulation of the battery voltage with reduced maintenance requirements. This is particularly necessary for vehicular and marine mobile communication equipment and fixed installations which rely on storage batteries for power supply, such as remote line communication repeater stations.

The drawbacks of the normal regulator system are well known; excessive operation of the regulator contacts switching the highly inductive generator field results in burning and pitting of contact material, leading to frequent adjustment and final failure. The effects of the consequent variations in battery voltage are detrimental; shortening life of the battery itself, lack of sensitivity and lower power output in communication equipment when voltage is low, and accelerated failure of power supply and other circuit components when voltage is high.

A system to overcome the above drawbacks should contain no moving parts and be adequately stabilised against wide variations in temperature such as can occur in the normal locations of the regulator unit.

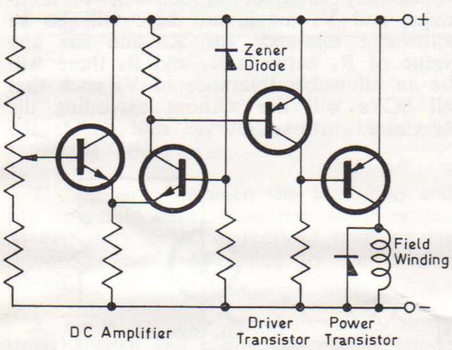


Fig. 1—Proportional Regulator

If a conventional DC generator is used, some form of current limitation must be provided by the regulator to prevent overloading of the machine when it is run at high speed and is providing current for a discharged battery. If an alternator is used, no such current regulation is required as the alternator has a self-regulating action due to the impedance of its windings increasing at high speed and thus high frequency. However, an alternator requires a greater range of field current for its effective control, so imposing more stringent requirements on the regulator unit.

The closest control of battery voltage will be obtained by the use of a proportional type regulator. In such a regulator the generator field current is a continuous function of battery voltage, field current increasing as battery volts decrease. In this type of system, the total loop gain is high, so that precise control of battery voltage is obtained.

Such a regulator would consist of a DC amplifier to provide an output proportional to the difference or error between a reference and a proportion of the battery voltage. This error signal is then amplified and used to control a regulating element consisting of a power transistor connected in series with the field. A basic circuit for this system is shown in Fig. 1. A suitable power transistor would be the Mullard type ADZ11, which has the required power rating.

A proportional regulator as described above can provide a particularly tight control of battery voltage, and is capable of special functions such as increase in charging current with reduction in temperature, to compensate for the battery characteristics. However, its application is severely limited by its complexity and attendant high cost of manufacture. It may be seen that the semiconductors required for the basic circuit of Fig. 1 are one zener diode, one diode, three transistors and one comparatively expensive power transistor. Clearly such a system, with its other components which must also be of high quality and stability, will have a very high cost indeed and will thus be unsuitable for quantity production.

Effective control of battery voltage may be obtained more economically by the use of a transistor switch which alternately connects and disconnects the generator field as the battery voltage decreases and increases. This operation is based on the same principle as that of the conventional relay type regulator, but possesses the favourable characteristics of circuits using semiconductors; switching frequency may be increased with resultant closer control, without the problems associated with moving contacts.

In this type of regulator the error signal between a reference and a proportion of the battery voltage is applied direct to the switching circuit. When the battery voltage exceeds the reference, this circuit disconnects the field which results in a fall in voltage. When this occurs the circuit switches the field on again. This cycle repeats many times per second, at a frequency limited only by the response time of the machine.

A diode reverse-connected across the field maintains a continuous flow of field current due to the high inductance of the field winding and prevents high switching voltages appearing on the transistors. The average field current will be determined by the ratio of 'ON-to-OFF' time of the circuit as the field current is switched between maximum and zero. The ratio of 'ON-OFF' times of the circuit is in turn determined by the battery voltage and its voltage-dependent internal resistance.

In this way a close control is maintained on the battery voltage under all conditions. As the transistors used in the switching circuit are either passing full current (bottomed condition) or are cut-off, their average dissipation is small and, hence, transistors of lower power rating and cost may be used here. A suitable transistor for passing the field current would be the Mullard type OC29.

The circuit is illustrated in its basic form in Fig. 2, from which it is evident that the number of semiconductors required is limited to one zener diode, one transistor and one high-current transistor. This represents a minimum number of components and, therefore, cost will be held to a reasonable limit.

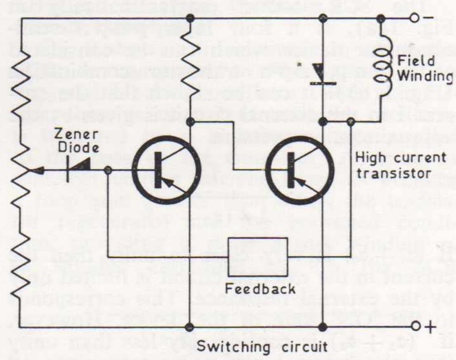


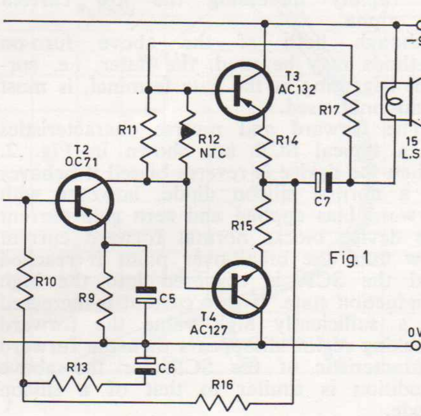
Fig. 2—On-off Regulator

Both the circuits described above are given as examples of the use of semiconductors for controlling battery voltage when the battery is charged from a rotating machine where output can vary widely. These circuits may be used for any similar applications involving the control of low voltages at high current. They both possess the inherent features of circuits using semiconductors, reliability and high response speed. The proportional regulator provides highly accurate control of battery voltage, while the 'ON-OFF' control provides full adequate regulation at an economical price.

J. G. Alexander
Applications Laboratory, Sydney

RED FACES

In the article entitled "A 300mW Complementary Audio Amplifier" (Outlook Vol. 6, No. 1, page 10a) a connection was wrongly shown in Fig. 1 from the junction of R14 and R15 to the collector of the driver transistor T2. A corrected diagram of the relevant portion of the circuit is shown below.



SILICON CONTROLLED RECTIFIERS

Firing Requirements, Firing Circuits and Motor Control Applications

Since its introduction only six years ago the silicon controlled rectifier (SCR) has won acclaim from all sections of the electronic industry. It was first used as a semiconductor equivalent for the thyratron in industrial and commercial applications, however extensive research and ingenious application engineering has led to the development of new products which rely entirely on the unique electrical characteristics of the SCR.

The SCR, shown diagrammatically in Fig. 1(a), is a four layer p-n-p-n semiconductor device which can be considered as a p-n-p, n-p-n transistor combination (Fig. 1(b)). It can be shown that the current I in the external circuit is given by the approximate expression

$$I = \frac{I_{co}}{1 - (\alpha_1 + \alpha_2)}$$

If $(\alpha_1 + \alpha_2)$ is very close to unity then the current in the external circuit is limited only by the external resistance. This corresponds to the 'ON' state of the device. However, if $(\alpha_1 + \alpha_2)$ is considerably less than unity then the external current is only some 10 to 100 times the leakage current I_{co} which for a silicon junction is itself extremely small. This corresponds to the 'OFF' state. The SCR is constructed in such a manner that $(\alpha_1 + \alpha_2)$ is less than unity so that when the device is forward biased it is normally in the 'OFF' state. To trigger the SCR into the conducting or 'ON' state necessitates an

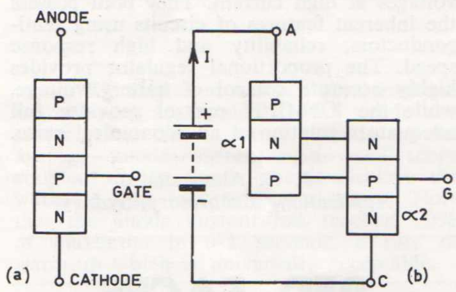


Fig. 1—(a) Diagram of SCR; (b) P-N-P, N-P-N transistor analogy

increase in the low current alphas of the p-n-p and/or n-p-n equivalent sections. This can be accomplished by either,

(a) increasing the forward bias until avalanche break down occurs, or
(b) increasing the emitter current thereby rapidly increasing the low current alpha.

Although both of the above turn-on methods may be used, the latter, i.e. current triggering at the gate terminal, is most commonly used.

The forward and reverse characteristics of a typical SCR are shown in Fig. 2. When the device is reverse biased it behaves as a normal silicon diode, however with forward bias applied and zero gate current the device blocks normal forward current flow until the break-over point is reached and the SCR is triggered into the high conduction state. If gate current is increased to a sufficiently high value the forward blocking region disappears then the forward characteristic of the SCR for the above condition is similar to that of a silicon diode.

Gate Firing Requirements

The design of a gate firing circuit is governed by the gate-cathode characteristic of the SCR. This characteristic, which is similar to a diode in series with a large resistance varies with temperature and with different SCRs of the same type, however it is always within the limits quoted in published data. In Fig. 3, the two heavy lines show these limits for a BTZ20.

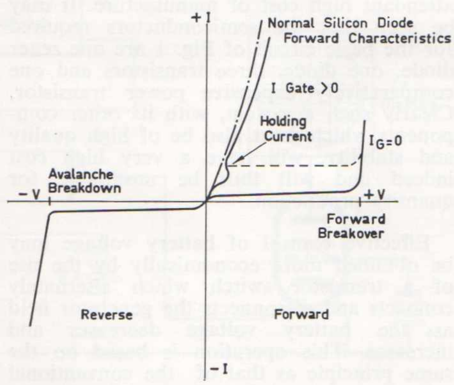


Fig. 2—SCR forward and reverse characteristics

A properly designed firing circuit should fire all SCRs of the type being considered without exceeding any of the maximum ratings of the device. The firing circuit should also make provision for the largest possible tolerance in the values of components used.

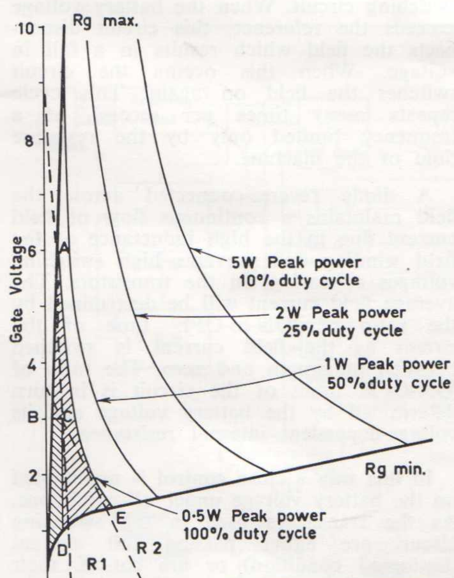


Fig. 3—Firing diagram for BTZ20 controlled rectifier

Gate Firing Methods

DC firing is accomplished by applying a positive DC voltage to the gate-cathode terminals. The upper limit of the gate firing point is set by the maximum average gate power rating and the lower limit by the uncertain firing boundaries. There is now an area in which all values of voltage and current will fire all devices of the type being considered at any temperature above the minimum working temperature. In Fig. 3, this is the area enclosed by points A, B, C, D, E.

The load lines R_1 and R_2 both pass through the extremity of the uncertain firing area (i.e. point C) and touch the limits of the maximum gate power curve. Projection of both R_1 and R_2 onto the voltage axis will give maximum and minimum limiting values of V_s for which no safety factor can be introduced and no tolerance can be applied.

For any value of V_s between V_s minimum and V_s maximum there will be an allowable tolerance for R_s and for any value of R_s between R_1 and R_2 there will be an allowable tolerance of V_s such that all SCRs will fire without exceeding the maximum ratings. Curves that enable the

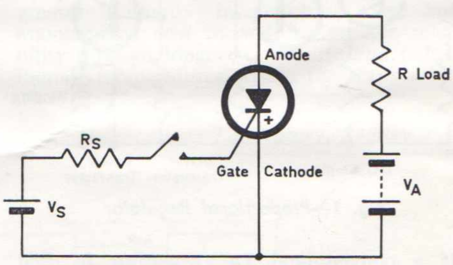


Fig. 4—Basic circuit for DC firing

designer to obtain values of V_s and R_s with tolerances under all conditions are shown in Fig. 5.

AC Firing

AC firing is the application of a sinewave voltage to the gate of an SCR through a suitable series resistance as in Fig. 6.

Compared to DC firing, AC firing has the advantage of a 25% duty cycle. The maximum peak drive power can be increased with a corresponding increase in maximum source voltage.

The reverse gate voltage swing must be suppressed to a value below the maximum reverse gate voltage rating. The diode D_1, in Fig. 6, is used for this purpose.

