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Commentary

THE Post Office has recently introduced a number of changes into the public telephone service of Great Britain. To the subscriber, the immediate effect of these alterations is that a number of calls which previously cost up to one shilling now cost only threepence, while the area in which numbers may be dialled direct has been considerably increased.

These changes are, however, only part of a much larger plan which is described in the recent White Paper, "Full Automation of the Telephone System". In this it is stated that the Post Office is taking two big steps towards full automation of the telephone system.

The first step, which has already been taken, is the introduction of simplified charges, and in this section of the Paper it is pointed out that since the first automatic exchange came into service in 1912 progress towards full automation has been held back by a system of charges designed for manual operation. This changeover, while calling for very little additional equipment, will cut telephone costs by about £2M a year.

The second step, for which the first step is designed to pave the way, concerns full automation. The first exchange to be converted will be Bristol and by the end of this year new equipment will allow subscribers in that city to dial about half the subscribers in the country. It is anticipated that, by the end of 1960, 40 other towns will be similarly equipped and that by 1970 three-quarters of all trunk calls will be dialled by subscribers.

To complete this second step a considerable amount of new and relatively complex equipment will be required; the "brain" of the new system being a machine devised by Post Office research engineers and known as GRACE (Group Routing and Charging Equipment). This equipment will interpret the numbers dialled by subscribers and route calls in the required direction. It will operate switches at intermediate telephone exchanges until the required exchange is reached and will then dial the number on that exchange. Automatic charging will take place and the charges for all calls will be recorded on the subscribers meters. To enable this to be accomplished the present three-minute minimum charge will be abolished and all calls will be charged in units of the local call charge (3d.). This will buy time according to distance. For example, when a Bristol subscriber dials an Edinburgh number his meter will register once when the call starts and again every 13 seconds until the call is finished; but on a call to Gloucester, which is considerably nearer, the meter will register only at 45 second intervals. It will, of course, be necessary to retain some operators to provide such specialized services as personal calls and reverse charge calls etc., but, by substituting machines for people for most of the routine work, great savings will be made and in ten years' time it is anticipated that savings of the order of £15M a year will be effected.

It is pointed out in the White Paper that the greatest problems of automation are not in the technical field, but that the human problems are much greater. Men and women are displaced, less staff is required and career prospects are jeopardized; all of which can cause much anxiety and hardship. However, the Post Office intends that full telephone automation shall be undertaken with the goodwill of the staff and that full regard will be paid to the interests of those who have given long service and to such added responsibilities as may be carried by staff in the future. There will be many such problems to be solved and these will be tackled together by the Post Office and the Trade Unions. The Trade Unions have declared their willingness to co-operate in the introduction of automation and, indeed, their members are already serving on working parties set up to consider its introduction and problems.

While few people will disagree with the statement that the greatest problems of automation are in the field of human relations, and while the Post Office are to be heartily congratulated on the way in which they have approached this problem, it must, nevertheless, be appreciated that the technological advances which have made these developments possible are of no small order. In fact, the research and engineering teams responsible are worthy of the highest praise for, in addition to the fairly complex nature of some of the equipment which has had to be devised, the reliability requirements for equipment used in so vital a public service are extremely stringent and, also, due to the very large capital expenditure involved, the equipment installed must have a projected life of many years.

If all goes according to plan, this will be a considerable technical and economic achievement and should prove a fine example of what can be achieved in the field of automation when the technologists and humanists are united in their efforts.

Automatic Control of Ground Instrumentation During the Launching and Flight of Experimental Guided Missiles

By R. J. Garvey*, B.Sc.(Eng.)

The article describes a programme switch associated with the launching and flight of experimental guided missiles. This unit switches control and recording equipment on and off automatically to a pre-arranged programme set up for the particular missile. The switching is effected by two uni-selectors which control output relays according to time settings on a selector board. There are 30 output channels and each one can be preset to switch remote equipments on and off at any time during a 14 minute automatic cycle.

In launching and flight testing an experimental guided missile, it is necessary to effect a sequence of switching operations on control and recording instruments in the various ground stations on the range and also to the missile before it is launched. The complexity of the missile and ground instrumentation, as shown by Figs. 1 and 2, and the necessity for accurate timing in some cases, requires that the switching be made automatically. The 'programme switch', illustrated by Fig. 3, has been designed to do this. It switches equipment to a pre-arranged sequence or programme set up for each test flight; a typical programme being given by Table 1.

It would be possible to control the switching manually but this introduces a risk of one or more of the personnel involved failing to operate the equipment at the correct time, or at all, particularly during the last few seconds before launching. This could result in the missile being launched in a faulty condition with inadequate ground instrumentation in service. A costly missile would then be wasted since it is normally expended on the single test flight. Manual switching for the more complex missiles has in fact proved impractical and automatic control is regularly used.

The programme switch operates automatically for a maximum period of 10min preceding the launching time followed by a 4min run to cover the flight of the missile. Each output channel can be quickly and independently preset to switch remote equipments on and off at any time during the 14min cycle. Thirty output channels are available with three outlets on each, permitting the control of a maximum 90 remote equipments. An indicator fitted to the programme switch counts off each second with respect to zero time, i.e. the time at which the round is launched. The officer controlling the shoot observes this and announces the time at suitable intervals over the range communication system. This officer is in close communication with all range stations and if necessary can halt the automatic operation of the programme switch.

The flight tests and whole range instrumentation depend on the correct functioning of the programme switch; due care has therefore been taken in designing the equipment to ensure that it operates reliably. All parts of the equipment are readily accessible for servicing and a packaged construction permits any vital faults to be quickly eliminated by plugging in replacement units.

The description here refers to equipment designed to meet a specific requirement. The basic circuit is, however, adaptable and could be used for the control of industrial processes. Another adaption would be suitable for use in

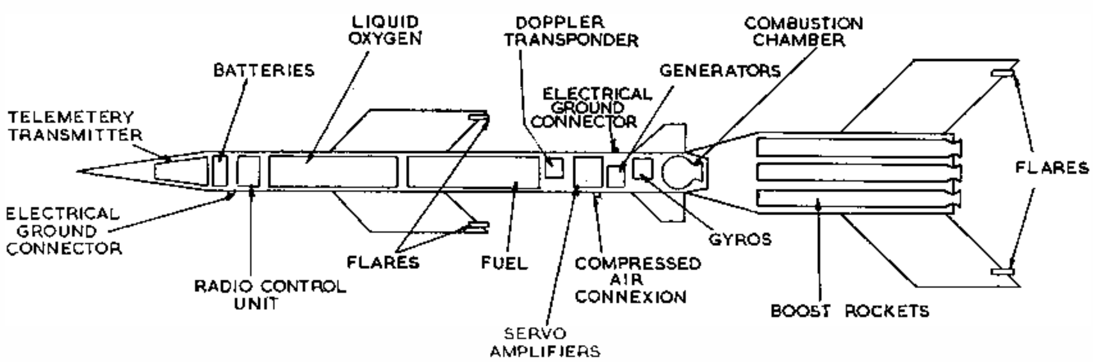


Fig. 1. Typical guided missile equipped with control unit and telemetry transmitter for test flight

TABLE 1
Typical Firing Programme

TIME	OPERATION
−5m	Electrical ground supplies switched to missile.
−4m 45sec	Compressed air supplied to gyros and servos in missile.
−4m 30sec	Telemetry aerial spinner switched on.
—	Operation of telemetry and control systems checked.
−2m	Missile switched on to internal supplies.
—	Operation on internal supplies checked.
−30sec	Fuel tanks primed.
−20sec	Electrical connexions to missile ejected.
−15sec	Compressed air line to missile ejected.
−4sec	Tail flares ignited.
−2sec	Cine-theodolites switched on.
−1sec	Launcher camera switched on.
−1sec	Radar recorder switched on.
−1sec	Telemetry recorders switched on.
−1sec	Doppler recorder switched on.
−1sec	Priming signal switched to central timer.
zero	Boost motor fired.
+1sec	Launcher camera switched off.
+30sec	Target cameras switched on.
+40sec	All circuits switched off.
+4m	Programme switch reset.

* *The Royal Aircraft Establishment.*

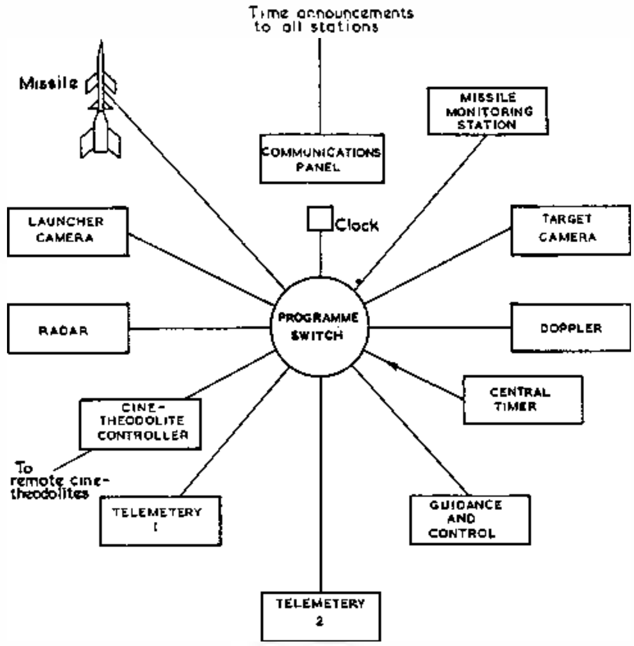


Fig. 2. Programme switch and range instrumentation

the field to fire operational missiles. Possible developments of the circuit design in this respect are discussed very briefly below.

General Description

A 1p/s operating signal is supplied to the programme switch from the range 'central timer'. This timer[1] consists of a 10kc/s oscillator and frequency dividing circuits which supply timing and reference signals to the various measuring and recording instruments used on the range; it functions as a separate equipment. The programme switch could be operated independently but it is better to synchronize all automatic switching operations with range time.

Referring to Fig. 4: the 1p/s operating signal is fed via a pulse amplifier to Post Office type uniselectors which switch voltages to a special selector board. Output relays control remote equipment in the missile and ground stations according to the time settings on each channel of the selector board. The uniselectors commence stepping in synchronism with the 1p/s signal when the operator presses the start button, run through a 14min cycle and automatically reset for another run. The operator can, however, interrupt the run for any length of time by operating the stop and start switches or, if necessary, reset the uniselectors before the cycle is complete.

Fig. 5 illustrates the method of operation in more detail. The selector board shown consists simply of a lattice of horizontal and vertical conductor bars arranged so that connexions can be made between them by inserting studs at the intersections. Uniselector *A* steps in synchronism with the 1p/s operating signal and switches a positive supply successively to 60 of the horizontal bars. Uniselector *B*, which is stepped once per minute by *A*, switches an earth in succession to the remaining 14 horizontal bars.