Continued on page 23

Continued from page 22

Pulse Firing

If the gate signal is the slow variation of a DC voltage or a sinewave, the firing point of the SCR will vary due to changes in junction temperature. Firing accuracy can be improved by using a short firing pulse and injecting a peak power into the gate in excess of the maximum average

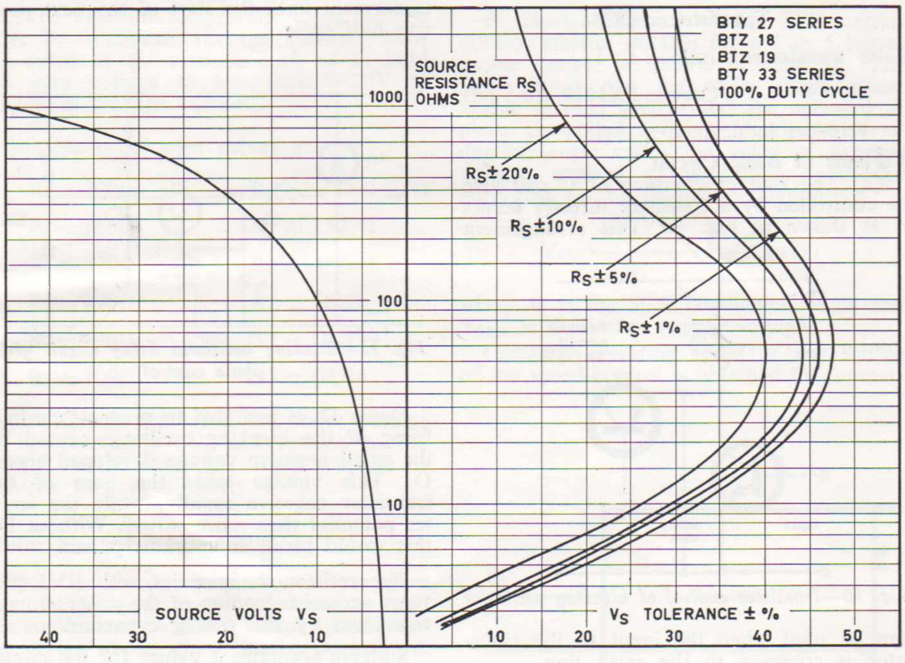


Fig. 5

gate power rating. Pulse firing has the following advantages:—

1. The area between the load line and the uncertain firing area can be increased as the duty cycle is reduced and greater tolerances can be applied to the driving circuit.

2. It is possible to reduce the delay time between gate signal and rise of anode current to a minimum with this method, thus making accurate timing possible.

3. As gate current approaches the firing level there is an increase in leakage current. With pulse firing the extra dissipation due to this is reduced to a minimum.

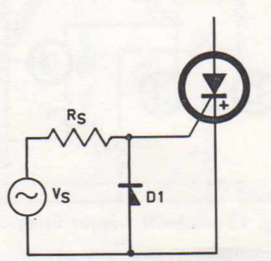


Fig. 6—Basic AC firing circuit

If the firing pulse duration is 6 µsecs or greater, the minimum voltage and current required for firing are the same as for DC, also as the duty cycle is reduced from the figure of 100% for DC it is permissible to increase the peak power delivered to the gate providing the average over a whole cycle does not exceed the maximum gate power rating quoted for the DC case and the peak power is limited to that calculated for 10% duty cycle.

V_s and R_s tolerance curves for a 10% duty cycle are shown in Fig. 7.

Firing Circuits

Junction transistors are very effective devices for firing SCRs since they are compact, stable and inexpensive. Three typical circuits, the triggered blocking oscillator, the free running blocking oscillator and the Schmitt trigger circuit, are described in detail below. Some of the ways in which these firing circuits can be used are also discussed. The circuits are suitable for firing SCRs in the BTY27 and BTY33 series as well as the BTZ18 and BTZ19 series.

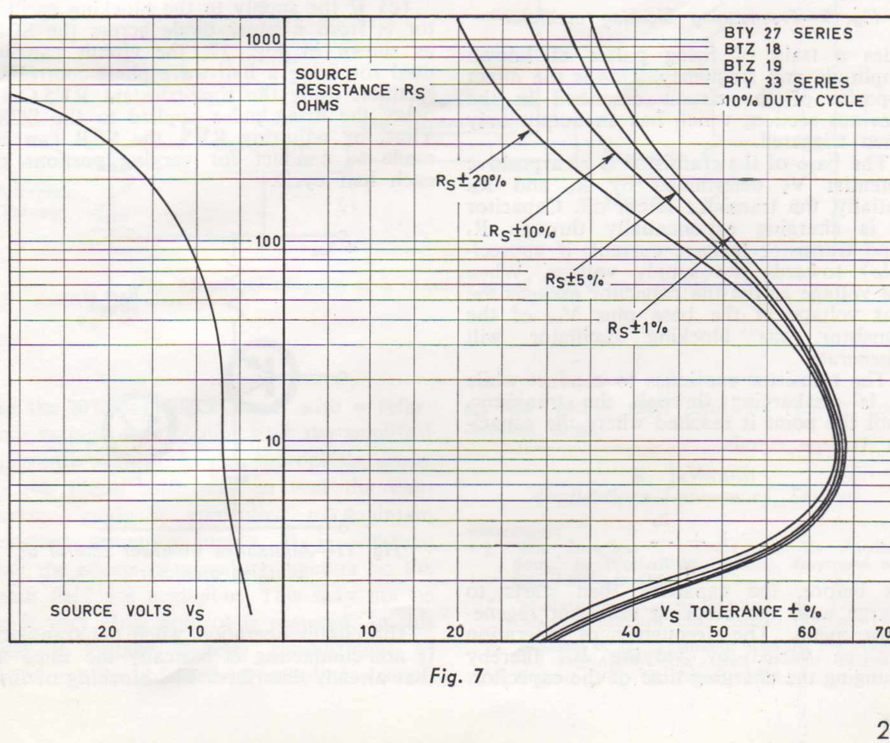


Fig. 7

These firing circuits will, in general, form part of a larger control system. In most cases the circuit designer is already familiar with the basic transistor amplifying and feedback circuits, or has available a suitable reference manual [1, 2, 3] to enable him to design the other parts of the system.

Triggered Blocking Oscillator
Circuit Description

Fig. 8 shows a firing circuit in which the SCR is connected to the output of a triggered blocking oscillator. With this circuit, there is no output until the oscillator is triggered by a negative voltage applied to the base of the transistor. As soon as sufficient emitter current flows to produce a loop gain greater than unity, the transistor regenerates into the bottomed condition, providing a pulse across winding n_1 of suitable amplitude to fire the SCR

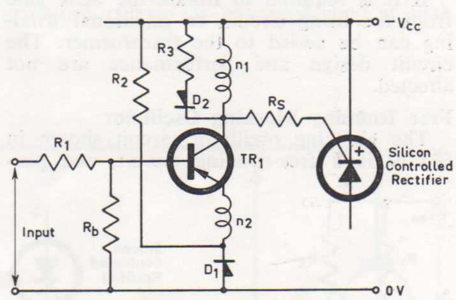


Fig. 8—Triggered blocking oscillator

through R_s. Collector current will now increase through the primary inductance of the transformer until the point αI_e is reached, when the transistor will regenerate

Continued on page 24

Continued from page 23

into the 'OFF' condition. The negative pulse appearing across n_1 when the circuit switches off is suppressed by R3 and D2.

Suitable component values, which will ensure firing of all SCRs of the types covered by this article, are as follows:

$V_{cc} = 9V$ $R_3 = 150\Omega \pm 10\%$

$R_1 = 2{\cdot}2k\Omega \pm 10\%$ $R_b = 330\Omega \pm 10\%$

$R_2 = 820\Omega \pm 10\%$ $R_s = 150\Omega \pm 10\%$

Transformer ratio $= n_1 : n_2 = 2$

Core = FX1638 (two off)

$L_{n1} = 50$ mH

Transistor = OC84

These values give a pulse duration of 200µs and a maximum recovery time of 1·3msecs. The input current to trigger the blocking oscillator is approximately 2mA.

Modification to Basic Circuit

If it is required to isolate the SCR gate from the firing circuit, an additional winding can be added to the transformer. The circuit design and performance are not affected.

Free Running Blocking Oscillator

The blocking oscillator circuit shown in Fig. 9 is a free-running circuit, and pro-

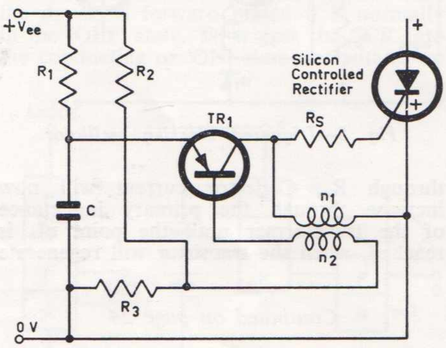


Fig. 9—Free-running blocking oscillator

vides a train of firing pulses of known amplitude and frequency. This is the direct opposite of the circuit described in the previous section, which has an output only when triggered.

The base of the transistor is at a positive potential V_b determined by R_2 and R_3. Initially, the transistor is cut off. Capacitor C is charging exponentially through R_1 (and transistor leakage current, if appreciable) towards the supply voltage. When the voltage across the capacitor exceeds the bias voltage at the base plus V_{be} of the transistor, the blocking oscillator will regenerate.

The transistor continues to conduct while C is discharging through the transistor, until the point is reached where the capacitor voltage equals

$$\frac{V_b}{1 + \frac{n_2}{n_1}}$$

As before, the capacitor then starts to charge until the blocking oscillator regenerates again. The frequency of operation can be varied by varying R_1, thereby changing the charging time of the capacitor.

For firing the types of SCR covered by this article, suitable component values are:

$V_{cc} = 12V \pm 2V$ $R_3 = 390\Omega \pm 10\%$

$R_1 = 3k\Omega$ for pulse duration 460µs, charging time 7·3ms $R_s = 82\Omega \pm 20\%$

$R_2 = 220\Omega \pm 10\%$ $C = 2\,\mu F$

Transistor = OC84

Pulse transformer ratio = 1 : 1

Core = FX1238 (two off)

Winding inductance = 135mH

Variants of Basic Circuit

(a) The rate of charging of C can also be controlled by a transistor directly across it as shown in Fig. 10. This arrangement

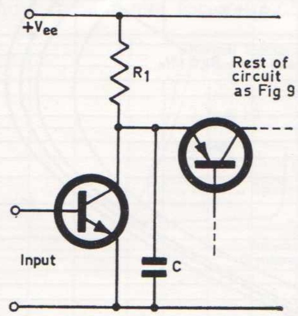


Fig. 10—Transistor control of blocking oscillator

can be used when the input to the transistor is common to the earth line.

Alternatively, the rate of charging of C can be controlled by a transistor in series with R_1, as shown in Fig. 11. This arrangement can be used when the transistor input is common to the firing circuit positive supply.

(b) The basic circuit can conveniently be adapted to feed a number of isolated outputs simply by adding further windings to the transformer.

(c) If the supply to the blocking oscillator is from a zener diode across the SCR as shown in Fig. 12, the circuit can be used for firing a half-wave phase-controlled rectifier, using the time-constant RV5.C to delay the firing pulse applied to the SCR. Thus, by adjusting RV5, the SCR can be made to conduct for varying portions of each half cycle.

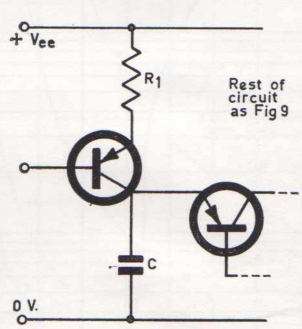


Fig. 11—Alternative transistor control of blocking oscillator

The circuit operation when the transistor is non-conducting is basically the same as that already described. The blocking oscillator will regenerate when the voltage across C just exceeds the bias voltage. When the blocking oscillator triggers, a pulse is fed to the SCR gate and the anode-cathode voltage of the SCR falls to about 1V. This means that the hold-off bias to the base of the transistor has been removed, and the oscillator will continue to oscillate because of the voltage on the capacitor C. Thus C is rapidly discharged and the circuit is dormant until the start of the next positive half cycle.

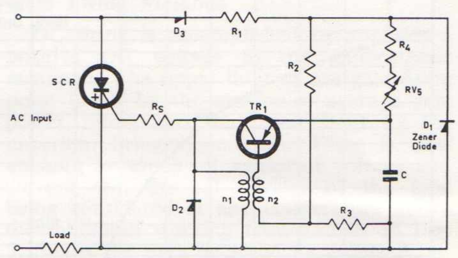


Fig. 12—Blocking oscillator firing circuit giving phase control

Diode D_3 is included to prevent spurious firing of the blocking oscillator caused by the small negative voltage developed across D_1. This voltage takes the base of the transistor negative rapidly, while the emitter potential rises more slowly. Without D_3, this would result in spurious pulses.

The resistor R_4 in series with RV5 prevents excessive loading of the zener voltage waveform by the timing capacitor.