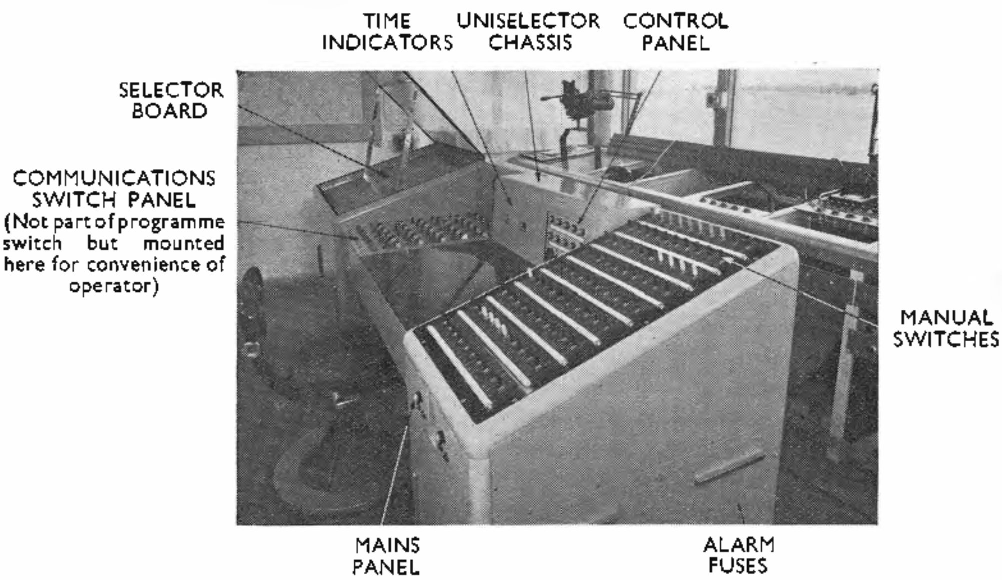


Fig. 3. Programme switch installed in central control room

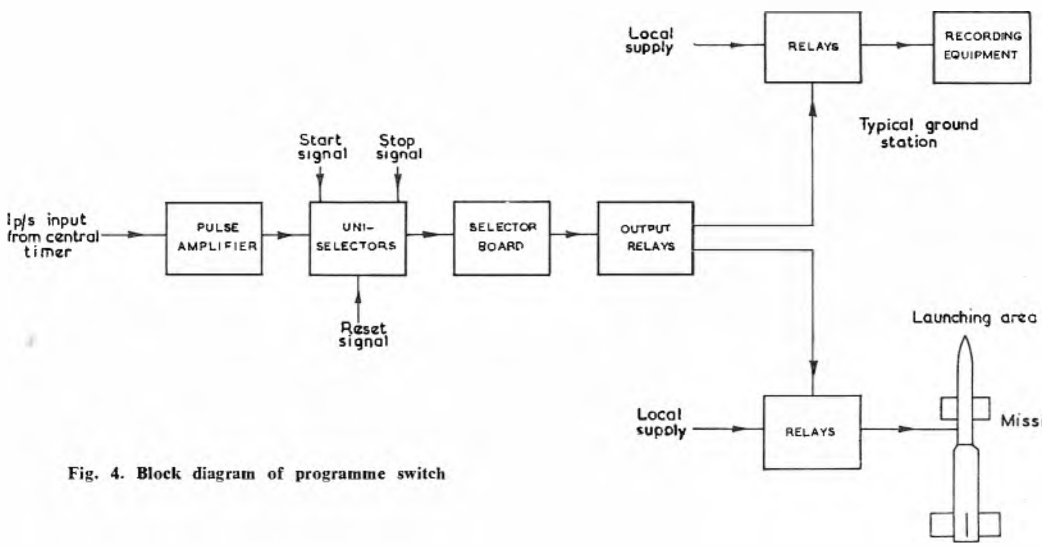


Fig. 4. Block diagram of programme switch

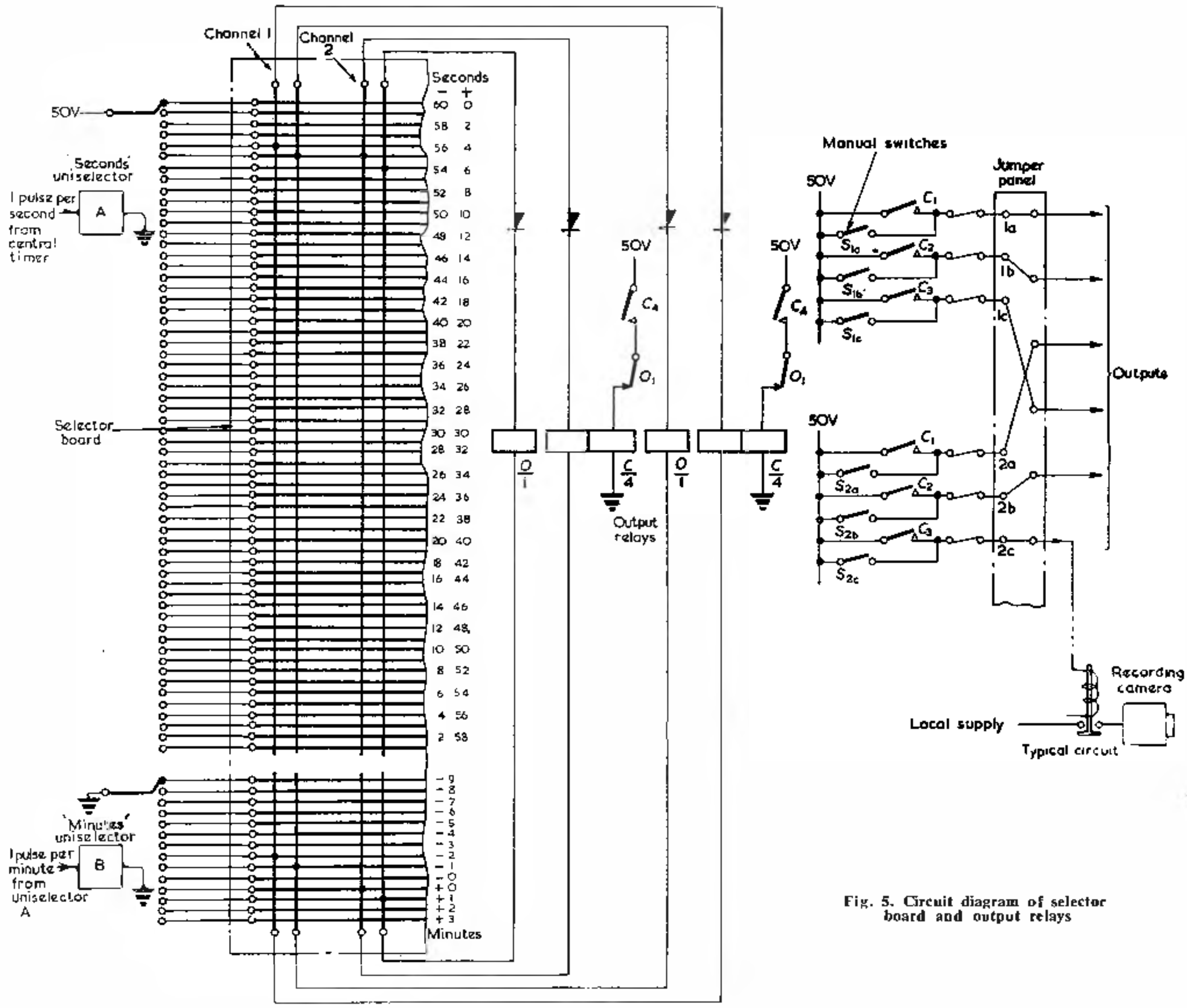
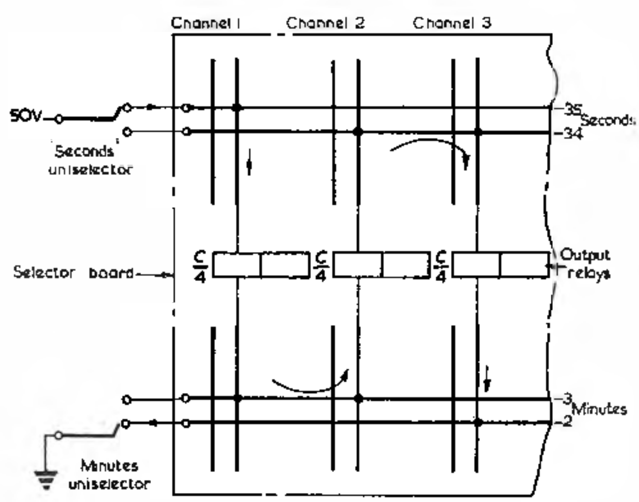


Fig. 5. Circuit diagram of selector board and output relays

Output relays connected to the vertical bars are switched on and off during the automatic run according to the positions of the studs inserted at the intersection of the conductor bars. As an example, the setting shown for channel 1 will cause relay C to switch on via the uniselectors at 2min 56sec before zero time; a second winding then holds C on until it is unlatched by relay O at 1min 55sec before zero time. The setting shown for channel 2 will result in the output relay switching on at 5sec and off at 1min 6sec after zero time. The rectifiers shown in the relay circuits block reverse currents that would occur with certain settings on the selector board, as shown by Fig. 6; these reverse currents would result in mis-timed operation of the relays.

Fig. 6. Fault current in relay circuit illustrating need for rectifiers
(All three relays will operate at the wrong time)



The uniselectors step automatically for a maximum period of 10min preceding the nominal zero time, i.e. the time at which the round is normally launched, followed by a 4min run to cover the flight of the missile. Time settings on the selector board are made with respect to zero time; a setting before zero having, conventionally, a negative value. It will be seen that a double annotation for the "Seconds", on the selector board, is necessary to suit both negative and positive settings.

Each output relay controls three contacts which switch pilot voltages to remote relays and contactors. Thirty output channels with three outlets on each are thus available so that a maximum of 90 remote equipments can be controlled. Manually operated switches are connected in parallel with the relay contacts, permitting the remote circuits to be controlled manually for test purposes. These

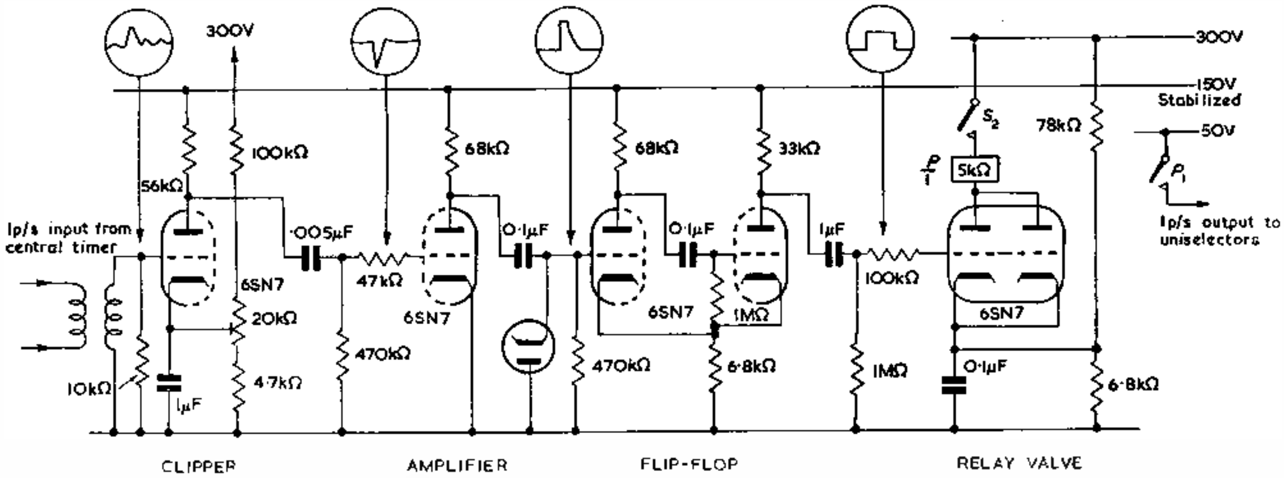


Fig. 7. Pulse amplifier

switches can be seen on the right-hand side of the console shown by Fig. 3. Ivorine labels are mounted beneath the switches so that they can be annotated according to the particular remote equipments which they control. The jumper panel shown on Fig. 5 permits remote circuits to be readily connected to any one of the 90 outlets so that maximum use can be made of the limited number of independently timed channels.

Pulse Amplifier

The function of the pulse amplifier is to amplify and widen the 1p/s pulses from the range central timer so that they operate a relay which steps the uniselectors. The pulses are transmitted from the remote timer over a long cable terminated at each end in balanced line transformers. A balanced line transmission is employed to eliminate the 'pick up' of extraneous signals that would cause mis-stepping of the uniselectors. The receiving end transformer is incorporated in the pulse amplifier.

The circuit, which is shown by Fig. 7, consists of a triode biased beyond cut-off to clip any small amplitude noise from the signal, followed by an amplifier and pulse widening flip-flop. The output of the flip-flop, which is a pulse of about 60msec duration, is fed to two triodes connected in parallel, with a relay in the anode circuit. This relay switches a voltage to the drive magnet of one of the uniselectors. The anode circuit of the relay valve is normally open, being closed by contact S_2 when the start switch is pressed to initiate the automatic operation of the programme switch.

The pulse amplifier, being a vital part of the equipment, is a self contained, plug-in unit with a built-in power pack. It can be quickly replaced with a spare in the event of a fault.

Uniselector Unit and Control Circuits

The uniselector chassis is illustrated by Fig. 8. Like the pulse amplifier it is a plug-in unit that can be replaced in the event of a fault.

Referring to Fig. 9; uniselector A is a special one having 30 outlets on each switch bank instead of the usual 25 ways. The wipers are staggered and the banks connected in pairs to give the required 60 outlets per revolution. It is stepped once per second by the pulsing of the relay P in the pulse amplifier. This operation commences when the control relays S and R are energized via the start switch and held on by the latching circuit through R_1 and S_1. Relay S switches on the relay valve in the pulse amplifier and R de-energizes the uniselector homing circuits. Operation of the stop switch releases S so that the uniselectors cease stepping, while operation of the reset switch releases both S and R so that the uniselectors halt and home to the start position ready for another run. Uniselector B is a standard 25 way unit, only 14 ways being required to complete the switching cycle. It is stepped by

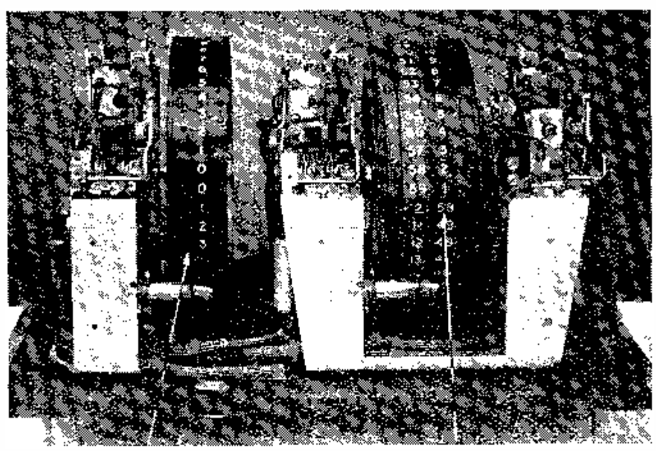


Fig. 8. Uniselector chassis

relay M which is energized once every 60sec by uniselector A.

The output connexions to the conductor bars of the selector board are made from the switching banks A_3, A_4 and B_2. It will be seen that the positive supply to these banks is fed via relay contacts R_2 and D_1. The contact R_2 isolates the outputs from the supply when the reset relay R is released; this ensures that there are no output voltages fed to the selector board as the uniselectors are homing. Contact D_1 protects the wipers by interrupting the supply while they are moving; it is operated by a slow release relay D which is pulsed by the signal from the pulse amplifier.

The uniselectors home to the starting position, when R is released, on the switching banks A_1, A_2 and B_1. Uniselector B homes according to the setting of a shorting switch. This presets the uniselector so that it starts from 1 to 10min before zero time. At the end of a normal uniselector run, i.e. at $+4$min, an earth is switched via B_1 to trip the reset relay so that the uniselectors are automatically reset ready for the next run.

Uniselector banks A_5, A_6 and B_3 operate relay K. This relay keys the output valve of a 1kc/s oscillator so that warning 'peeps' are transmitted over the range inter-

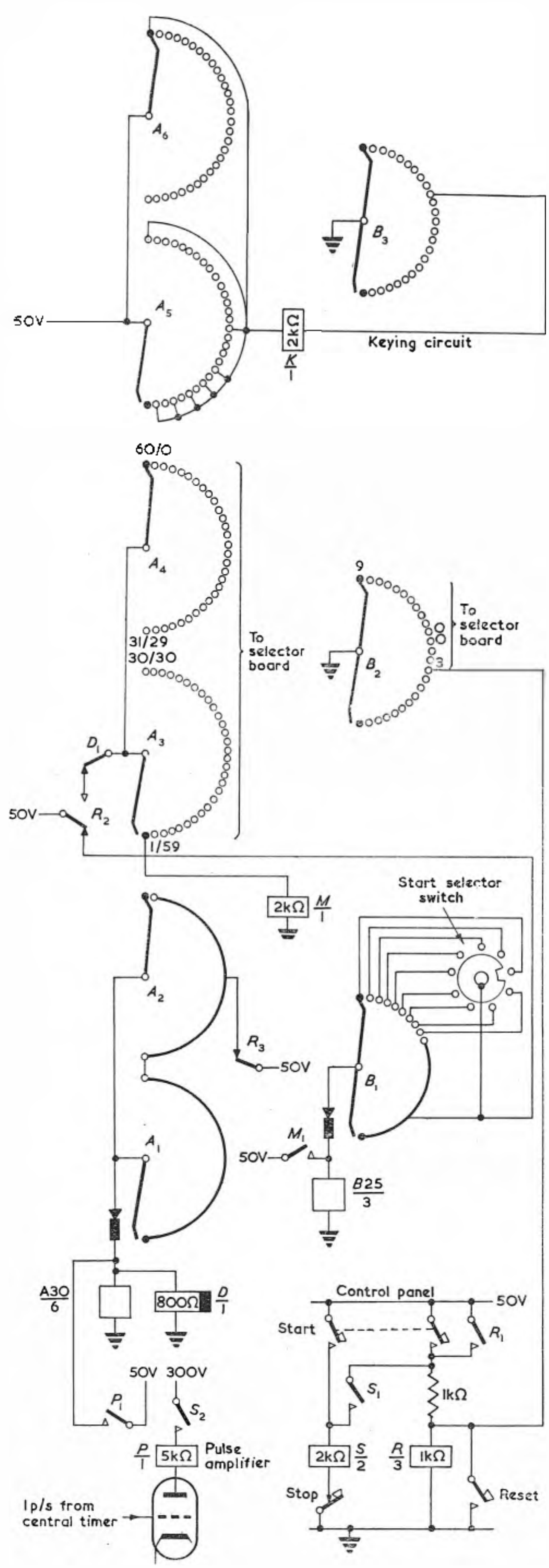


Fig. 9. Circuit diagram of uniselector chassis

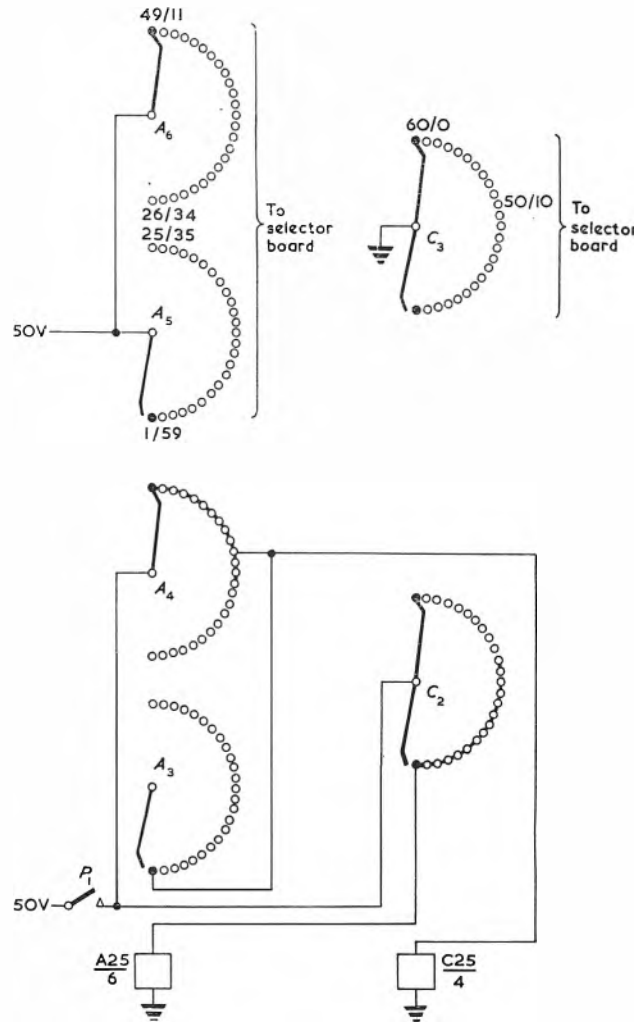


Fig. 10. Method of adapting two 25 way uniselectors to give a 60 way operation

communication system during the shoot. The time at which the 'peeps' are transmitted depends on the connexions to the uniselector banks; with the arrangement shown they are transmitted at -60, -30, -15, -10, -8, -6, -4 and -2sec before zero time. The outlets from the switch banks are wired to a small jumper board so that the time connexions can be easily altered if required.

Reference has been made to a non-standard, 30 way uniselector. While this is made it was not available when the programme switch was built and a circuit was improvised which permits the use of 25 way uniselectors. The circuits employing the 30 way uniselector has, however, been described in detail since it is the best arrangement and illustrates the principle of operation most clearly. The method of using 25 way uniselectors to give the required 60 way operation is shown by Fig. 10. Two cross connected uniselectors are used; one providing 10 and the other 50 outlets. Uniselector *C* steps through the first 10 positions and completes the circuit to drive magnet *A*. Uniselector *A* then steps through the remaining 50 positions while *C* returns to its original position ready for the next 60sec cycle.

Indicator drums, engraved in minutes and seconds, are fitted to the uniselector spindles as illustrated by Fig. 8. They are viewed through a suitable aperture and count off the minutes and seconds of the operation with respect to zero time. The drums mounted on the two 'seconds' uniselectors are concentric; the inner drum being viewed through an aperture cut in the outer one. A single 'seconds' drum would suffice when using a 30 way uniselector. Negative time, i.e. time before zero, is engraved on a black and positive time on a red background. A double set of figures is engraved on the drums indicating seconds, corresponding to negative and positive time. The method of indicating time directly off the uniselectors ensures that the indication is always in phase with the switching operations; a false indication and announcement of time over the range intercommunication system would, of course, completely disorganize the shoot.

A requirement has arisen, since installing the programme switch, for time indication at a number of remote positions; various methods of meeting this are being considered. The problem is to adopt a simple method which will ensure that the indications are always in phase with the stepping of the programme switch. One solution is to wire a bank on each uniselector to function as a Desynn or Magslip transmitter, with two receivers at each remote position to indicate minutes and seconds.

Selector Board

This is illustrated by Fig. 11. It consists of a lattice of Inconel bars mounted on an insulating board. The bars are drilled and tapped so that that they can be connected at the intersections by screwing in 4 B.A. studs as shown. The horizontal bars on the rear of the board, which cannot be seen in the illustration, are in two groups; 60 of them are connected to the uniselectors stepping at second intervals and 14 to the uniselector which steps each minute. The bars on the front of the board are wired to the 30 output channels. The board is annotated in minutes and seconds according to the times the uniselectors switch to the horizontal bars. The times are repeated in four columns since it would be difficult to align studs against a single column across the whole width of the board.

Alarm Circuits

Failure or faulty operation of equipment in the missile and ground stations could result in a misfire or in the missile being launched with inadequate ground instrumentation in service. Alarm facilities are, therefore, provided so that during a shoot an operator at any of the remote stations can warn the control room if he is not satisfied with the operation of his equipment. The operator concerned presses a stop switch which sounds a buzzer and lights a warning lamp on the programme switch. The automatic operation is then interrupted or cancelled at the discretion of the programme switch operator.

The outlets of the programme switch are individually fused, primarily to protect the contacts of the output relays An automatic run with one or more of these fuses blown could result in a misfire or premature launching. It is therefore, important that a warning be given should any of these fuses blow either before or during a shoot. Alarm fuses, as used by the G.P.O., are therefore fitted. Each one consists of a contact spring anchored to a fuse wire. Rupture of the fuse wire releases the spring so that it contacts a common bus bar and energizes a warning buzzer.

Operation

The function of the programme switch is illustrated now by describing the procedure for an actual shoot.

A programme to suit the particular missile is set up on the selector board and any necessary alterations made to the jumper panel connexions. A typical programme is given by Table 1. The connexions to the remote stations

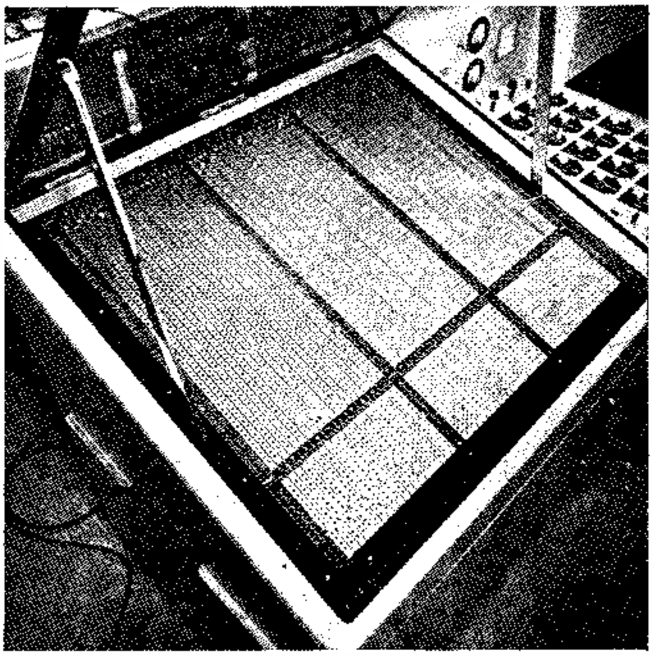


Fig. 11. Selector board

are then proved by operating the manual switches and checking that each equipment functions correctly and is connected to the right channel. Dangerous circuits such as those associated with the ignition of the rocket motor are checked as far as possible but voltage supplies are not completed to the equipment until all personnel are clear of the launching area prior to firing the missile. Safety links are then inserted and continuity checked with an ohmmeter.

The start selector switch is set to the time at which the automatic operation is to commence. For example, if the earliest setting on the selector board is -4min 30sec, the start selector will be set to -5min. The programme switch is now ready and the control officer stands by while final tests are made on the missile. If the programme is a complex one the firing is preceded by a dummy run. This duplicates the live run, as far as possible, to check the instrumentation and ensure that personnel are familiar with the procedure and ready for firing.