Suitable component values for the circuit of Fig. 12, for 50 c/s phase control of the SCR types covered by this article are:

$R_1 = 1{\cdot}8k\Omega$ Transistor = OC84

$R_2 = 340\ \Omega$ D_1 = OAZ207

$R_3 = 1{\cdot}0k\Omega$ D_2, D_3 = OA200

$R_4 = 82\ \Omega$ Pulse transformer ratio = 1 : 1

$R_s = 68\ \Omega$ Core = FX1238 (two off)

$RV5 = 10k\Omega$ Winding inductance = 36mH

$C = 2\ \mu F$

The arrangements described under headings (a) and (b) can also, of course, be used with this circuit.

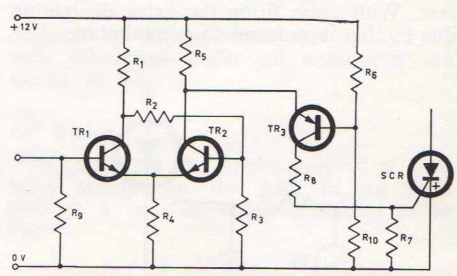


Fig. 13—Schmitt trigger firing circuit

Schmitt Trigger Firing Circuit
Circuit Description

Fig. 13, shows a Schmitt trigger circuit in which Tr1 and Tr2 are followed by an output stage Tr3. With no voltage applied

Continued on page 25

Continued from page 24

to the base of Tr1, this transistor is non-conducting, while Tr2 is hard on. The emitter of Tr3 is negative to its base. The SCR is fired by the collector current of Tr3.

At high temperatures it is necessary, when DC coupling a transistor to the gate of an SCR, to include a resistance from gate to cathode to prevent firing of the SCR by transistor leakage current. Thus the value of R_7 must be such as to hold the gate voltage to less than 0·25V at maximum leakage current.

Suitable component values are:

$R_1 = 100\Omega$	$R_7 = 1k\Omega$
$R_2 = 180\Omega$	$R_8 = 150\ \Omega$
$R_3 = 270\Omega$	$R_9 = 4{\cdot}7k\Omega$
$R_4 = 100\Omega$	$R_{10} = 160\Omega$
$R_5 = 100\Omega$	Tr1, Tr2 = OC139
$R_6 = 100\Omega$	Tr3 = OC84

This circuit fires at a DC input voltage of approximately 6V, and returns to the normal state when the input voltage is reduced to approximately 4·6V.

Motor Control Applications

The increased complexity of modern manufacturing processes produces a need for continuously variable motor speeds, and for drives whose speed can be determined, partially or wholly, by some other process. Silicon controlled rectifiers are being used in increasing numbers to provide such control.

The conversion of an alternating voltage into a variable direct voltage to control the operation of a DC motor is a popular form of speed control. Normally, a separately excited shunt motor is used, and SCRs are employed to provide a variable rectified supply, derived from the AC supply mains,

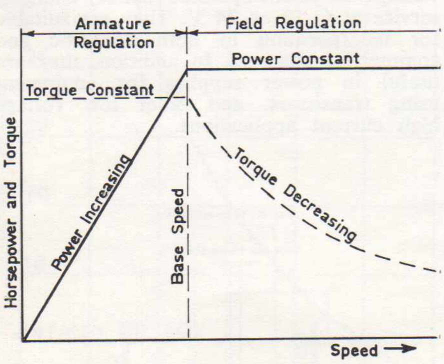


Fig. 14—Torque and power characteristics

to the armature, to the field or both, according to the characteristics required. Fig. 14 shows the torque and power characteristics which can be obtained with either armature or field control. Up to full rated speed or base speed the voltage applied to the field is kept constant, and that to the armature is controlled to give constant armature current and therefore constant torque. The output power increases as speed increases. Above the base speed, the armature voltage is maintained at a constant value, and the voltage applied to the field is controlled to ensure constant power and a torque characteristic which falls hyperbolically with increase of speed.

A block diagram of an SCR motor control system, suitable for a 1 to 5 horse-power range, is shown in Fig. 15. The SCRs form one leg of a single phase bridge rectifier system and are operated as phase controlled switches. Varying the conduction angle varies the value of average voltage applied to the motor armature according to the following relationship.

$$V_a = \frac{\sqrt{2}\, V_{RMS}}{\pi} (1 - \cos\Theta)$$

where Θ is the SCR conduction angle and V_{RMS} is the bridge supply voltage.

Constant operating speed at any setting of the speed control is obtained by comparing the armature back e.m.f. with a reference voltage, then feeding back the amplified difference voltage, in the appropriate sense, to the phase shift unit so that the conduction angle is controlled to maintain constant armature supply. It is assumed that the effects of armature reaction on the main field are negligible. This may not be so if very close control is required. In this case a voltage directly proportional to speed can be obtained by coupling a small tacho-generator to the motor and using its output in place of the armature feedback voltage.

The speed of a DC motor operating with constant excitation is not proportional to the applied armature voltage V, but to $V - I_A R_A$ where $I_A R_A$ is the voltage drop in the armature. Although $I_A R_A$ may be small and even negligible at high speeds, it can be comparable with V at low speeds. Thus to obtain a linear voltage/speed characteristic over a wide speed range, it is necessary to compensate for the $I_A R_A$ drop. This can be achieved by increasing the reference voltage by an amount proportional to the armature drop, thereby increasing the armature supply by an amount equal to $I_A R_A$.

In any control system for DC motors the armature current should be limited in order to avoid excessive starting current and to protect the motor. Current limit is applied to the SCR motor control in a manner similar to IR compensation except that the feedback voltage is held off until maximum armature current is exceeded, and it acts on the phase shift unit to reduce conduction angle.

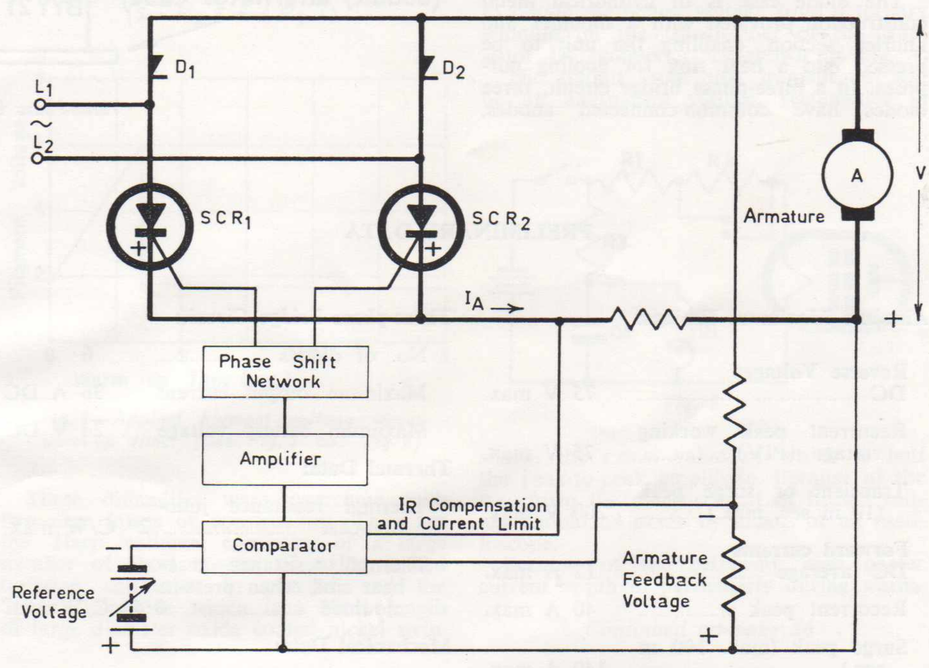


Fig. 15—Block diagram of a variable speed SCR motor control system

R. T. Smith
Applications Laboratory, Sydney

References:

1. 'The Junction Transistor and its Applications.' E. Wolfendale (Editor). Heywood and Co. Limited, 1958.
2. 'Mullard Reference Manual of Transistor Circuits.' Mullard Ltd., Second Edition, 1961.
3. Blecher, F. H. 'Transistor Circuits for Analog and Digital Systems'. Bell System Technical Journal, Vol. 35, March 1956.

Mullard

NEW RELEASE — BYY20/21 SILICON DIODES

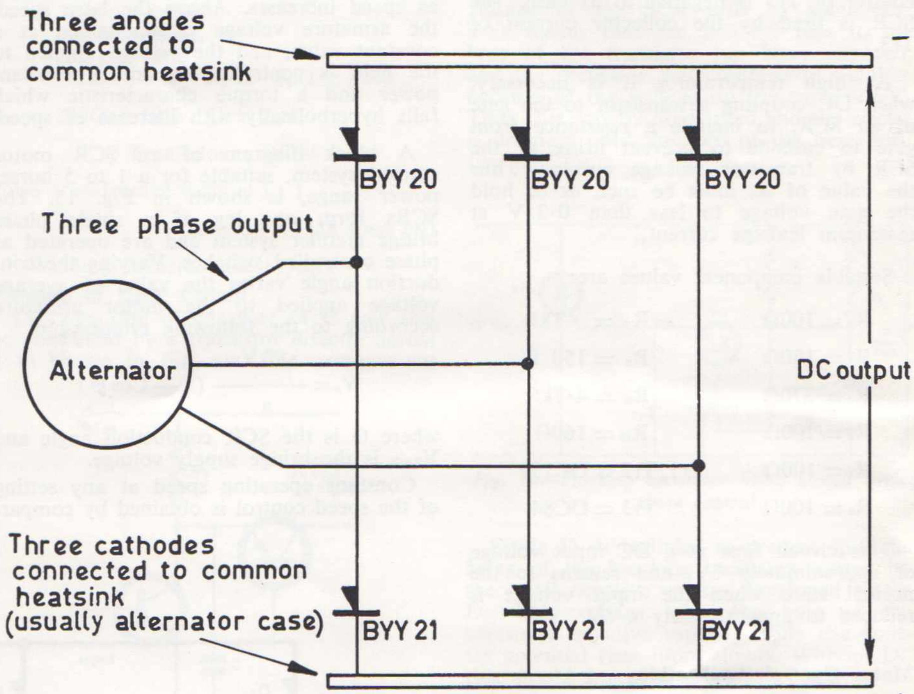
Two new diodes introduced to the Mullard range are the BYY20 and BYY21, diffused silicon units developed particularly for the automotive industry.

There is an increasing trend towards the use of an AC generator, or alternator, for the generation of electrical power in motor vehicles due to the inability of the conventional DC generator to handle the ever-increasing power demands of the modern vehicle. Normal technique uses three phase generation in the alternator, and a three phase bridge circuit to convert the alternator output to direct current.

The electrical properties of the BYY20 and BYY21 have been particularly tailored for this application. Arranged as a three-phase bridge, six diodes can provide a charge power to a 12 V storage battery of 450 W, which is more than sufficient to meet the demands of cars fully equipped with current-consuming accessories. The voltage ratings of the diodes are such that they may be used in 24 V systems, with a corresponding power output of approximately 850 W.

The diode case is of cylindrical metal construction provided with a shoulder and knurled section, enabling the unit to be pressed into a heat sink for cooling purposes. In a three-phase bridge circuit, three diodes have common-connected anodes, while the other three diodes have common-connected cathodes. To enable all of the diodes to be mounted onto a minimum number of two heat sinks, they are therefore manufactured in two versions: the BYY20 has anode connected to case, while the BYY21 has cathode connected to case. The two types are otherwise identical. If required, the flat bottom of the diode may be soldered onto a heatsink.

These diodes can also be used with advantage in mains-operated battery-charging service at 6, 12 or 24 V. They are suitable for incorporation in both domestic and commercial chargers. In addition, they are useful in power supplies for equipment using transistors, and other low voltage high current applications.



Three-Phase Bridge Circuit

PRELIMINARY DATA

Absolute Maximum Ratings:

Reverse Voltages

DC	75 V max.
Recurrent peak working voltage (PIV)	75 V max.
Transient or surge peak (10 m sec. max.)	200 V max.

Forward currents

DC average	12 A max.
Recurrent peak	40 A max.
Surge peak (max. 100 m sec.)	140 A max.

Temperatures

Storage	−65 to 150°C. max.
Junction	175°C. max.

Characteristics:

Forward voltages at Tj = 25°C.

at 12 A forward current	1·15 V max.
at 40 A forward current	1·25 V max.

Reverse current at PIV Tj = 140°C.	4mA max.

Three-phase Bridge Circuit:

No. of diodes	6
Maximum output current	36 A DC
Maximum battery voltage	24 V DC

Thermal Data:

Thermal resistance junction-case	2° C/W max
Thermal resistance case-heat sink when press-in mounted	0·5° C/W max.

Mechanical Data:

(See diagram)

Press Mounting

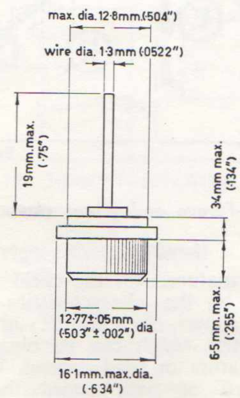
Diameter of hole in heat sink	0·4965″– 0·4985″ (12·61 — 12·66 mm.)
Maximum force to press-in diode	660 lbs. (300 kg.)