When all is ready, the launching area is cleared, the safety links inserted and final continuity tests made. The control officer alerts range personnel over the intercommunication system and presses the start switch. The operation then proceeds automatically while time is announced at suitable intervals. Apart from the time announcements it is normal to maintain silence on the range intercommunication system so that it is free for emergency use. The sequence is interrupted in response to an alarm signal, or for any other reason, at the discretion of the control officer; the sequence continues when the start switch is pressed. If a failure in the instrumentation or the missile require the shoot to be cancelled, the reset switch is pressed. All remote circuits are then de-energized and the programme switch is ready for another run. The sequence normally continues to cover the flight of the missile and the uniselectors reset automatically.

at any time during a 24h cycle. The hour and minute selector switches of each channel are concentrically mounted with the switch pointers displayed against a clock face. A feature of the basic design which applies to this variation is the facility for operating a large number of independently timed channels off the two uniselectors. The number is limited only by the current rating of the uniselector wipers and outlets which supply the output relays.

A process control system using uniselectors has been described by Watson and Wiggs[2]. This system is of independent design to that described here; it is a good example of the general application of automatic switching techniques.

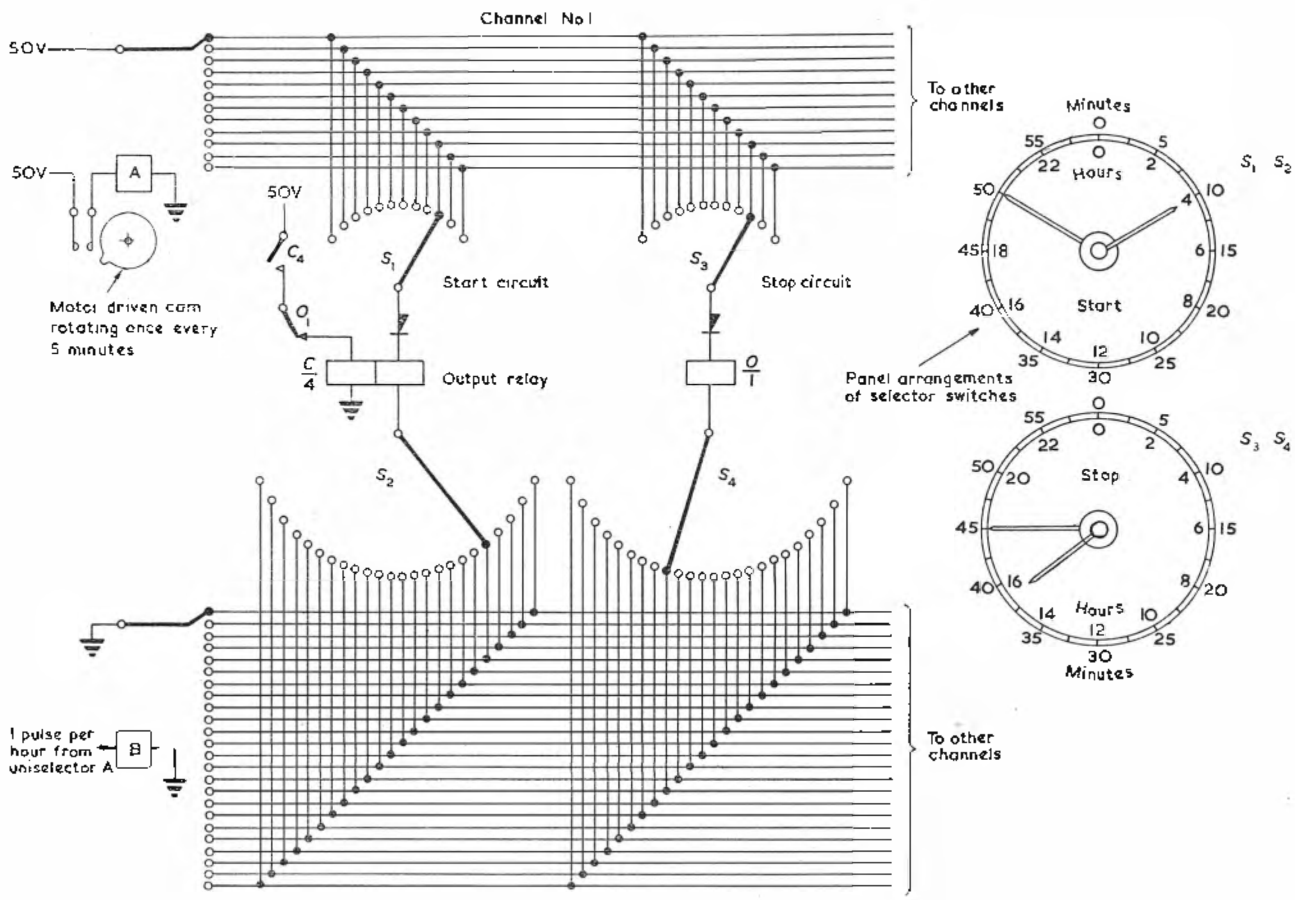


Fig. 12. Variation in basic circuit design

The time at which certain instruments should be switched off cannot be foreseen exactly since it depends on the behaviour of the missile during its flight. High speed cameras and recorders switched on by the programme switch are therefore switched off manually by local operators. This saves film which is used at the rate of 8ft/sec in some instruments.

Possible Developments

The equipment described was designed to meet specific requirements. However, the basic circuit could be readily adapted for other applications. An adaption, for example, that could be used for industrial process control is shown by Fig. 12. Here uniselector A is stepped every 5min and B every hour. Uniselector A is connected to recycle after every 12, and B after every 24, steps. Programmes are set up on a bank of multi-way switches which replace the selector board in this case. This arrangement permits equipments to be switched on and off, to the nearest 5min,

The basic circuit could be adapted for firing operational missiles. The arrangement would be simpler in this case since the equipment would be designed to suit the limited requirements of a particular type of missile. Switches could be used to set up the firing programme. Alternatively a selector board would permit the use of ready-made programmes consisting of pre-arranged patterns of selector studs mounted on plug-in panels.

Acknowledgments

A prototype of the programme switch was originally designed in the Guided Weapons Department of the Royal Aircraft Establishment. The equipment described, however, is the result of subsequent development and engineering by the Plessey Manufacturing Co. Ltd.

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Junction Transistor Sawtooth Waveform Generators

By K. P. Padmanabhan Nambiar, M.Sc., D.I.C.

The bootstrap and Miller integrators are two well-known thermionic valve circuits for generating linear sweeps or sawtooth waveforms. In this article these circuits and their equivalents using junction transistors are discussed. It is shown that sufficient current swing can be obtained for applications such as electromagnetic scanning of miniature television camera tubes using currently available low power transistors.

TWO well-known circuits in valve practice for generating linear sweeps or sawtooth waveforms are the Miller and the bootstrap integrators. The two forms of circuit are very similar in operation and differ mainly in the choice of earthing points. It is found that junction transistors may be successfully employed in either of the two forms of circuit to generate linear sweeps, the percentage non-linearity in either case being less than 1 per cent in certain circuits[1]. In the following, circuits of the bootstrap type, recurrently generating sawtooth voltages and currents, are discussed; it is shown that the current and voltage swings are limited mainly by the dissipation rating of the transistor. Sufficient current swing for applications such as electromagnetic scanning of miniature television camera tubes (for example, vidicon photo-conductive tube), may be obtained with currently available low power transistors of the OC72 type.

Generation of Linear Sweeps

The action of the bootstrap circuit using either a junction transistor or a valve may be seen by reference to Fig. 1. In Fig. 1(a), X_1 is an emitter-follower similar to the cathode-follower V_1 of Fig. 1(b), and exhibits a voltage gain, A, close to unity, from the base to the emitter load R_e. Assume that the switch S is closed in the steady state and then suddenly opened. The diode D_1 conducts and a current I flows into the capacitor C (which is initially at earth potential) building up a voltage V_i across it. For a pnp transistor the supply voltage is actually negative so that I flowing into C, and the potential V_i built up across C, are also negative quantities. The action of the valve circuit follows very closely to the bootstrap circuit using an npn transistor where the polarities of voltages and currents are all positive. Owing to the emitter-follower action, an output voltage $V_o = AV_i$, appears at the emitter of X_1; if the feedback capacitor C_1 is so large that it offers negligible impedance to the potential changes at the emitter of X_1, V_o will appear at the anode of D_1, cutting off the diode. Since the feedback capacitor C_1 is charged to the supply potential V_c in the steady state this acts as a 'floating battery' supplying the current which flows into the charging capacitor C, while at the same time maintaining the feedback of V_o to the junction of the resistor R and the diode D_1. These two functions performed by the capacitor C_1 enable a continuous supply of charging current I to flow. This current would be a constant and equal to V_c/R, if A were unity, R_i infinite and the output resistance of the emitter-follower zero; the voltage built up across the capacitor C would then be a perfectly linear sweep.

In practice the voltage V_i developed across the capacitor, and consequently the output voltage V_o, is non-linear owing to the voltage gain A being less than unity and the input resistance R_i of the emitter-follower having a finite though large value. Both these factors alter the magnitude of the current I through C. For any value of V_i corresponding to time considered, a value of A less than unity reduces the magnitude of the current flowing into the capacitor C from V_c/R to $[(V_c/R) - V_i(1-A)/R]$. V_i may be assumed to be an approximately linear sweep if the change in the current I is relatively small so that

$$V_i = (V_c t/CR) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

The decrease in current through C at any time owing to non-unity gain is then given by

$$\frac{V_c t}{CR} \quad \frac{(1-A)}{R}$$

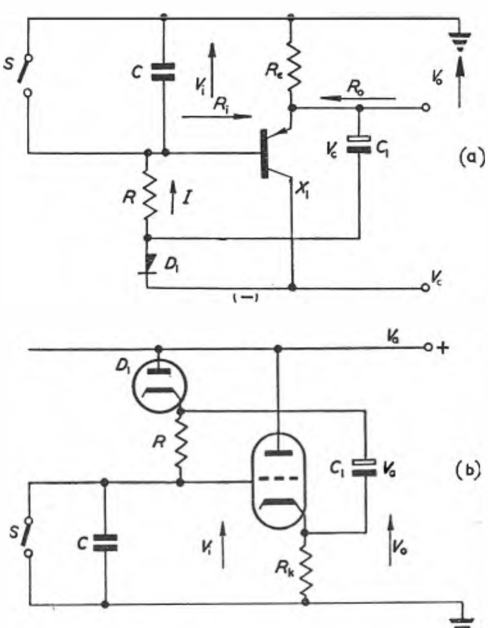

Fig. 1. Bootstrap linear sweep circuits
(a) Circuit using junction transistor
(b) Circuit using triode valve

The magnitude of the current flowing into C is also decreased from the initial value of V_c/R because of diversion of current into the base of the transistor, this decrease being dependent on the input resistance R_i. Thus again assuming V_i to be a linear sweep as above the decrease in current due to R_i is given by

$$V_i/R_i = (V_c t/CRR_i)$$

The total decrease in current ΔI through the capacitor C in relation to time is therefore

$$V_c t/CR\left[\frac{(1-A)}{R} + (1/R_i)\right]$$

The reduction in voltage across C due to this current is given by

$$V_d = (1/C)\int \Delta I \,.\, dt = \frac{V_c}{2\,.\,CR\,.\,CR}\left[(1-A) + (R/R_i)\right] t^2 \,\ldots\, (2)$$

The degree of non-linearity may be expressed as the deviation of the instantaneous output voltage of the actual sweep from the instantaneous voltage of an assumed straight line which closely represents the actual sweep. This deviation is expressed as a percentage of the actual sweep and is termed 'displacement error' or percentage non-linearity, denoted here by ϵ. The straight line is assumed to coincide with the beginning of the sweep and have a slope equal to the initial slope of the actual sweep. If the switch S of Fig. 1 is closed at a time duration $t = t_s$, the voltage across the capacitor C will consist of a linear sweep V_s equal to $V_c t_s/CR$, and a distortion term due to V_d, given by equation (2), where t_s is substituted for t. The percentage non-linearity is then given by the ratio of V_d and V_s so that

$$\epsilon = (V_d/V_s) \times 100 = (50t_s/CR)(\delta + k) \text{ per cent} \quad \ldots\ldots (3)$$

where $\delta = 1 - A$ and $k = (R/R_i)$.