Soldering

Max. time 30 sec. at max. solder temperature	235°C.

Connections

BYY20 (blue markings)	case-anode
BYY21 (red markings)	case-cathode



QUICK HEATING TRANSMITTING VALVES

Considerable advances have been made in the development of transistors suitable for VHF mobile equipment, making possible light weight compact equipment of very low battery consumption. However the semiconductor art is still not sufficiently advanced to provide medium power amplifier stages. Till now indirectly heated valves have performed this function but the wasted power during standby is considerable. This is particularly so in taxi equipment where the receiver is on for long periods, and the transmitter must be in a standby state ready for an immediate reply. In this case the transmitter is used for very short periods only.

The ideal solution has been found in a quick heating filament of special construction known as a 'Harp' cathode. With valves of this type savings of up to 90% in battery drain can be realised, providing a warm-up time of 0·8 seconds.

Using this technique Mullard have developed a complete new range of quick heating transmitting valves offering the following advantages:

1. As the filament is only switched on for short periods the working life of the valve only occurs during actual transmission.

2. The ambient temperature of an equipment is lower, thus simplifying the problems of providing adequate cooling and DC stabilization of associated semiconductor circuits.

3. The average power consumption is considerably reduced.

4. Interface effects which can occur with indirectly heated valves are non-existent.

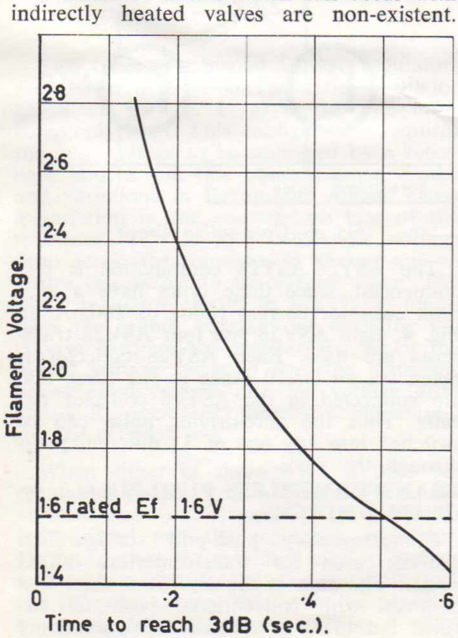


Fig. 1

THE 'HARP' CATHODE

The type of filament which has previously been used in small directly heated valves does not have the required characteristics, because it has a relatively large inductance and is not capable of withstanding the shock and vibration to which mobile equipment is subjected, because it is long and relatively thin.

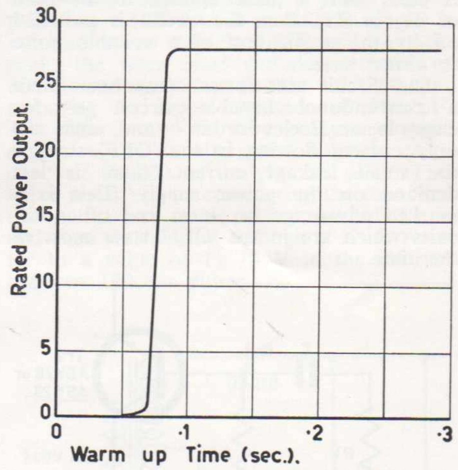


Fig. 2—Warm-up time of 28W PA (See Fig. 5 and Fig. 6)

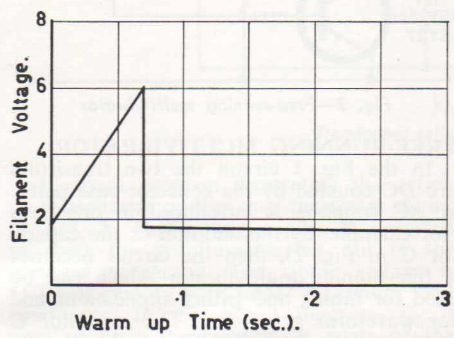


Fig. 3—Applied filament voltage versus warm-up time. (See Fig. 5 and Fig. 6)

These difficulties were overcome with two new forms of construction. They are the 'Harp' cathode, consisting of a large number of short lengths of oxide coated tungsten connected in parallel and the 'Ribbon' cathode which is a short length of large diameter oxide coated nickel strip. The 'Ribbon' cathode has a longer heating time than the 'Harp' construction but both types have a low inductance.

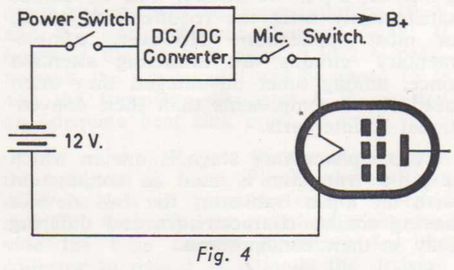


Fig. 4

FILAMENT SUPPLY VOLTAGE

The cathode structure of the quick-heating valves has been designed for the utmost in reliability and freedom from vibration faults such as microphony. For this reason the heater voltages are lower than usual which may appear to present heater supply difficulties, however there are several ready solutions. The most economical method is to obtain the heater voltage from a tertiary winding on the transformer of the transistor DC converter, as shown in Fig. 4. The output from this transformer will be an alternating voltage, rectangular in

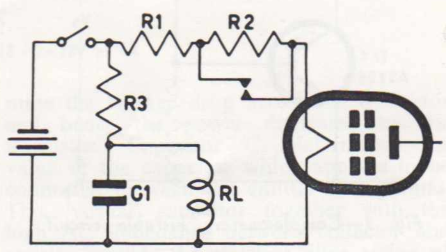


Fig. 5

shape, the r.m.s. value of which is half the peak-to-peak amplitude. Because of the waveform the measurement of heater voltage should be made by means of an oscilloscope.

Because of the relatively high heater current required, particularly during warm-

Continued on page 36

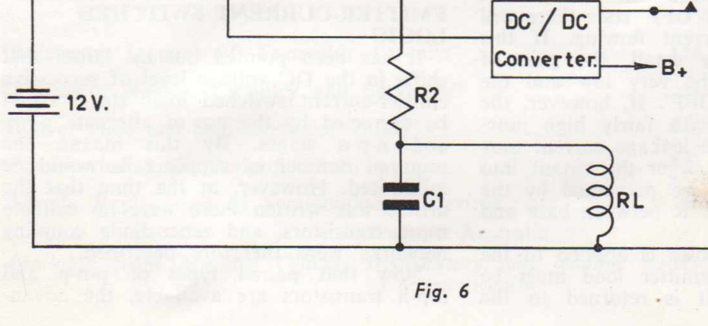


Fig. 6

COMPLEMENTARY CIRCUITS USING P-N-P AND N-P-N TRANSISTORS

COMPLEMENTARY CIRCUITS

Conventional circuit design, using only p-n-p or n-p-n transistors, can of course satisfactorily meet the requirements of all or most applications. However, 'complementary' circuits are attracting attention since, among other advantages, they often need fewer components than their conventional counterparts.

A complementary stage is one in which a p-n-p transistor is used in conjunction with an n-p-n transistor, the two devices having similar characteristics and differing only in their configuration.

When successive p-n-p and n-p-n circuits are used in cascaded linear amplifying stages, DC coupling can be used without the introduction of voltage level difficulties. The circuits of Figs. 3 and 5 illustrate this. Not only do voltage shifts cancel out; the effects of V_{be} temperature coefficients tend to cancel out as well.

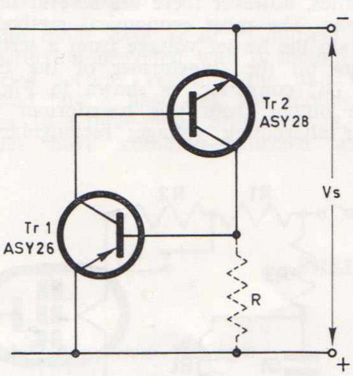


Fig. 1—Complementary bistable circuit

BISTABLE CIRCUIT

A simple bistable switching circuit is shown in Fig. 1. This circuit has been fully analysed by Wolfendale[1]. Its operation, in brief, is as follows.

The circuit may be switched on by the application of a triggering pulse to either the emitter of the p-n-p transistor Tr_1 or the base of the n-p-n transistor Tr_2: All the collector current of Tr_1 is then available as base current for Tr_2; and the collector current of Tr_2 is available as base current for Tr_1.

In the 'ON' state, under normal operating conditions with several milliamperes available, the two transistors will be hard on and bottomed. In the 'OFF' state there will be only leakage current flowing. If this current is sufficiently small, the gain of each transistor will be very low and the circuit will remain 'OFF'. If, however, the circuit is operating with fairly high junction temperatures, the leakage current may be great enough to trigger the circuit into conduction. This can be prevented by the inclusion of a resistor R between base and emitter of Tr_1.

If the triggering pulse is applied to the emitter of Tr_1, an emitter load must be provided. The circuit is returned to the 'OFF' state by interruption of the current flowing in the emitter of Tr_2.

If the alternative method of triggering is used, with a pulse applied to the base of Tr_1 or Tr_2, then the circuit is switched off by the application of a suitable pulse to either base.

The circuit uses fewer components than a conventional bistable circuit — for example, an Eccles-Jordan — and, since the only current flowing in the 'OFF' state is the small leakage current, there is less demand on the power supply. This is a marked advantage in alarm and other circuits which are in the 'OFF' state most of the time.

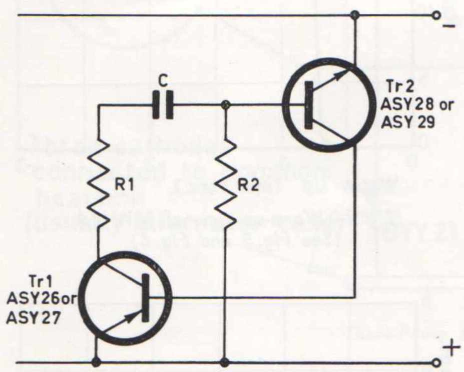


Fig. 2—Free-running multivibrator

FREE-RUNNING MULTIVIBRATOR

In the Fig. 1 circuit the two transistors are DC coupled by the collector-base paths. If AC coupling is introduced in one path (for example, by the addition of the capacitor C in Fig. 2), then the circuit becomes a free-running multivibrator which can be used for timing and gating applications and for waveform generation. The capacitor C is charged through R_2. The discharge current is limited by R_1.

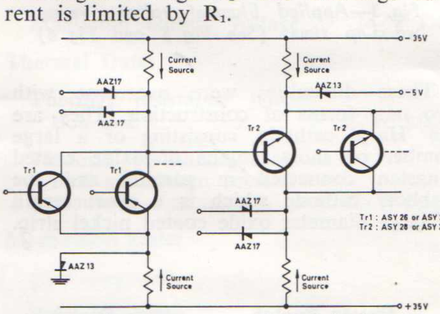


Fig. 3—Emitter-current switched logic, with alternate p-n-p and n-p-n stages

EMITTER-CURRENT SWITCHED LOGIC

It has been pointed out by Miles[2] that shifts in the DC voltage level of successive emitter-current switched logic stages could be corrected by the use of alternate p-n-p and n-p-n stages. By this means, the required number of supply rails would be minimised. However, at the time that the article was written there were no suitable n-p-n transistors, and zener-diode coupling networks were therefore described.

Now that paired types of p-n-p and n-p-n transistors are available, the advantages of emitter-current switching (notably a 50% improvement in switching speed over base-current switching) can be retained while eliminating the zeners.

Typical p-n-p and n-p-n stages are shown in Fig. 3. In this circuit the ASY26 may be followed by the ASY28, or the ASY27 by the ASY29. The ASY27-ASY29 combination gives a higher branching factor. The AAZ17 act as catching diodes.

CORE-GATING

If, in a core-gating circuit, only p-n-p or n-p-n transistors are used, then the base drive voltage must swing through the full voltage that appears across the store. In a p-n-p/n-p-n complementary circuit, such as that shown in Fig. 4, the required swing is only the V_{be} of the transistors.

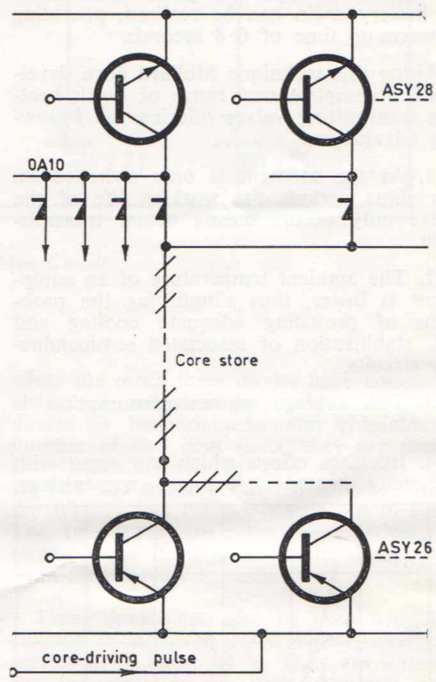


Fig. 4—Core-gating circuit

The ASY26-ASY28 combination is recommended, since these types have a high peak collector current rating of 200mA. In Fig. 4, eight ASY26 and four ASY28 transistors are used. Each ASY28 collector is taken via an OA10 diode to the cores that are connected in the ASY26 collector circuits. Thus the core-driving pulse can be switched into any one of 32 different paths through the store.