In terms of the small signal a.c. parameters of the junction transistor

$$A = \frac{R_e}{r_e + R_e + r_b(1 + A_i)} \quad \ldots\ldots\ldots\ldots (4)$$

where

$$A_i \simeq \frac{\alpha r_c}{R_e + r_c(1 - \alpha)} = \frac{\alpha}{(1 - \alpha) + R_e/r_c} \quad \ldots\ldots (5)$$

so that

$$\delta = (1 - A) \simeq \frac{r_e + r_b/(1 + A_i)}{R_e} \quad \ldots\ldots\ldots (6)$$

and

$$R_i = r_b + R_e(1 + A_i) \quad \ldots\ldots\ldots\ldots\ldots (7)$$

r_e, r_b and r_c are the emitter, base and collector resistances, α, the short-circuit current amplification factor of the grounded base stage, and A_i the circuit current gain from base to collector of the transistor.

On closing of the switch S the capacitor C discharges rapidly and the potential across it falls to zero. The rapidity of the discharge is governed by the magnitude of the current that may be supplied by the switch S—which, in the practical circuits to be described, is also a transistor.

Linearization of the 'Run-Down' by Compensating Feedback

In the triode valve bootstrap circuit of Fig. 1(b) the non-linearity of the sweep is almost entirely the result of non-unity voltage gain A; the value of k of equation (3) approximates to zero since the input impedance of the valve is of the order of many megohms. The maximum voltage gain of a triode cathode-follower is defined by the amplification factor μ of the valve so that

$$A \leqslant (\mu/(\mu + 1))$$

In the junction transistor emitter-follower circuit the voltage gain $A = (A_t/(1 + A_t))$, where A_t is the voltage gain of a grounded emitter stage with R_e as its collector load. This voltage gain A_t is given by $A_i \, . \, (R_e/R_{in})$, where R_{in} is the input resistance of the grounded emitter stage. In practice the magnitude of the amplification factor μ of the triode valve is only of the order of the current gain A_i of the transistor so that by increasing the ratio of R_e/R_{in} within limits by appropriate choice of R_e the voltage gain A may be made much closer to unity in the case of the emitter-follower than for the triode cathode-follower.

For example, the current gain A_i of a transistor may typically be equal to 40 with $R_e = 10\text{k}\Omega$ and $R_{in} = 1\text{k}\Omega$.

Then $A_t = A_i \, . \, (R_e/R_{in}) = 400$

so that $A = (A_t/(1 + A_t)) = 400/401 = 0{\cdot}997$ for the emitter follower.

Assuming a value for μ, equal to that for A_i, in the valve case, the voltage gain A of the triode cathode-follower will be equal to or less than 40/41 so that $A \leqslant 0{\cdot}976$ for the cathode-follower.

Thus the approximation of A to unity is much more valid for the emitter-follower than for its valve counterpart.

However, due to the resistive path between the base and the emitter and the finite value of the current gain A_i the approximation of R_i to infinity is not valid in the emitter-follower and this is mainly responsible for the non-linearity in the generated sweeps.

In valve circuits, the sweep non-linearity resulting from departure of A from unity may be considerably reduced by a resistive feedback method of compensation[2]. Fortunately, this compensation technique is valid[1] regardless of the cause of non-linearity provided it be largely attributed to the t^2 term given in equation (2), so that effective compensation for the finite value of R_i and for the non-unity

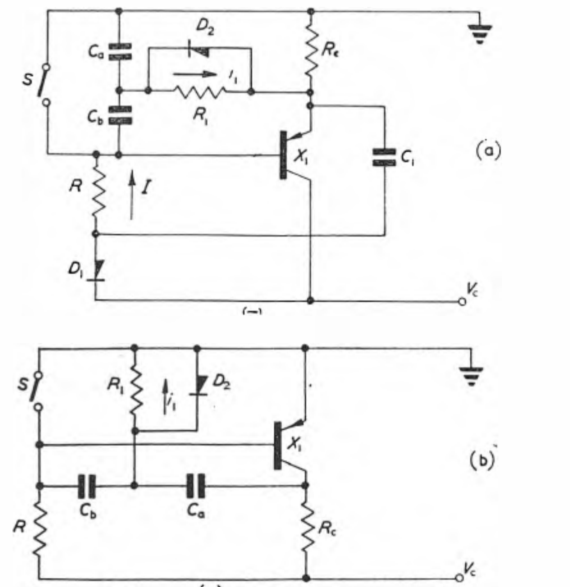

Fig. 2. Resistance compensation for the improvement of sweep linearity
(a) Bootstrap
(b) Miller

gain A of the junction transistor circuit is feasible. The method of compensation is illustrated in Fig. 2(a). Here the charging capacitor C is split into two, C_a and C_b and a resistor R_1 connected between the junction of the capacitor and the emitter of X_1. The potential difference across R_1 drives a current i_1 through it; assuming this is linear with time t, a 'compensating' potential proportional to t^2 will be developed at the base of X_1 as a result of integration of i_1; this integration is performed mainly by C_a because of the relatively high resistance paths (i.e. via R or R_i) associated with the charging of C_b. This compensating potential is in opposite phase to the initial distortion term of equation (2) and the magnitude may be made equal to the initial distortion by appropriate choice of R_1.

Assuming $C_a = C_b$, R_1 is given by the relation

$$R_1 = \frac{R}{4(1 - A + (R/R_i))}$$

If $(1 - A) \ll R/R_i$ as is usual in emitter-follower circuits

$$R_1 = (R_i/4) \simeq \frac{R_e(1 + A_i)}{4}$$

In the Miller* version of the sweep circuit the compensating resistor R_1 may be connected between the junction of

* *This form of compensation for Miller sweep circuits does not appear to have been mentioned in any literature; by employing this compensation, however, the linearity of Miller circuits using valves or transistors may be considerably improved.*

the capacitors and earth as shown in Fig. 2(b). The value of R_1 is then given by $R_c(1 + A_1)/4$, where R_c is the collector load resistance.

In practice a potentiometer may be employed for R_1 and used as a linearity control. The diode D_2 is non-conducting throughout the sweep and helps to reset the potential at the junction of C_a and C_b at the end of the flyback.

RECHARGE OF C_1

In all bootstrap circuits C_1 acts as a 'floating battery' during the sweep, supplying current for integration by the charging capacitor C and consequently loses some of its charge in the process. In valve circuits the charge lost is considerably smaller owing to the large value that can be employed for R (since $R_i \to \infty$) thereby reducing the charging current supplied by C_1. In the transistor circuit, however, for good linearity the value of R should be relatively low and much smaller than the finite value of R_1 so that much larger currents flow during the sweep. Consequently the charge lost by C_1 may become appreciable and C_1

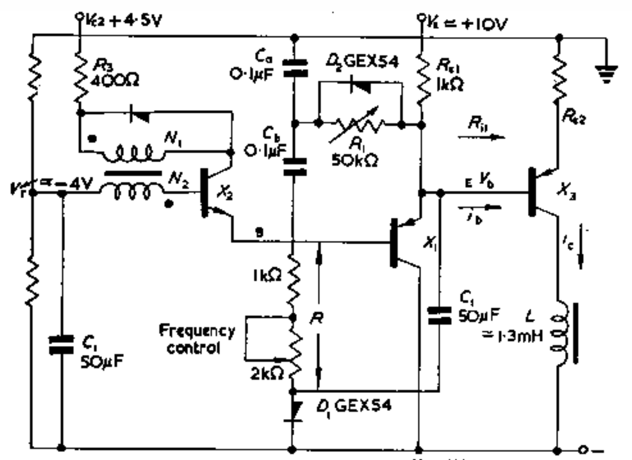

Fig. 3. Sawtooth waveform generator using pnp and one npn transistors.

The value of R_{e2} defines the maximum current swing through L, X_1 and X_3: Type OC72 (cooling fins fixed to the metal chassis or appropriate heat sink) X_2: npn GD2517. R_{e2} as required but not less than 27Ω. Blocking oscillator transformer: core: potcore type LA1. N_1 — 250 turns, N_2 — 125 „ } 40 s.w.g.

Frequency of operation — 10kc/s

should be recharged before commencing the subsequent sweep. This may be done by a current drawn via the output of the emitter-follower or via R_e from a positive supply voltage V_c or by the addition of a transistor switch at the emitter of X_1, the switch being in the cut-off state during the 'run-down' and 'on' state during the flyback; in the 'on' state large currents may be made available for the recharge of C_1.

As stated earlier the base of the transistor X_1 returns to the steady state (earth potential) by the current supplied by the switch S.

Sawtooth Voltage and Current Generator Circuits

Fig. 3 shows a form of sawtooth generator-employing junction transistors. Here X_1 acts as the emitter-follower and X_2 as the switch S of Fig. 1. X_2 is an npn transistor and is initially held cut off by a negative reference potential V_r at its base; the bootstrap action and linear 'rundown' occur as discussed previously until the potential at the base of X_1 (i.e. V_i) approaches V_r. When $V_i \simeq V_r$ emitter current begins to flow in X_2, the current increasing regeneratively owing to the blocking oscillator action until limited by the resistor R_3. X_2 is thus in the 'on' state supplying large currents for the discharge of the charging capacitor C_a and C_b so that the base of X_1 recovers to earth potential. At the same time the emitter potential of X_1 rises and recharge of C_1 is also completed aided by the current supplied via R_e. At the end of the flyback X_2 is again cut off so that the 'run-down' process is repeated. Thus a recurrent sawtooth voltage waveform appears across the emitter load R_{e1} of X_1.

X_3 stage in Fig. 3 converts the voltage sweep into current sweep, as discussed in more detail later, the linear rise in current being obtained through the coil L and the peak swing defined by the value of R_{e2} in the emitter of X_3. Using type OC72 transistors for X_1 and X_3 (with their cooling fins connected to a heat sink) a current swing up to 150mA has been obtained through the coil as R_{e2} is decreased to 27Ω, the run-down and flyback periods being of the order of 94 and 6μsec respectively, corresponding to a television line scanning frequency of approximately 10kc/s. The linearity of the scan is adjusted by the potentiometer R_1. Non-linearity present is mainly due to insufficient compensation for the variations of the current amplification factor α of X_3 and in parameters of X_1 during the sweep. The parameter α falls off with increasing emitter current, $(1 - \alpha)$ being roughly proportional to the emitter current at high currents so that the effective a.c.

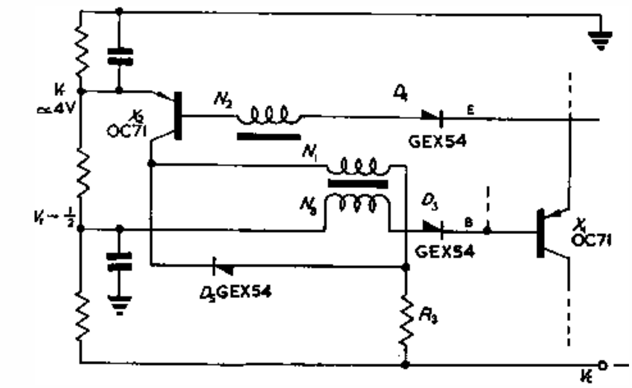

Fig. 4. Switching arrangement using pnp transistor: type OC71

emitter load resistance of X_1 varies correspondingly resulting in variations of R_1 and A of X_1. Distortion limited to about 1 per cent or less is estimated to be possible in all the practical circuits discussed.

Fig. 4 shows a somewhat different switching arrangement for a circuit employing a pnp transistor for X_2. Here the emitter of X_2 is initially at the negative reference potential V_r so that X_2 is cut off, the current I flows through R into the integrating capacitor C_a and C_b of the bootstrap circuit initiating the linear 'run-down'. When the sweep amplitude V_s appearing at E (Fig. 4) equals V_r, emitter current flows in X_2 and the blocking oscillator fires. The diode D_3 is reverse biased throughout the sweep by returning one end of the winding N_3 to a potential negative to V_r by at least half a volt; this biasing prevents the conduction of D_3 before the firing of the blocking oscillator comparator. During the pulse large collector current flows in X_2, a portion of this current being transferred for the discharge of the capacitors connected at the point B via the winding N_3 of the pulse transformer, the diode D_3 conducting in the process. Apart from the presence of an additional winding for the pulse transformer and two extra diodes D_3 and D_4 the switching arrangement of Figs. 3 and 4 are very similar. D_4 prevents any negative 'spike' appearing at the emitter of X_1 when the regenerative action of X_2 ceases.

In the above circuits synchronization may be achieved by applying the synchronizing pulses at the base of X_2. There are applications, however, where a recurrently driven circuit is more suitable than an oscillatory one. In that case, the transformer may be eliminated and X_2 functions as an externally driven switch as shown in Fig. 5. Here X_2 is held cut off by a positive voltage V_b at the base so that the sweep action in X_1 proceeds as already discussed. When the period of the run-down equals the required period of the active scan a negative going pulse is applied at the base of X_2; large collector current flows in X_2 discharging the capacitors C_a and C_b connected at the point B, to earth potential. On removing the pulse, the base of X_2 returns to the positive potential set by V_b and X_2 is cut off initiating the subsequent run-down.

It should be noted that, in all the circuits discussed above, the sweep amplitude at B cannot exceed $V_c R_i/(R + R_i)$, the potential drop across R_i in the absence of C.

Conversion of Voltage into Current Sweeps

Normally in valve circuits sawtooth voltages of the order of 3 to 5V applied to the grid will be quite inadequate to produce any appreciable linear current rise in an inductance placed in the anode of the valve owing to the non-linearity of the transfer characteristics, resulting in variation of g_m and also due to the approximation to a constant

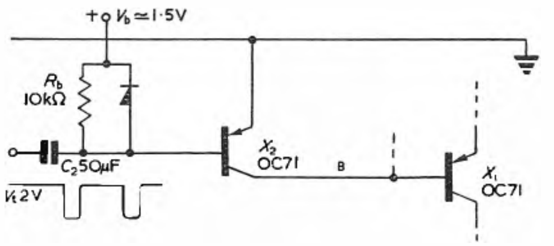

Fig. 5. Switching arrangement for externally driven sawtooth generator
Note in this case the potentiometer forming the frequency control of Fig. 3 functions as an amplitude control

current generator being far from true except in the case of pentodes. It would appear, however, that despite the variation in the current gain A_i with increasing emitter currents, the transistor is very well suited to transform linear voltage sweeps of relatively low amplitudes into large current sweeps, the peak current swings being readily defined by the emitter resistance R_{e2} (Fig. 3) employed in the circuit. With reference to Fig. 3 stage X_3

$$(i_c/i_b) = A_i$$

and $R_{i1} = R_{e2}(1 + A_i)$

The base current i_b due to the sweep voltage V_b at the base of X_3 is given by

$$i_b = \frac{V_b}{R_{e2}(1 + A_i)}$$

so that $i_c = i_b \cdot A_i = (V_b/R_{e2}) \cdot \dfrac{A_i}{(1 + A_i)} \simeq (V_b/R_{e2})$

assuming $A_i \gg 1$.

Thus the collector current X_3 is largely independent of the current gain A_i, provided V_b is unaltered by variations in R_{i1} and

$$R_{e2}(1 + A_i) \gg r_b + (1 + A_i) r_e$$

or

$$R_{e2} \gg \frac{r_b}{1 + A_i} + r_e$$

(The short-circuit output resistance of the X_3 stage).

r_e is quite negligible for emitter currents of a few milliamperes or more so that R_{e2} of a few tens of ohms would be adequate to satisfy this condition.

However, the variations in input impedance of X_3 stage due to variation in α alters the dynamic emitter load resistance of the emitter-follower X_1, changing the magnitude of its R_i thereby distorting the sweep. This effect will be relatively insignificant provided

$$R_{e1} \ll A_i . R_{e2}$$

where A_i corresponds to the current gain of X_3. Thus provided R_{e1} and R_{e2} are chosen to satisfy the above requirements a single stage transistor (such as X_3 stage) can affect the conversion of sawtooth voltage into current with good accuracy.

Application to Electromagnetic Scanning

The coil L in Fig. 3 represents the effective value of the inductance in the collector circuit of X_3; the scanning coil

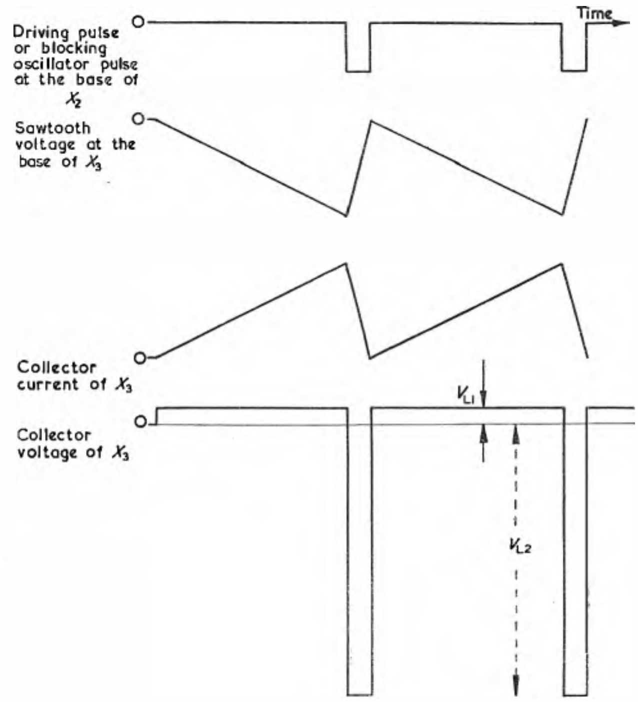

Fig. 6. Idealized waveforms of the circuits discussed

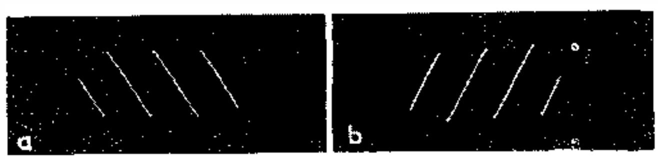

Fig. 7. Base voltage and collector current waveforms of X_3

itself is not normally placed in the collector circuit in order to avoid the positional shift encountered in the circuits having a d.c. voltage path. The value of the effective inductance that may be employed will depend upon the rate of change of current during the flyback owing to large negative voltage appearing at the collector during this period.

The dynamic collector voltage V_L of X_3 consists of two components V_{L1} during the 'run-down' and V_{L2} during the flyback where

$$V_{L1} = L(d_i/dt) = L . (I_s/t_s) \text{ volts}$$

and $V_{L2} = L(di/dt) = -L . (I_s/t_f)$ volts

where I_s is the peak sweep current through the collector of X_3, t_s and t_f the scanning period and flyback period respectively. If the scanning period is very much greater than the flyback $V_L \simeq -L(I_s/t_f)$.

Thus the inductance L of the coil in the collector of X_3 is limited by the maximum negative voltage permissible across the collector junction during the flyback and the rate of charge of the flyback current. In the circuits discussed above the effective value of L is just over 1mH being the inductance of scanning coils suitable for line scanning of a vidicon camera tube. For a flyback period of 6μsec and a maximum swing of 150mA, V_L is of the order of -30V.

Idealized waveforms are shown in Fig. 6 and the oscillograms of the actual waveforms of the base voltage and collector current of X_3 are shown in Fig. 7.

Thus where scanning current requirements are relatively low, current sweeps for magnetic deflexion may be generated by the circuits described above. It must be mentioned, however, that a value of 27Ω for R_{e2} is almost the limiting case for an OC72 type transistor, 150mA swing dissipating nearly its peak rated power. The conditions governing the choice of R_{e2} and R_{e1} are better satisfied by higher values for R_{e2} thereby limiting current swings in the circuits described. The variations of input impedance and voltage gain of the bootstrap sweep generator stage and the dissipation and maximum collector voltage ratings of the output stage are the main limitations to the quality and magnitude of the current sweeps. Owing to the need to replenish the charge lost by the feedback capacitor C_1 during the sweep before the subsequent sweep can commence, very high repetition rates cannot be achieved in the bootstrap type circuit which appears to be its basic disadvantage. The effect of variation in the collector cut off current I_{co} due to temperature is negligible in circuits where the charging current I is relatively very much higher. For a temperature range of 20 to 50° C, the variation in I_{co} for the above transistors was found to be less than 50μA.

Acknowledgments

In preparing this article the author has made use of some of the information and diagrams in the paper entitled "Junction Transistor Bootstrap Linear Sweep Circuits" by himself and Dr. A. R. Boothroyd, which was first published by the Institution of Electrical Engineers as Paper No. 2228R in January 1957 and republished in Part B of the Proceedings of that Institution.

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A Time-Multiplex Oscilloscope for Electroencephalography

By T. J. McDermott*, M.Sc., A.R.C.S.

A small instrument has been designed, using pulse time multiplex techniques, to amplify and display simultaneously on one c.r.t. six electroencephalographic potentials.

As part of a programme of investigation of the use of pulse time multiplex techniques for the simultaneous amplification and display of a number of electrophysiological potentials, it was decided to design a small instrument for early diagnostic purposes in electroencephalography. The use of the multiplex techniques entailed the use of a cathode-ray tube for presentation of the e.e.g. signals. It was considered, however, that a small inexpensive instrument which could display six channels of e.e.g. information on the screen of a c.r.t. with an after-glow phosphor would be quite useful for some e.e.g. purposes. The following instrument was designed and a laboratory model made, so that trials of its diagnostic value could be held.

The display of six e.e.g. signals on one c.r.t. was achieved by the use of a seven step echelon waveform of approximately 1msec period to produce a vertical deflexion of the trace to each of seven spots. From the echelon waveform six pulse trains, individually derived from each of the six steps, are used as pulse carrier of the e.e.g. signals. The separate amplitude modulated pulse trains are combined and amplified in the 'slicer-amplifier' and the amplified pulses elongated. These elongated amplitude-modulated pulses are added to the echelon waveform. Thus each spot of the c.r.t. trace has a vertical deflexion, proportional to a channel signal. A slow time-base which can be either repetitive or triggered by some external means, is applied to the X plates of the c.r.t. Thus the deflexions of the six spots are traced out across the face of the c.r.t. at speeds corresponding to those at which a normal e.e.g. records on paper. The record, of course, is visible only for a few seconds at a given time, and the operator has to make judgments of the nature of the record in that time.

The shortness of the time available for judgments to be made, suggested a departure from standard e.e.g. practice, that was used in the prototype. The localization of foci in e.e.g. practice depends on the recognition of 'phase reversals' in the bipolar recording. It was considered that such phase reversals would be very difficult to identify on the c.r.t. screen, so the system of unipolar recording with an 'electrical mean reference potential' was used. With such a system, foci would be identified by voltage maxima, and it was considered that this would be more satisfactory in practice.

Circuit Design

THE TIME DIVISION SYSTEM

The c.r.t. used was the DP13-2 a 6in diameter electrostatic deflexion tube, with a long persistence phosphor, and post deflexion acceleration.

A block diagram of the overall system is shown in Fig. 1 and a detailed circuit diagram in Fig. 2. Valve V_{E1} is a master-oscillator controlling the 10kc/s pulse repetition frequency of the whole system. The oscillator transformer T_{E1} has a secondary winding across which a 10kc/s sine wave of 2·0kV is developed. Two e.h.t. voltages of + 2·5kV and $-$ 1·7kV for the post deflexion accelerator anode, and cathode of the c.r.t. are produced by rectifiers 36eht 100 and the reservoir capacitors C_{E23} and C_{E24}. The oscillation (Fig. 3(a)), fed through diode V_{E2a}, is also used to control the frequency of the blocking oscillator V_{E1}, which produces a train of short pulses at the 10kc/s repetition frequency (Fig. 3(b)). The duration of these short

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pulses is determined by the short-circuited section of the delay line DL and their amplitude is limited by the diode V_{B2b}. The pulses produced in the cathode of V_{B1} are delayed by the section of the delay-line terminated by R_{B13}. The pulses occur across R_{B13} 1·5μsec after the current pulse in the anode resistor R_{B14}. This delay is necessary for the operation of the pulse amplitude detector system V_{A4} and V_{A5}.

From the train of 1μsec pulses developed across R_{B13}, an echelon waveform (Fig. 3(d)) of seven steps of period 700μsec is derived by the transitron controlled Miller-integrator V_{B4}. The echelon waveform appearing across the anode load resistors R_{B43} and R_{B44} is limited in the positive and negative directions by the diodes $V_{B5a,b}$. The third anode of the c.r.t. is taken to the junction of R_{B51} and R_{B52}, so that its potential is the mean potential of the anode of V_{B4}. The echelon waveform developed at the anode of V_{B4} is inverted by the anode-follower V_{B6}, and the Y plates of the c.r.t. are connected to these two anodes to produce a symmetrical deflexion of the c.r.t. spot.

A fraction $\dfrac{R_{B43}}{(R_{B43}+R_{B44})}$ of the echelon waveform developed at the anode of V_{B4} is fed via the cathode-follower V_{B7} to the anodes of the slicing diodes V_{K4b}. Resistors R_{K42} are returned to a sequence of tappings on a long resistance chain made up of the parallel series combination of resistors R_{K43} and R_{K44} with R_{K45-49}. These tappings are at the potentials V_{n1}-V_{n6} indicated in Fig. 3(d). The diodes V_{K4b} slice the echelon applied to their anodes to develop waveforms such as those shown in Fig. 3(e$_{3,4}$). The anodes of diodes V_{K4a} are returned via resistors R_{K41} to the sequence of potentials V_{n1-6}, and thus pulses, whose negative going edges correspond to the steps of the echelon waveform are produced across resistor R_{K41} as shown in Fig. 3(f,i). These pulses are differentiated and inverted by capacitors C_{K32} and ferroxcube pulse transformers T_{K1} to produce the channel pulses (Fig. 3(g,j)). These latter are applied to the grids of the channel modulator triodes V_{K3} in addition to the channel signal that is developed across resistor R_{K31}. Thermionic diodes type EB91 were used for the slicing diodes $V_{K4a,b}$ since point-contact germanium diodes were found to have a large low frequency random variation of their back-resistance.