TRANSFORMERLESS PUSH-PULL OUTPUT STAGES

Complementary push-pull configuration can be used for transformerless output stages. The gain is slightly lower than that obtained with conventional push-pull circuits; but there is a considerable economy in components, since a phase-splitting stage (transistor or transformer) is not required. Even if an n-p-n pre-amplifier is introduced to compensate for the reduced gain, there is still a saving in components.

Continued on page 36

TRANSISTORS AS POWER SUPPLY FILTER ELEMENTS

Until recently, transistorised equipment has been almost exclusively battery operated. Now, however, both domestic and industrial equipment is required to produce and run at higher power levels and, consequently, there is an increasing tendency to design transistorised equipment for mains operation.

Transistor power supplies are usually required to provide large currents and low voltages. Once the current exceeds a few hundred mA, the normally accepted filter components are no longer adequate. The largest problem besetting the designer of high current filters is to obtain a suitable choke. Consequently the specified inductance is often quite low and may even be in the mH range. In order to compensate for this, the filter capacitors must be raised from hundreds to thousands of μF. This in turn raises the problem of providing rectifiers with suitable peak current ratings.

The size and cost of these filter units makes it desirable to simulate them by other components. Regulated supplies are suitable for this purpose and the complexity involved may well be warranted even in a simple supply.

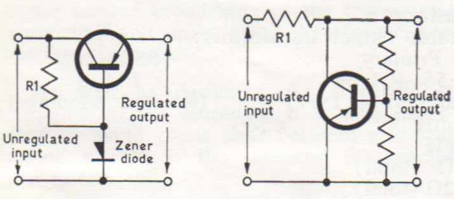


Fig. 1

Two types of constant voltage regulators are shown in Fig. 1. The series regulator as shown in Fig. 1 (a) is the one most commonly used. This configuration requires the base voltage to be stabilised by a reference source (in this case a zener diode) and variations in the output voltage cause a variation in the emitter-base bias of the transistor. This in turn causes the voltage drop across the transistor to change and so return the output to its original value. The shunt regulator shown in Fig. 1 (b) maintains the output constant by changing the current through the resistor R_1. When the output voltage tends to rise, the transistor draws extra current and the voltage drop across R_1, increases, thereby compensating for the original rise.

When either of these types is used for regulation purposes, it is customary to use at least one amplifying stage between the error signal and the regulating transistor. However when the system is used to replace a capacitor, this is not necessary. Indeed, the reference source may even be replaced by a capacitor as shown in Fig. 2. This arrangement operates as if a capacitor were connected between the emitter of the transistor and ground. The value of this capacitor is approximately equal to that of capacitor C_2 multiplied by the current gain of the transistor. It may be seen then, that capacitors of thousands of μF may be simulated economically with standard components, thus decreasing the size of the equipment and increasing the reliability.

In order to demonstrate fully the properties of this circuit, consider a power supply providing 12 V and 1 A. If the ripple is required to be less than 250 mV peak-to-peak, the filter must contain a choke of 10 H and output capacitor of 200 μF, assuming an input capacitor of 500 μF. This filter would be very bulky and quite expensive. As an alternative a single capacitor of 40,000 μF might be used. This value is derived by considering the capacitor to be charged to a peak value of 12 V and then discharging through a 12 Ω resistor to a value of 11·75 V in the interval between 100 c/s pulses.

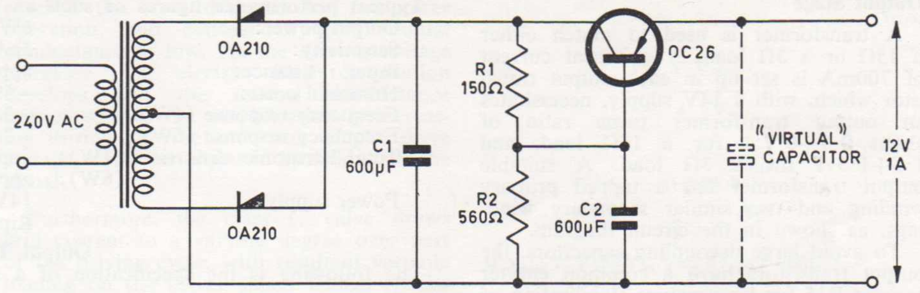


Fig. 2

Transformer rating: 18 – 0 – 18V at 1A

A circuit diagram illustrating the action of capacitance multiplying circuits is shown in Fig. 2. This power supply is designed to meet the requirements set out in the previous paragraph. The ripple voltage is 150 mV peak-to-peak under maximum load conditions. In the circuit diagram, C_1 is used to limit the conduction angle of the diodes. The value of this capacitor is determined by two conflicting factors. If the capacitor is made too large, the peak current necessary to charge the capacitor will exceed the peak current rating of the diodes. On the other hand, if the capacitor is made too small, the ripple voltage will exceed the ripple rating of the capacitor. As a rule of thumb, the ripple should not exceed one eighth of the DC working voltage of the capacitor. In this case the ripple is 2·9 V r.m.s. This value is well within the capabilities of the 25 VW type specified. Increasing the size of this capacitor also reduces the ratings necessary for the transistor. As it is desirable to have a minimum of 2 V between collector and emitter of the transistor, the minimum voltage at the collector must be 14 V. If the peak-to-peak ripple is 10 V, the r.m.s. input voltage will be approximately 19 V and the transistor will dissipate 7 W. If however, the ripple is reduced to 2 V peak-to-peak, the r.m.s. input will be about 15 V and the transistor dissipates 6 W so that an adequate heat sink must be used.

The resistors R_1 and R_2 are chosen with three factors in mind. Firstly, the resistors must be low in value to ensure thermal stability. Secondly, the resistors must provide the base current necessary for the collector to pass 1 A. Finally the division ratio of the two resistors will help to determine the voltage drop across the transistor and hence the power dissipated by the transistor. Capacitor C_2 determines the value of the capacitor which appears to be connected between the emitter and ground. This 'virtual' capacitor together with the high AC impedance between collector and emitter provide the effective filter action of this configuration. It should be emphasised that this 'virtual' capacitor is seen only by the load. It does not affect the rectifying circuit, even though its value is quite large.

It may be seen then that despite the size of the heat sink required, the cost of the transistor and the complexity of the circuit, this configuration has considerable advantages to offer over the conventional filter on the grounds of space, economy and reliability.

R. R. Donohoe
Applications Laboratory, Sydney

HIGH PERFORMANCE AMPLIFIER - 3.5 c/s to 80 kc/s

This audio frequency amplifier is suitable for industrial applications; for example, energising servo systems, transducers, or low-resistance strain gauge systems, where wide frequency response and low harmonic distortion are required.

The OC22 is an alloy junction germanium transistor with cut-off frequency $f\beta$ typically as high as 11kc/s, and a permitted collector dissipation on a 2°C/W heat sink of over 8W at 45°C ambient. This enables a power amplifier using two OC22 transistors in Class A push-pull to deliver 6W output at low distortion.

The circuit diagram of such an amplifier is shown. DC or "semi-DC" interstage coupling is used throughout except for the output transformer itself and, in most applications, the input circuit. This permits good low frequency performance without the use of very large capacitors.

Output Stage

A transformer is used to match either a 15Ω or a 3Ω load. A quiescent current of 700mA is set up in each output transistor which, with a 14V supply, necessitates an output transformer turns ratio of about 0·7+0·7:1 for a 15Ω load, and 1·5+1·5:1 for a 3Ω load. A suitable output transformer has a tapped primary winding and two similar secondary windings, as shown in the circuit diagram.

To avoid large decoupling capacitors, the output transistors have a common emitter resistor R15 for temperature stabilisation of the operating point.

The peak current under sinewave conditions in the circuit shown will be 1·4A. The permitted current in the OC22, averaged over 20ms, is 1·0A. Thus maximum power operation may be maintained above 30c/s. The derating below 30c/s may be linear with frequency to a permissible 3W at 2·5c/s.

Driver Stage

The OC22 has a collector current versus base input voltage characteristic of good linearity when forward biased at 700mA. OC42 transistors connected as emitter followers are, therefore, used to provide a low source-impedance drive for the output transistors.

The 560Ω resistors R13 and R14 in the emitters of the driver transistors are sufficient to prevent reverse bias of the driver stages by leakage current I_{co} flowing in the base of the output transistors in the working temperature range.

Phase Splitter and Input Amplifier

A conventional resistive phase splitter is used, the transistor being an OC42 operated at about 5mA with equal AC collector and emitter loads. The operating point is chosen so that stable voltage bias may be applied to the base of each driver transistor by adjusting R3 and R10. This voltage bias, of course, defines the collector current in the output transistors.

The input amplifier is also conventional, being an OC42 transistor in grounded emitter.

Feedback

25dB negative feedback is applied from the secondary of the output transformer to the emitter of the input stage. It is noted that the base bias potential divider of this stage (R1, R2) is referred to the emitter degeneration resistor R6. Thus feedback is also applied to the base circuit, raising the amplifier input impedance.

Performance

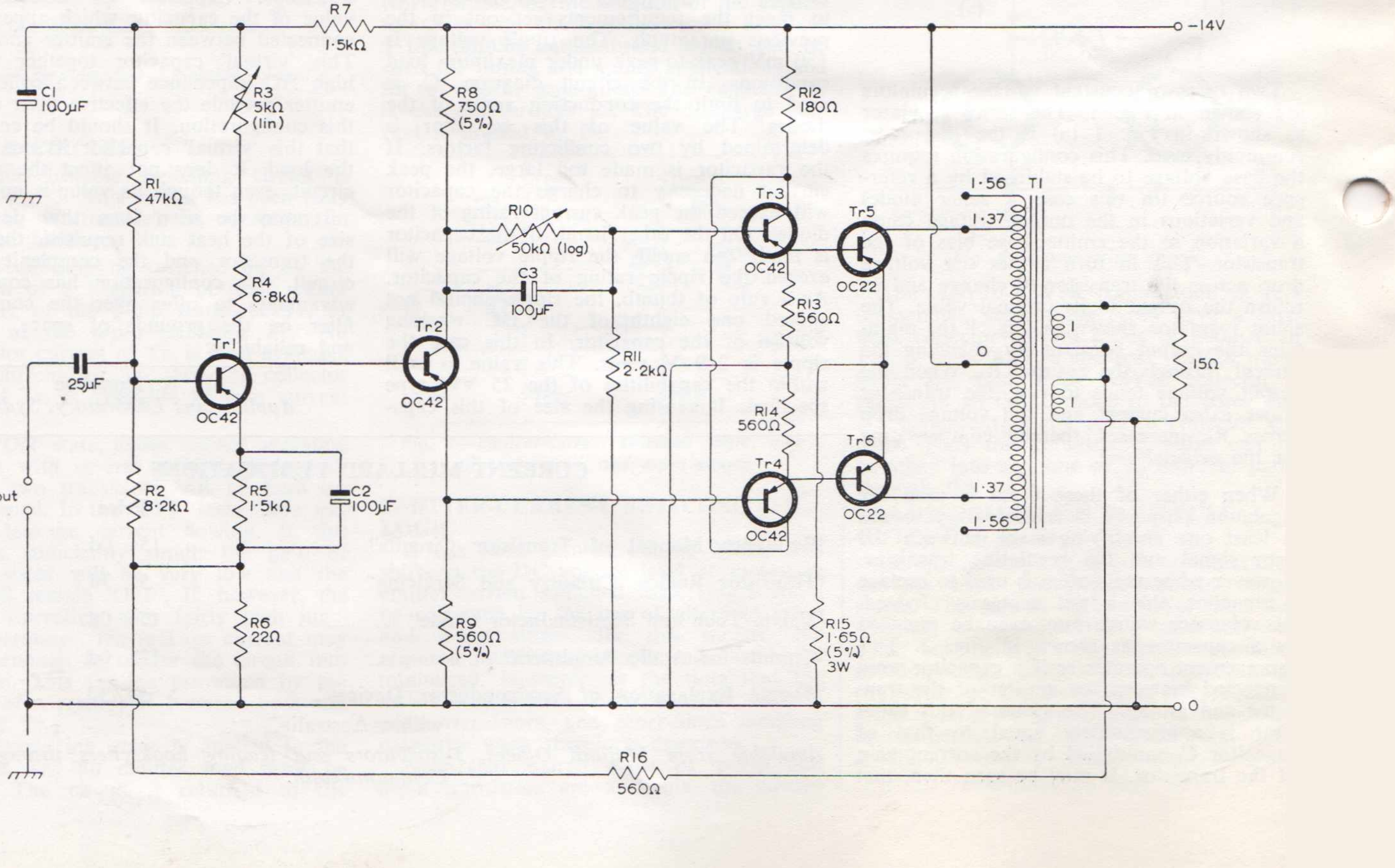
Typical performance figures of such an amplifier are:

Output power:	6W (nominal)
Sensitivity:	5μA (r.m.s.) for 6W output (gain = 68dB)
Input resistance:	approx. 40kΩ
Hum and noise:	−59dB referred to 6W output
Frequency response (1W):	−3dB, 3·5c/s to 80 kc/s
Frequency response (6W):	−3dB, 8c/s to 40 kc/s
Total harmonic distortion (1W):	approx. 0·05%
(6W):	approx. 0·1%
Power supply:	14V DC at 1·5A Ripple < 100mV (r.m.s.)