The modulator triodes V_{K3} are maintained biased beyond cut off by the voltage across R_{A11} in the cathode of pentode V_{A1}. Each triode V_{K3} only conducts for the duration of the positive-going channel pulse applied to its grid, and then in conjunction with pentode V_{A1}, forms a long-tail pair of amplifiers. The current pulses developed in the cathode of channel triodes V_{K3} are separately amplitude modulated by the channel signals applied across resistors R_{K31}, from these pulses, positive voltage pulses are produced across the anode resistor R_{A12} of valve V_{A1}. These pulses are amplified in valves V_{A2} and V_{A3}. For this purpose double triodes are used as long-tail pair amplifiers. By their use the pulses are maintained as positive-going pulses applied to the input grids, which are negatively biased well beyond 'cut-off'. It is essential to avoid the flow of grid current in the slicing circuits, as the variation of the mean potentials which would result from the amplitude modulation of the pulses would introduce inter-channel cross-modulation. Consequently, the transfer of negative going pulses through a capacitance-resistance coupling with the slicing achieved by the flow of grid-cathode or diode current, is to be avoided. The interstage coupling has a time-constant of approximately 20μsec, sufficiently long to produce little distortion of the pulse shape, but short enough that the cross-modulation produced by the pulse tail is negligible. The balanced current taken in the long-tail pair reduces the difficulties of feedback to the common load resistance of the power pack, and no decoupling is necessary.

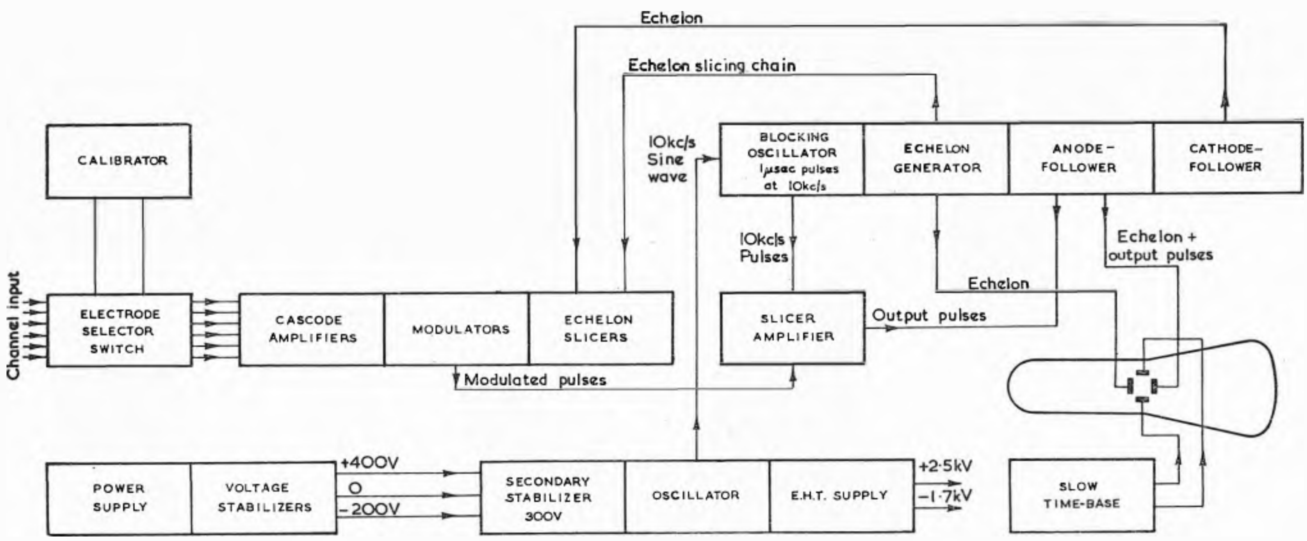


Fig. 1. General arrangement of the instrument

The amplified pulses are fed to the 'pulse amplitude detector' V_{A4}. The positive-going pulses from the anode of V_{A3} (Fig. 3(h)) are fed to the grid of the cathode-follower V_{A4a}. A capacitor C_{A42} is connected as the cathode load of V_{A4a} as is also the grid of cathode-follower V_{A4b}. So long as C_{A42}/g_m is small compared with the duration of the applied pulse, the capacitor is charged to the peak potential of the applied pulse and remains so charged. At the end of a channel interval, a negative pulse, from the anode of V_{B1} is applied to the capacitor C_{A42} through the diode V_{A5a}. The capacitor is thus discharged to a fixed negative potential 1·5μsec before the next channel pulse arrives to recharge it. Thus the output voltage across the cathode load R_{A4} of cathode-follower V_{A4b}, is maintained for the 100μsec channel interval, at the peak potential of the channel pulse. The resultant pulse train is illustrated in Fig. 3(i). This pulse train is added to the echelon waveform by the grid resistance network $R_{B61,62,63,64,65}$, of the anode-follower V_{B6}. Consequently the vertical deflexion of each spot of the c.r.t. will be proportional to the amplitude of the appropriate channel pulse.

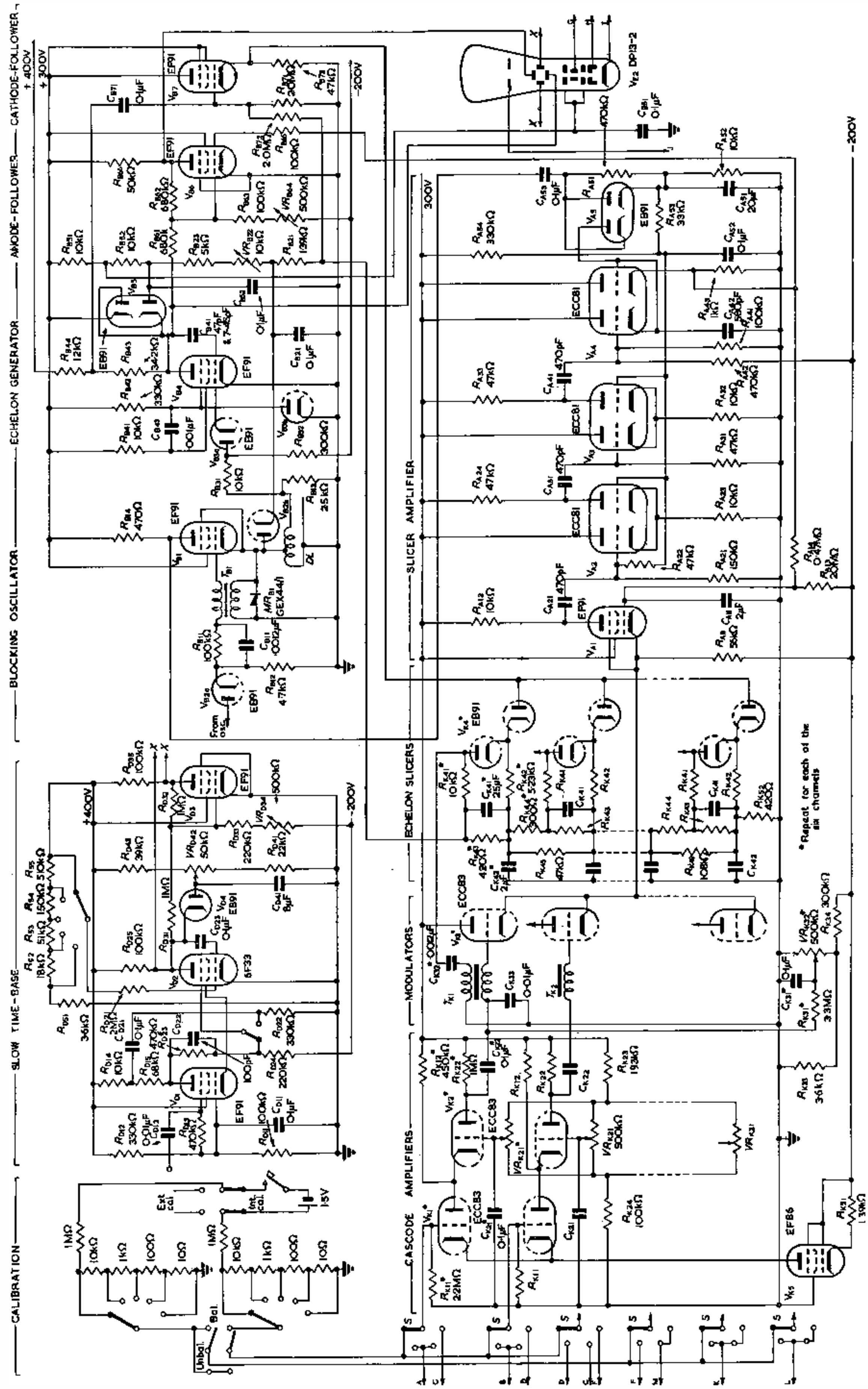


Fig. 2(a) The oscilloscope circuit *(continued on page 68)*

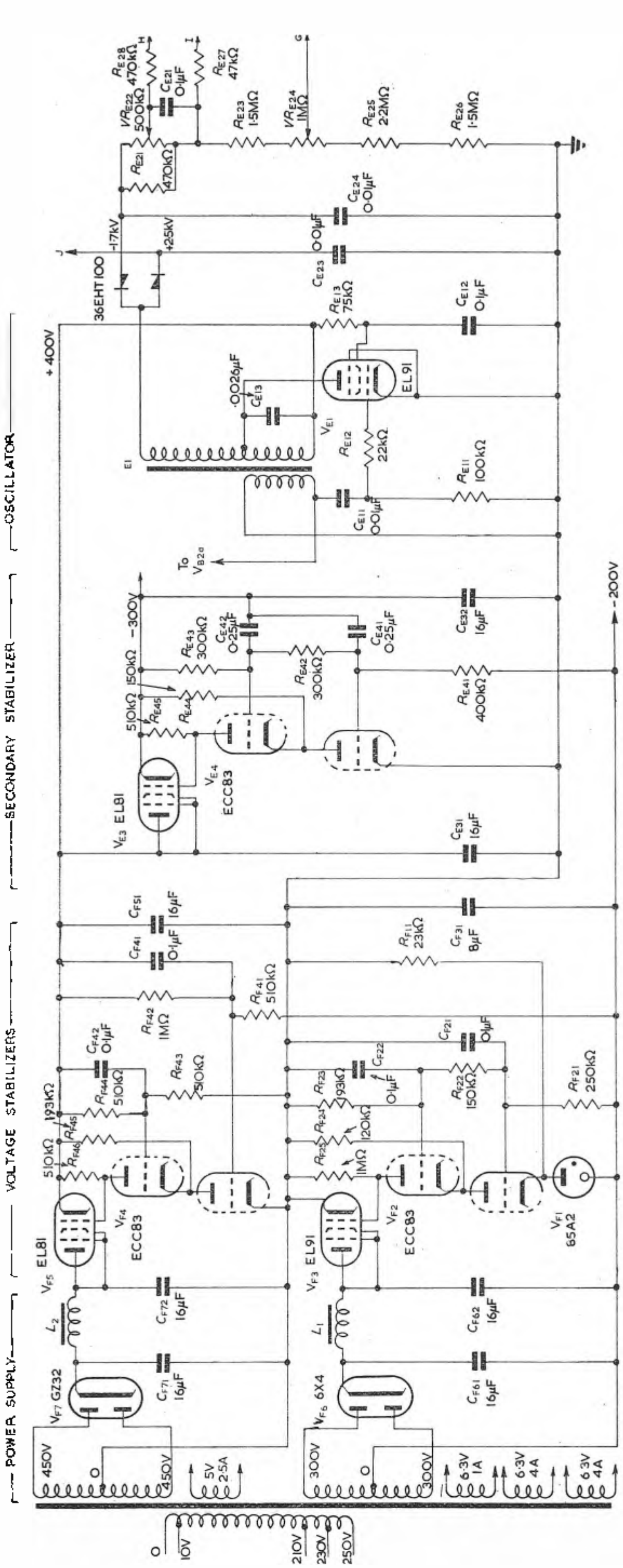


Fig. 2(b). The power supply unit and oscillator

Negative feedback which serves both to stabilize the output with respect to variations in slicer-amplifier gain, and also to refer the channel outputs to a reference 'weighted' mean potential is incorporated. The pulse train across cathode resistor R_{A13} is integrated by a long time-constant resistance-capacitance circuit (R_{A14}, C_{A11}) and applied to the grid of V_{A1}. The mean potential so produced is thus subtracted from the channel signal across R_{K31}, in the pulse modulator pair V_{K3}-V_{A1}. Deriving the feedback potential from a common output system instead of using a separate control channel reduces the number of valves necessary, but complicates the operation of the system. Unless the feedback time-constant is large compared with the product of a channel gain and pulse interval, the control voltage will vary from channel to channel and a high frequency inter-channel cross-modulation will be produced. The feedback time-constant must therefore be long and the resultant output becomes out of balance at high frequencies.