Output Transformer

The following is the specification of a suitable output transformer:

	Primary	Secondary
Turns ratio:	1·56 + 1·56 (tapped 1·37 + 1·37)	1 + 1 (isolated windings)
Primary inductance:	> 100mH	
Leakage inductance:	< 50μH	
Primary resistance:	< 1·2Ω (total)	
Secondary resistance:	< 0·42Ω (total)	



VALVES FOR LINEAR AMPLIFIERS AND SINGLE-SIDEBAND TRANSMISSION

The aim in designing a linear amplifier is to obtain in the output an exact reproduction of the input waveform, but with a considerable gain in power. This criterion applies directly to audio-frequency amplifiers and vibration frequency generators; in class B telephony or low-power amplitude-modulated transmitters it is the waveform of the envelope which is important, although the contained RF wave and the modulation envelope are reproduced or modified together. A linear amplifier can be used for amplifying AM, FM, FSK (frequency shift keying), PM, and SSB signals. In single-sideband operation the amplitudes of the sideband frequencies must reproduce the varying amplitude of the modulating signal, but the RF output must remain sinusoidal to avoid production of harmonics and intermodulation products—that is, significant output at radio frequencies other than those applied. Such intermodulation frequencies cause serious broadening of the transmitted spectrum, and may cause interference on nearby channels.

As well as designing for linearity of response in the amplifier, it is also necessary to avoid phase shift between grid and anode signals as the sideband frequencies are raised.

Valve Requirements

The requirements outlined above may be achieved by utilising a valve where the anode current is a linear function of the grid voltage; where input and output impedances remain constant throughout the driving cycle, and in which no capacitive or inductive coupling, which might produce positive feedback, occurs between the output and input circuits.

Class A. Operation of almost any valve in class A can provide the required degree of linearity over a large part of its characteristic, but at the cost of efficiency (a maximum of 42·5 per cent for full c.w.). Fig. 1 shows how a change in grid voltage, ΔV_g, results in a change in i_a equal to $\Delta V_g . g_m$, where g_m is the mutual conductance in mA/V.

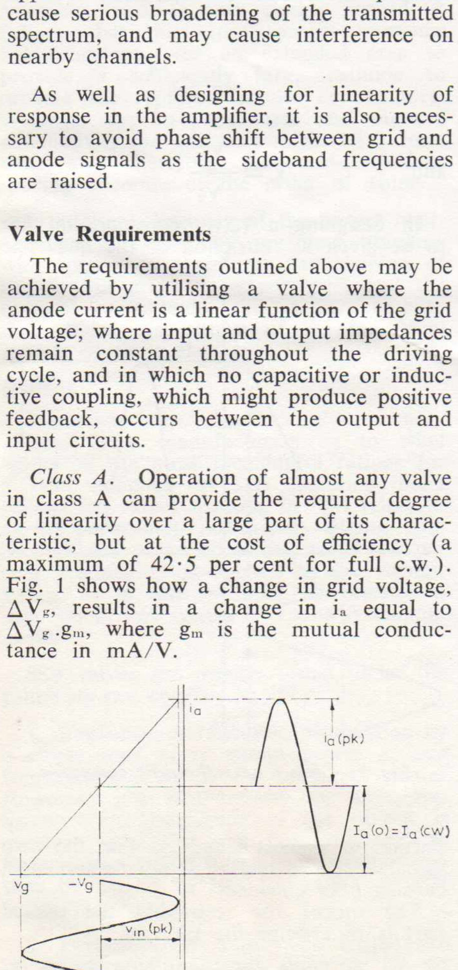


Fig. 1—I_a/V_g curve

Class C. Valves generally provided for class C amplification have not received much attention in design to linearity of characteristics, as high efficiency has been the main aim. Ruggedness and minimum manufacturing costs require safe electrode spacings, and this tells against linear amplification. With such spacings, when the interelectrode potentials due to a driving signal are such that current conduction is very low, the electric field distribution has not reached the state designed for full-power operation, and consequently the mutual conductance is low. As the driving voltage increases, the electric field distribution develops and the mutual conductance improves; thus the amplification is increasing throughout the excursion of the drive voltage with consequent distortion of the signal.

Furthermore, the class C valve draws grid current to a varying degree over part of the driving cycle, with resultant variable loading on the driver stage. Unless special provisions are made, the latter is normally a high impedance source, and consequently the drive voltage on the amplifier is not a linear function of the signal into the driver.

Class B operation provides a compromise between these two. Fig. 2 shows the relationships in the ideal case of a completely

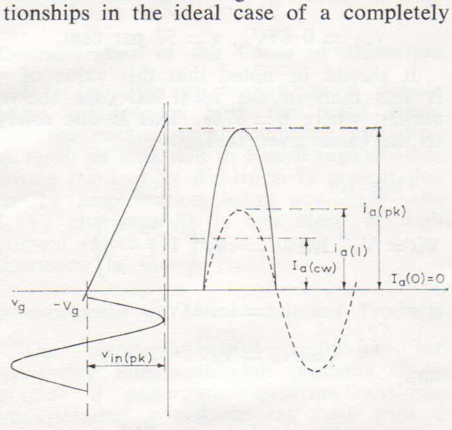


Fig. 2—Class B operation with linear characteristic (single tone)

linear characteristic. A sinusoidal voltage between grid and cathode produces a half-sine pulse of anode current per cycle of driving voltage, with a peak value

$$i_{a(pk)} = v_{in(pk)} . g_m ;$$

and an average value per cycle of

$$I_a = \frac{i_{a(pk)}}{\pi} .$$

In a tuned amplifier the fundamental harmonic is

$$i_{a(1)} = \frac{i_{a(pk)}}{2}$$

The efficiency is

$$\eta_a = \frac{P_{out}}{P_{in}} = \frac{\frac{1}{2} v_{a(1)} \cdot i_{a(1)}}{V_a \cdot I_a} = \frac{v_{a(1)}}{V_a} \cdot \frac{\frac{1}{2}\frac{i_{a(pk)}}{2}}{\frac{i_{a(pk)}}{\pi}}$$

$$= \frac{\pi}{4} \frac{v_{a(1)}}{V_a}$$

$$= 0 \cdot 785 \frac{v_{a(1)}}{V_a}$$

If $v_{a(1)} = 0 \cdot 85 V_a$, $\eta = 67$ per cent.

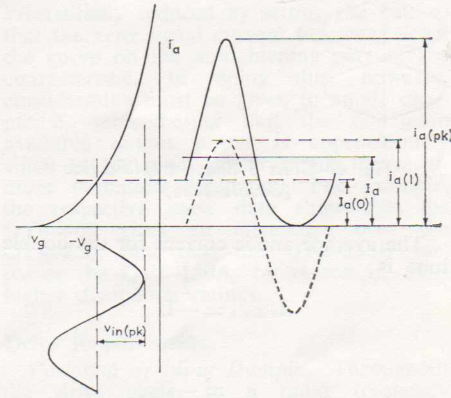


Fig. 3—The effect of curvature on linearity

The practical case is shown in Fig. 3, wherein the curvature of the characteristic causes distortion, but this situation is greatly improved by setting the starting point clear of the curvature. In such conditions, before the driving signal is applied the valve draws anode current $i_{a(o)}$ and dissipates power in the anode given by

$$P_a = V_a \,.\, i_{a(o)} \,.$$

As the driven current rides on this pedestal, it is evident that part of this dissipation must be deducted from the total power available for conversion, and the efficiency is therefore reduced. Hence a compromise has to be made between low distortion at low efficiency and somewhat higher distortion at high efficiency. This condition of operation is class AB. (Class AB_1 when the grid is not allowed to go positive, and class AB_2 when the grid is driven positive.)

The previous comments refer to an unmodulated CW output. Normally, there would be modulation and resultant side frequencies; in a typical single-sideband transmitter the amplifier would be driven by signals at the frequencies ordinarily contained in one sideband, and in its simplest form where only two radio frequencies, whose difference is an audio frequency, say 400c/s, are combined. The driving voltage as a function of time is then

$$V_g = v_{in(pk)}(\cos\omega_1 t + \cos\omega_2 t).$$

This effect is seen in Fig. 4 where, for clarity, class B operation with an ideally linear characteristic is shown. (Only the envelopes are drawn.) The anode pulses rise to a maximum value $i_{a(pk)}$. At this level the mean (or DC) anode current is

$$I_a = \frac{i_{a(pk)}}{\pi}$$

and the crest value of the fundamental harmonic is

$$i_{a(1)} = \frac{i_{a(pk)}}{2}.$$

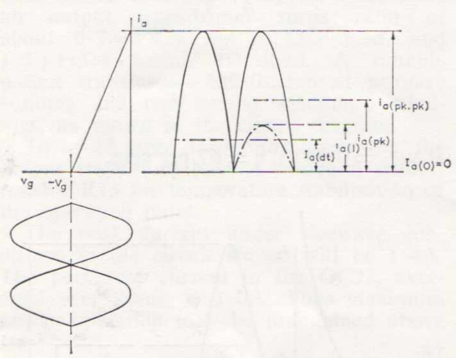


Fig. 4—Class B linear amplification (double tone)

The average anode current for the double tone is

$$I_{a(d.t.)} = \frac{2}{\pi} I_a$$

$$= \frac{2}{\pi^2} i_{a(pk)}$$

The average input power is then

$$P_{in(d.t.)} = V_a \, . \, I_{a(d.t.)}$$

$$= \frac{2}{\pi^2} V_a \, . \, i_{a(pk)}$$

and the average output power is

$$P_{out(d.t.)} = \tfrac{1}{2} P_{out(s.t.)}$$

$$= \tfrac{1}{4} v_{a(1)} \, . \, i_{a(1)}$$

$$= \tfrac{1}{8} v_{a(1)} \, . \, i_{a(pk)} \, .$$

The efficiency is

$$\eta_{(d.t.)} = \frac{\tfrac{1}{8} v_{a(1)} \, . \, i_{a(pk)}}{\frac{2}{\pi^2} . V_a \, . \, i_{a(pk)}}$$

$$= \frac{\pi^2}{16} \cdot \frac{v_{a(1)}}{V_a}$$

Allowing $v_{a(1)} = 0{\cdot}85 V_a$, $\eta_{(d.t.)} = 52{\cdot}5$ per cent.

Translating this ideal class B case into class AB, the zero-signal DC current

$$I_{a(o)} = \alpha V_g^2,$$

where α is the exponent of the characteristic curve.

When the valve is driven by a single-frequency sine wave the average current becomes

$$I_a = \alpha V_g^2 + \frac{\alpha}{2} v_{in}^2$$

and

$$i_{a(1)} = 2\alpha V_g^2 \cdot v_{in}$$

When

$$v_{in} = V_g, \; I_a = \frac{3}{2} \alpha V_g^2.$$

Then the CW efficiency is

$$\eta_{(c.w.)} = \frac{\tfrac{1}{2} v_{a(1)} \, . \, i_{a(1)}}{V_a \, . \, I_a}$$

$$= \frac{v_{a(1)} \alpha V_g^2}{V_a \frac{3}{2} \alpha V_g^2}$$

$$= \frac{2}{3} \frac{v_{a(1)}}{V_a}$$

If

$$v_{a(1)} = 0{\cdot}85 V_a \, , \; \eta = 57 \text{ per cent.}$$

It should be noted that this value of η is less than in the ideal s.t. case shown earlier, where $\eta = 67\%$. This is due solely to the value given to $I_{a(o)}$.

Thus for

$$v_{in} = V_g \, , \; I_{a(d.t.)} = \alpha V_g^2 \left(1 + \frac{1}{\pi} \right)$$

$$= 1{\cdot}32 \alpha V_g^2,$$

$$P_{in(d.t.)} = V_a \, . \, 1{\cdot}32 \alpha V_g^2,$$

and

$$\eta_{(d.t.)} = \frac{\frac{\alpha}{2} v_{a(1)} V_g^2}{V_a 1{\cdot}32 \alpha V_g^2}$$

$$= \frac{0{\cdot}38 v_{a(1)}}{V_a}$$

When

$$v_{a(1)} = 0{\cdot}85 V_a \, , \; \eta = 32 \text{ per cent.}$$

In a multi-electrode valve the electron flow/voltage relationship does not follow the pure Langmuir-law curve with an exponential value of $\frac{3}{2}$, as there is a varying division of current among the several electrodes. Designers take advantage of this fact to try to mould the shape of the anode current characteristic to suit the application.

Accepting the inevitable—that there must be curvature of some kind—the good compromise for a class B linear valve is a characteristic curve of composite function, the tail having a quadratic form ($i_a = \alpha v_{in}^2$) and the main part being a linear tangential extension $i_a = 4\alpha(v_{in} - V_{g1})$. V_{g1} is the value of grid voltage at which the tangent, projected, intersects the axis (see Fig. 5).

This combination operates in two overlapping phases:

(a) The projected V_{g1} provides a biasing point for class A working at low signal level.

(b) As the drive voltage increases, the conditions change smoothly into class B where the straight part predominates.

Driving with a single frequency signal produces an anode current whose mean value is

$$I_{a(c.w.)} = \frac{\alpha V_g^2}{\pi} \left[3 \tan \omega t_{(1)} + (2 + V^2) \left(\frac{\pi}{2} - \omega t_{(1)} \right) \right]$$

where

$$\omega t_{(1)} = \text{arc} \cos \frac{1}{V}$$

and

$$V = \frac{V_{in}}{V_g} \, .$$

In designing a valve, consideration has to be given to restriction of the quadratic

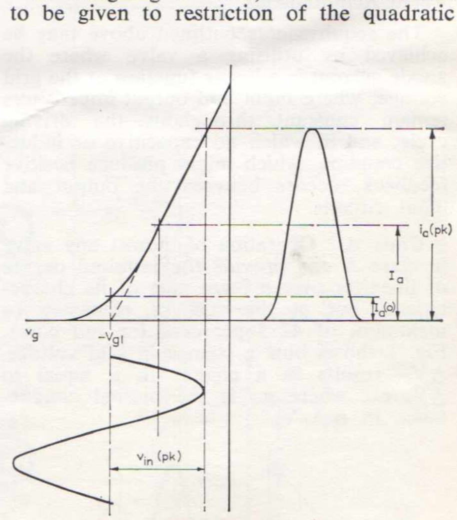


Fig. 5—Class B operation with composite characteristic

portion to a short length so that the zero signal anode current shall not be too high, causing poor efficiency.

The means for restricting the curved part is to arrange the ratio of

$$\frac{\text{grid winding pitch}}{\text{grid-to-cathode spacing}}$$

to limit what we know as the 'island' effect. This name describes pockets of electron starvation which appear between the cathode and the grid wires due to uneven field, and which therefore reduce the average electron density and hence the mutual conductance.

To straighten the upper portion of the curve it is necessary to arrange a gradual taking over of the electron current by a grid, and this is mainly a matter of screen grid geometry. There is another drawback here, however. If the screen grid draws high current at the peak of drive the anode is robbed so that the mutual conductance of the valve falls away again. So the ratio of anode current to screen grid current must be kept high.

Referring now to the grid loading, to obtain the required peak anode current with $V_{in(pk)}$ never exceeding the bias voltage, the screen grid of a tetrode has to raise a sufficiently strong positive potential in the plane of the control grid to replace the positive excursion no longer included in the driving signal. Therefore the valve must be designed for high V_{g2} and low inner μ which means a close grid-to-screen spacing and a low ratio grid wire diameter/grid pitch. For low island effect the ratio grid pitch/g–k spacing must be low, and from these considerations the required ratio grid wire diameter/g–k spacing is decided.

It can be seen from this that the overall assembly becomes very closely spaced; it follows that this close spacing must be held constant over an extended area to provide a sufficiently large cathode to provide the required mutual conductance. With an eye to ruggedness and thermal conductivity the grid wire might be 0·2mm dia., and therefore the grid-to-cathode spacing becomes of the order of 1mm.

In addition, linearity of the output characteristic is aided by ingenious thinking with respect to the form and shaping of the control grid, the screen grid, and the filament or cathode, and in particular to the relative dispositions of the individual strands of these electrodes.

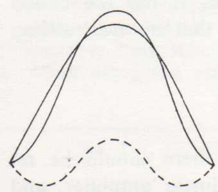
Ratings

There has been some lack of uniformity among valve manufacturers as to what might be the most meaningful ratings for a valve designed for SSB operation. Most ratings for valves working in AM relate to carrier operating conditions, taking into account the peaks occurring when 100 per cent modulation is applied to the carrier. When, as in SSB working, the carrier is suppressed, this system of rating is inapplicable.

SSB valves are usually tested under the following two conditions:

1. *Single-tone modulation.* Modulation by a single tone gives output power at one frequency which is the nominal carrier frequency plus (or minus) the tone frequency. The amplitude of the output is constant, and this can be considered as linear CW amplification. The amplifier can thus be driven to full telegraphy or FM limits.

2. *Two-tone test.* Two radio frequencies of equal amplitude but differing by an audio frequency are fed to the amplifier under test. The peak-to-mean ratio of this complex signal is two to one. Thus the PEP is twice the mean power obtained. With normal speech, the peak-to-mean ratio is between three and five, and to limit the peak condition to the CW rating for the valve in speech application could be wasteful of the valve's capabilities.



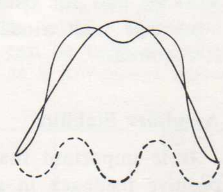


Fig. 6—Third order intermodulation distortion arising from second harmonic

Fig. 7—Fifth order intermodulation distortion arising from second and third harmonics

The Mullard data will therefore include the following:

1. An SSB linear amplifier for signals with a peak-to-mean ratio between one and two. Operating values will be given for both single and double-tone modulation, but the peak envelope power for the double-tone case will be no more than the CW power level of the single-tone case.

2. An SSB linear amplifier for signals with a peak-to-mean ratio of two or greater. Here the permitted maximum envelope power will be stated. It is important that in this case a single frequency should not be used to the full PEP rating. Individual data sheets will give full information.

The intermodulation distortion figures given for d_3 and d_5 will be referred to the amplitude of one of the two-tone frequency signals. If referred to the PEP, these figures would be increased by 6dB.

Demonstration of the Cause of Distortion

It is possible to display on a double beam oscilloscope both the input signal and the output signal from the amplifier. A result as indicated in Fig. 6 suggests that strong third order distortion is present due to the amplification being non-linear—the I_a/V_g characteristic is too steep at high current values. In Fig. 7, strong fifth order distortion is shown.

Measurement of Intermodulation Products

The most successful technique for measuring intermodulation products is to employ a panoramic spectrum analyser incorporating a cathode ray tube with a time base arranged to display the various frequency components and a vertical deflection graticule calibrated to compare the amplitudes of the components. With suitable equipment, up to 60dB can be measured.

Application Notes

Bias

For an SSB linear amplifier, the bias is set to class AB_1 with the possibility of running slightly into grid current on the drive peaks. The anode tuned circuit maintains the sinusoidal wave form on the negative-going peaks so that it is not necessary, as with audio operation, to run two valves in push-pull with this class of bias.

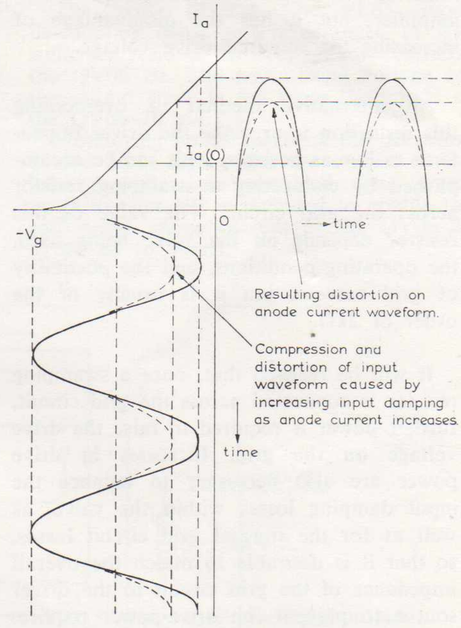


Fig. 8—Variation of input damping

When a valve is chosen whose anode current characteristic possesses a prolonged region of curvature at the low-current end, waveform distortion and resultant intermodulation products can generally be substantially reduced by setting the bias so that the zero signal current lies away from the curve on the straightening part of the characteristic. In doing this, however, consideration must be given to anode dissipation, remembering that the maximum available output power is approximately equal to the input power minus the maximum permitted dissipation. For example, the respective valve data show that the QY4–400 gives an advantage over the QY4–250; and that the QV2–250B surpasses the QV1–150A, by reason of their higher dissipation ratings.

Drive Requirements

Variation of Input Damping. Throughout the drive cycle, in a radio frequency amplifier biased at AB_1, the input damping changes, and these changes are reflected back to the driver valve, as shown in Fig. 8. The changes in input damping are caused by the change of input capacitance of the valve with increasing anode current. As the anode current starts to flow, electrons move away from the cathode and pass through the grid on their way to the anode. The proximity of this increasing stream of electrons to the grid wires modifies the grid-cathode capacitance. This increased capacitance requires additional current to charge it from the driver source, and as there is always some resistive loss associated with the capacitance, more power is required.

An additional difficulty is the grid-circuit tuning effect. This is often encountered in IF amplifiers, and a usual method of reducing this change of capacitance is to leave part of the cathode resistor unbypassed. This method gives a degree of negative feedback which may assist in reducing the distortion in an SSB linear amplifier, but it has the disadvantage of increasing the required drive voltage.

An alternative method of overcoming this distortion is to make the driver impedance as low as possible. This can be accomplished by connecting a swamping resistor across the grid circuit. The value of this resistor depends on the valve being used, the operating conditions, and the possibility of grid current, but it is usually of the order of 2kΩ.

It will be realised that, once a swamping resistor is connected across the grid circuit, further power is required to raise the drive voltage on the grid. Increases in drive power are also necessary to balance the input damping losses within the valve, as well as for the normal grid circuit losses, so that it is desirable to match the overall impedance of the grid circuit to the driver source to prevent the drive-power requirement becoming too high. The swamping resistor assists when the valve is run slightly into grid current, and also improves stability.

Grid Current. The characteristics of certain valve types are such that it is advantageous to drive the valve slightly into grid current. In this instance, a swamping resistor will assist in masking the effects of input impedance changes when grid current starts to flow.

Summary of Drive Requirements

Summarising the drive requirements for a class AB_1 linear amplifier for SSB operation, it may be said that drive power is required to supply the following:

1. Grid circuit losses and coupling circuit losses;
2. input damping losses of the valve;
3. loss in the grid circuit swamping resistor; and
4. the small amount of grid current which may flow.

Common Grid Operation

An alternative arrangement is provided by the common grid (or grounded grid) circuit. In this technique the driving impedance is low (approximately equal to $\frac{V_{in(pk)}}{i_k}$) and the variable loading of any grid current has a proportionally diminished effect.

The drive power requirement is increased to

$$P_{load(driver)} = \frac{V_{in(pk)} \cdot i_{k(pk)}}{2} + P_{ct},$$

but of this

$$\frac{V_{in(pk)} \cdot i_{k(pk)}}{2}$$

is transferred to the output circuit.

This system is of particular value in AB_2 working and for triodes. It has the added advantage with triodes that no neutralising is required.

Amplifier Stability

It is important that there should be no positive feedback in a linear amplifier, and neutralising may be necessary with some valves, depending on their construction and power gain. The QV06–20 usually requires neutralising, as do some of the valves which may be used in low-power stages.

Valves in the medium-power tetrode series, QY3–125, QY4–250, QY4–400, do not require neutralising when adequate shielding is provided between input and output circuits.

Although the discussion here relates to operation of the amplifier valve, a complete transmitter may incorporate waveform correction circuits.

Anode Tuning and Matching

The maximum output from a tuned amplifier should always occur at the anode current *dip* when the anode circuit is being tuned. In practice, it is found that there is a certain amount of compensation in the tuning and matching, where the maximum output can be obtained for practically the same valve input. It would appear, therefore, that the tuning and matching of a linear amplifier is a simple operation. However, when the output from the amplifier is continuously monitored for intermodulation distortion products (whilst the tuning and matching is being carried out), it is seen that only one tuning and loading position gives the lowest distortion, and up to 10dB increase of intermodulation products can result with apparently correct tuning.

In practice, it will usually be found that the lowest intermodulation distortion figures can be obtained when the fifth and third order products are nearly equal in amplitude. This is because some cancellation of the third order product can be achieved by careful alignment; normally, of course, one would expect the third order product to be considerably worse than the fifth. The reason for this partial cancellation of the third order product can be explained by reference to Figs. 9a, 9b and 9c.

Fig. 9a shows a typical I_a/V_g curve with the curvature exaggerated. Fig. 9b is a curve of drive voltage against output voltage of a linear amplifier showing the effect of the increasing input damping with medium rate of change of curvature when grid current starts to flow. Fig. 9c shows the overall amplification characteristic including the effect shown in Figs. 9a and 9b (the curvature is again exaggerated to illustrate the effects). The result of this compound characteristic (Fig. 9c) is that the third order intermodulation product is largely cancelled and there is little or no effect on the fifth order product.

When an amplifier is set up to give this partial cancellation of the third order product, then as the drive is progressively

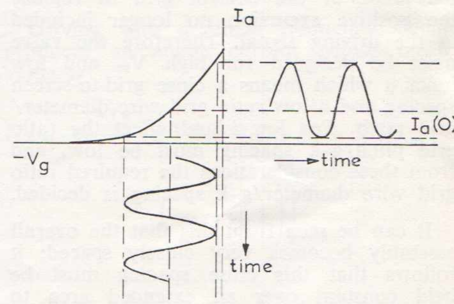
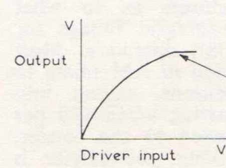


Fig. 9a—Typical I_a/V_g curve with exaggerated curvature

reduced the fifth order product decreases in amplitude, but the third order product stays constant, or even increases slightly in amplitude, before it finally reduces. The degree of increase in amplitude of this third order product is usually about 4dB.



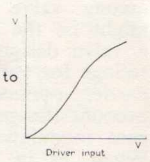


Fig. 9b—Output/driver input curve showing effect of input damping

Fig. 9c—Output/driver input curve showing overall amplification characteristic

The cancellation effect of the third order intermodulation product is much more pronounced when a small amount of grid current is flowing, and thus it is possible to eliminate this intermodulation entirely. There is little to be gained from this elimination, however, as the fifth order product remains unaffected.

Two Stage Linear Amplifier Operating at 3·1Mc/s

This amplifier gives an output of 20W PEP, with intermodulation products better than 50dB. The amplifier has a power gain of 250 without using negative feed-back. The circuit is shown in Fig. 10.

Circuit Description

The input to the amplifier is fed from a 75Ω source, and a drive control RV2 provides an adjustment covering a minimum of 2dB change of drive voltage to the grid of EL85. The source is arranged to give an output which can be varied in 2dB steps by an attenuator, so that RV2 permits accurate setting to any desired output level. R5 damps the anode circuit of the EL85 and also swamps the change in input damping of the QV06–20. A circuit to tune and match the anode to the load is included in the output stage. Also included in the output stage is a variable load which can be used if desired when using the unit as a low-power driver.

The value of C2 (the EL85 anode de-coupling capacitor) is arranged to give sufficient residual RF voltage to neutralise the QV06–20. The neutralising voltage for the output stage is fed from the anode of the QV06–20 via C6.

Monitoring points to check the distortion of the input signal and the output of the final amplifier are provided. The preset potentiometer R9 is included to enable the zero-signal anode current of the QV06–20 to be set to 50mA.

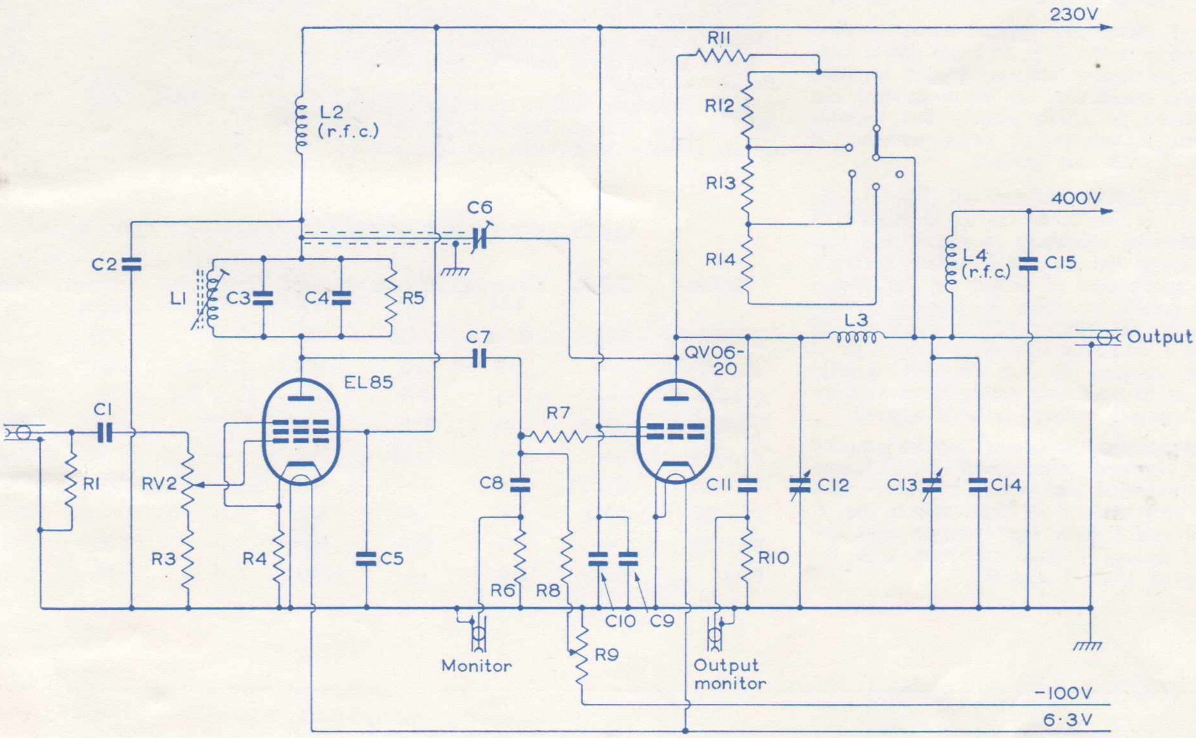


Fig. 10—Two-stage linear amplifier

THE RANGE OF MULLARD TRANSMITTING VALVES FOR SSB

Linear RF Power Amplifier, Class, AB_1 SSB Suppressed Carrier Service. Typical Operation at f<30Mc/s. 'Two Tone' Modulation, Max. Signal Conditions.

	QV06-20	QY3-65	QV1-150A	QV08-100	QY3-125	QV2-250B	QY4-250	QY4-400	QY5-500	QY5-800	QY5-3000A	
V_a	600	3000	1250	750	3000	2000	4000	4000	5000	4000	5000	V
V_{g2}	200	360	300	310	600	350	500	700	700	600	1000	V
V_{g1}	–42	–85	–45	–45	–108	–56	–93	–120	–90	–110	–117	V
$I_{a(o)}$	26	15	100	130	23	100	50	85	56	150	200	mA
I_a	72	45	180	270	77	190	115	180	180	330	380	mA
I_{g2}	5.0	1.5	4.4	26	7.0	4.0	3.0	5.0	15	40	25	mA
I_{g1}	0	0	0	0	0	0	0	0	0	0	1.5	mA
$V_{in(pk)}$	47	85	45	45	108	56	93	115	90	100	134	V
$P_{load(driver)}$	0.25	1.0	1.0	1.5	1.0	1.5	1.0	1.5	2.0	1.5	2.5	W
p_a	20	70	140	93	117	220	188	427	495	670	1115	W
p_{g2}	1.0	0.5	1.3	8.0	4.2	1.4	1.5	3.5	3.8	24	25	W
PEP (out)	46	130	172	220	228	320	454	586	810	1300	1570	W
$P_{out(mean)}$	23	65	86	110	114	160	272	293	405	650	785	W
η_a	54	48	38	55	49	42	56	41	45	49	41	%
PEP (load)	38	111	146	187	194	272	383	498	688	1110	1335	W
$D_{i.m.}$	30	28	33	28	29	28	36	40	28	35	30	dB

QUICK HEATING TRANSMITTING VALVES

Continued from page 27

up, care should be taken to keep lead resistances to a minimum. A fall in heater voltage below the rated value at the time of switch-on will considerably lengthen the warm-up time. On the other hand, the warm-up time may be reduced by increasing the voltage during the switch-on period.

Fig. 1 shows the typical warm-up time required to reach 70% of rated power output versus heater voltage. It can be seen from this graph that the warm-up time can be reduced to approximately 0·1 seconds if a surge voltage of twice nominal is applied during this period.

In the circuit diagram of Fig. 5, the relay RL is shorted across R_2 (a portion of the resistance necessary to reduce the supply voltage to the rated heater voltage). Upon application of power by the microphone switch, the relay is delayed in pulling in by the timing circuit consisting of R_3 and C_1. During this delay the surge of voltage applied to the filament quickly brings it to operating temperature and the normal heater voltage is then applied.

Alternatively, the heater may be supplied from a filament winding of the DC converter, switched into circuit by the delayed relay. This method is illustrated in Fig. 6. Figs. 2 and 3 show the warm-up time and applied heater voltage obtained with the circuits of Figs. 5 and 6.

As the frequency of the filament voltage may be in the audio band, hum modulation caused by AC heating of the filament can occur. Measurements indicate that the hum modulation caused by a 500 c/s heater supply is as follows:

1. Transformer made to commercial specification with one side of the filament earthed — 30 dB.
2. Transformer made to commercial specification with centre-tap earthed — 50 dB.
3. As above with suitable hum bucking circuit — 60 dB.

Slight detuning with a consequent reduction in drive voltage, especially in the grid circuit of an output stage, can increase the hum level by approximately 10 dB.

Fig. 7 illustrates a satisfactory type of hum bucking circuit, the HT negative terminal being earthed.

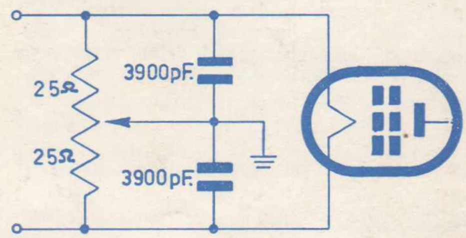


Fig. 7

MULLARD QUICK-HEATING TRANSMITTING VALVES

Mullard	E.I.A.	Filament (V)	Ratings (A)	V_a max (V)	Anode Dissipation (W)	Freq. (Mc/s)	Warm-up Time (sec)	Output (W)
QQZ03-10	7983	3·15	1·65	250	2 × 7	200	0·8	14
QQZ03-20	—	1·5	4·5	600	2 × 10	200	0·8	45
QQZ06-40	—	2·3	4·5	600	2 × 20	200	0·8	85
QZ06-20	8042	1·6	3·2	400	25	175	0·4	35
YL1000	—	1·1	1·05	200	—	175	0·4	2·8
YL1020	8118	1·6	—	600	2 × 12·5	200	0·5	35
YL1080	8348	1·6	—	300	2 × 4·0	—	0·5	—
YL1130	—	1·1	3·1	300	—	500	0·5	8·0
8300	8300	2·0	—	1000	2 × 34	30	0·1	140 PEP

COMPLEMENTARY CIRCUITS

Continued from page 28

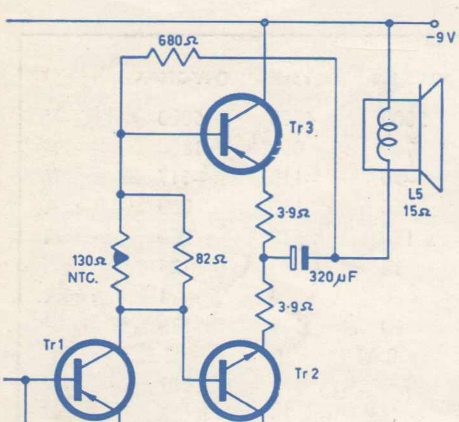


Fig. 5—300mW transformerless push-pull output stage

A 300 mW amplifier including an n-p-n common emitter input stage has been described by Chorley[3]. The output stage of this amplifier is shown in Fig. 5.

COMPLEMENTARY TRANSISTOR TYPES

The ASY26 and ASY27, p-n-p transistors, have an established position in the field of switching applications. Two new types, the ASY28 and ASY29, are n-p-n counterparts of the earlier types. The characteristics of the four transistors fall into two matched group as shown below.

R. T. Smith
Applications Laboratory, Sydney

REFERENCES:

Ref. 1. WOLFENDALE, E. 'The Junction Transistor and its Applications' (pp. 292 to 301). Heywood and Co., London, 1958.

Ref. 2. MILES, R. J. 'Output-coupling Networks for use with Logical Circuits of the Emitter-current Switching Type'. Mullard Technical Communications, Vol. 5, No. 47, March, 1961. (pp. 295 to 300).

Ref. 3. CHORLEY, R. B. 'A 300 mW Complementary Audio Amplifier'. Mullard Outlook, Vol. 6, No. 1, January-February, 1963.

P-N-P	N-P-N	V_{CB} max. V	V_{CE} max. V	I_{CM} max. mA	P_{tot} max. mW	f_1 Mc/s	h_{FE} (I_c = 10 mA)
ASY26		— 30	— 25	200	125	4	30
	ASY28	+ 30	+ 25	200	125	4	30
ASY27		— 25	— 20	200	125	6	50
	ASY29	+ 25	+ 20	200	125	6	50

All four types are in TO 5 encapsulation with the base connected to the envelope.

Printed in Australia by The Langlea Printery Pty. Ltd. 31 Princes Highway, St. Peters