THE PRE-AMPLIFIER

In order to make the system sufficiently sensitive for e.e.g. work a pre-amplifier was introduced for each channel. It was known from previous work that a random noise voltage equivalent approximately to 100μsec peak-to-peak could be expected at the input of the channel modulators. For e.e.g. purposes an input noise level of 1 to 3μsec peak-to-peak was necessary. It was decided to use a pre-amplifier, the amplified input noise of which would be much greater than the 100μsec at the channel modulator input. For this purpose a double triode V_{K1} and V_{K2} per channel was used, as a cascode amplifier. The resistor R_{K12} by-passes the second triode V_{K2}, so that 0·5mA anode current can flow in V_{K1}, giving an effective mutual conductance of 0·6mA/V, while only approximately 0·1mA flows in V_{K2} and its anode load resistor R_{K22}. Thus an anode load resistance of 1MΩ can be used, and an overall gain of 600 is achieved. To achieve the desired unipolar electrical mean properties, the cathodes of the six channel amplifiers are connected together as the anode load of the EF86 pentode V_{K5}. The common cathode line becomes a 'weighted' electrical mean reference potential given by

$$\delta v_k = \frac{r_a \sum_1^n g_m \, \delta v_{g1}}{1 + r_a \sum_1^n g_m}$$

where v_k = potential of common cathodes

r_a = internal anode resistance of V_{K5} $\simeq$ 1MΩ.

g_m = mutual conductance of each triode $V_{K1} \simeq 0{\cdot}6$mA/V.

v_{g1} = input grid-ground potential.

For six channels $\sum_1^n g_m \simeq 3{\cdot}6$mA/V

Therefore $\qquad r_a \sum_1^n g_m \simeq 3\,600$

Therefore one can put this as $\delta v_k = \dfrac{\sum_1^n g_m \, \delta v_{gi}}{\sum_1^n g_m}$

which is a weighted mean of the input potentials. A preset adjustment VR_{K31} of the grid potential of the triodes V_{K2}, enables one to equalize their anode currents and thus those of the triodes V_{K1}. Since for triodes V_{K1}, g_m/i_a is approxi-

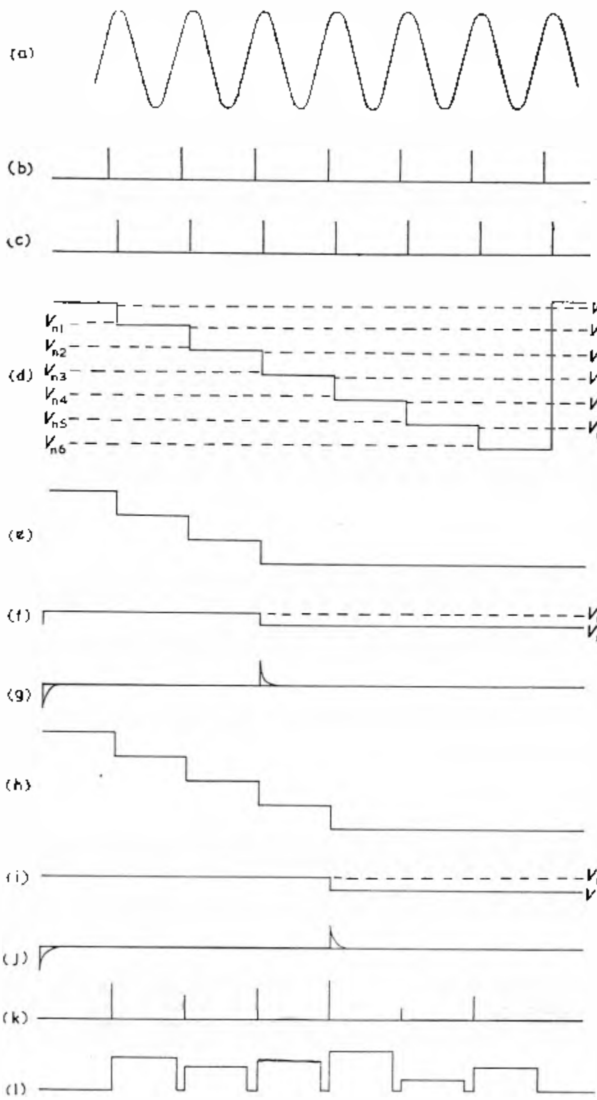


Fig. 3. Waveform diagram

mately the same, this arrangement was introduced to provide a method of equating the channel gains. For practical purposes the important consideration is the rejection of unwanted hum and muscle potentials that are common to all channels. If the input grid signals are expressed as a common signal δv_g plus an individual channel signal δv_{gi}, then

$$\delta v_{gi} = \delta v_g + \delta v_{gi}$$

The output from a given channel anode is

$$R_a g_m \, (\delta v_{gi} - \delta v_k)$$

$$= R_a g_m \left(\frac{\delta v_g}{1 + r_a \sum_1^n g_m} + \delta v_{gi} - \frac{\sum_1^n g_m \, \delta v_{gi}}{\sum_1^n g_m} \right)$$

The factor $(1 + r_a \sum_1^n g_m)$ gives the out of balance rejection ratio which was shown to be approximately 3 600. This rejection ratio is obtained in practice, so long as no phase change or frequency distortion takes place before the grids of the input amplifiers.

The restriction of the frequency band necessary for e.g. purposes is achieved in the coupling between the input amplifiers and the channel modulators. In the prototype switching of the time-constant and upper frequency response has not been incorporated. A fixed time-constant of: $C_{K21} = 0{\cdot}1\mu F$, $R_{K31} + R_{K22} = 3{\cdot}3 + 1{\cdot}0M\Omega$, i.e. $T = 0{\cdot}25$sec. is used, and the upper frequency response determined by the anode load resistor $R_{K22} = 1{\cdot}0M\Omega$ in parallel with $R_{K31} = 3{\cdot}3M\Omega$ and the by-pass capacitor $C_{K33} = 0{\cdot}01\mu F$, gives a cut-off frequency (3dB) of approximately 20c/s. Such a low cut-off frequency was incorporated as a precaution against 50c/s pick-up.

The only limitation experienced in the use of the six cascode input amplifiers, is the rather high noise level. Using unselected 12AX7 (ECC83) which had been given a 300h ageing period during which 6·3V d.c. was applied to the heaters, a fundamental input noise level of 5μV peak-to-peak is obtained with the above input bandwidth 0·5 to 20c/s. This is obtained using a 12V d.c. heater supply, stabilized only by the use of a constant voltage transformer.

The Ancillary Circuits

The system of seven spots in a vertical line on the face of the c.r.t., six of which are vertically deflected, is moved horizontally by the slow time-base. A balanced deflexion potential is derived from the anodes of a Miller-transitron V_{D2} and an anode-follower V_{D3}. The speed of the trace is switchable to values 3, 10, 30 or 100cm/sec by returning the grid resistor of V_{D1} to various tappings of the resistor chain R_{D51-55}. The coupling from the screen grid to the suppressor grid of the Miller transitron V_{D2} is effected via the cathode and anode of a pentode V_{D1}. By this means the system can be maintained indefinitely with the anode of the Miller-transitron V_{D2} cut off, without exceeding the power dissipation of the screen grid. The circuit can thus be switched from a free-running to a single stroke condition. The suppressor grid of V_{D1} can be used to inject pulses to trigger or synchronize the time-base.

The power supply for the equipment consists of a stabilized negative supply of -200V which is used as a reference potential to control a stabilized 400V supply and a second doubly stabilized $+300$V supply. A stabilizer type 85A2 is used as a reference for the -200V supply, and cascode amplifiers are used in all the stabilizers. The $+300$V supply is used only for those circuits which require a highly stabilized supply to minimize inter-channel cross-modulation and random variations of the pulse amplitude.

The Performance of the Laboratory Model

Linearity of Response

Since the signals are displayed on the face of a c.r.t. and no power amplifier stages or pen recorders are involved, the major source of non-linearity would be the pulse amplitude detector. A Lissajous figure method of measurement on the face of the c.r.t. showed no appreciable non-linearity.

Amplification

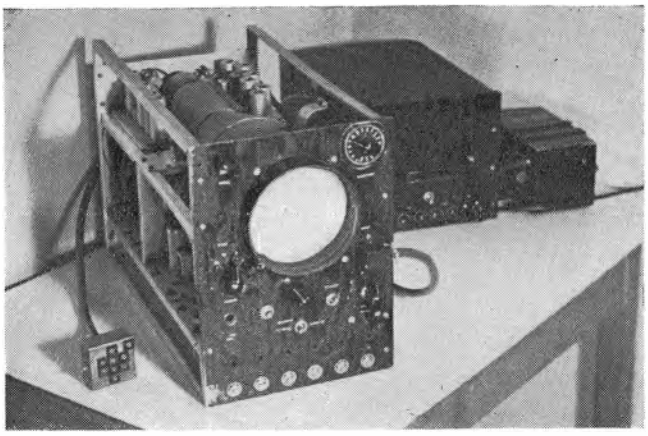
No overall gain control was incorporated in the model.

The variation of individual channel gain, achievable by adjustment of the cascode 2nd grid potential proved to be not sufficient to compensate for the variations of channel modulators. A more satisfactory method of adjustment of the modulator transconductance must be developed for a future model. A signal of $50\mu V$ applied to each channel produced a spot deflexion of approximately 10mm. For each channel the maximum working deflexion is 1·3mm, the spot size is approximately 0·3mm, thus a ratio:

$$\frac{\text{Maximum working deflexion}}{\text{Minimum visible deflexion}} = 40 \text{ is obtained}$$

FREQUENCY RESPONSE

The limitations of the design are such that with an echelon repetition frequency of 1kc/s, an upper response frequency of 300c/s is attainable. The lower response frequency is determined solely by the inter-stage capacitor resistor coupling between the cascode anodes and the channel modulators. This model had a fixed upper response frequency of 25c/s (3dB) and a time-constant of 0·4sec for e.e.g. use.



The complete instrument

BASIC IRREGULARITY

With the inputs connected to ground the channels have a 'noise' variation of approximately $3\mu V$.

DISCRIMINATION

With all the channel inputs connected together, a potential applied between the inputs and ground produces a deflexion no greater than 0·1 per cent of that produced by the same potential applied between one electrode and all the rest.

CALIBRATION

A small switched calibrator was included which gives balanced or unbalanced inputs of $7{\cdot}5\mu V$, $75\mu V$, $750\mu V$, 7·5mV.

WRITING SPEEDS

The time-base could be used in either a self-running or single-stroke condition at speeds of 1cm/sec, 3cm/sec, or 10cm/sec.

PHYSICAL DIMENSIONS

The instrument consisted of a display unit 18in by 12 in by 9in of 20lb weight and a power supply unit 12in by 9in by 8in of 22½lb weight. The total power consumption was less than 150W.

The Use of the Instrument for E.E.G. Diagnosis

Since no permanent record of the e.e.g. is taken, this instrument cannot be used as an ultimate reference for diagnostic purposes. However, it can play an active part, in places where the cost of a large e.e.g. machine is prohibitive, in selecting patients for reference to the distant e.e.g. departments.

A clinical survey of the validity of this instrument for diagnostic purposes, is still in progress. No adequate conclusions can yet be drawn, but the following observations appear.

(1) The appearance of the traces is indistinguishable from an ordinary e.e.g. record, but the short time available for observing any section of the record, make greater demands on the intelligence and quickness of the operator. It is important that the operator should know what to look for.

(2) Without any additional time marking system, it is possible, after a little practice, to estimate the frequency of an oscillation in the range of 1 to 15c/s to approximately 1c/s by simple observations and estimations by eye of a few cycles. A time marking pulse was applied to the c.r.t. blacking out all the traces for a few milliseconds at a controllable frequency, this by itself proved to be of little value. It would be of more value if the repetitive time-base were synchronized to several periods of the time marker.

(3) The assumption that it would be impossible to detect phase reversals proved invalid, although it is impossible to detect them by observation of the yellow afterglow trace. By watching the spots themselves, and the direction of their vertical movement, it is comparatively easy to detect phase reversals, at least of adjacent traces.

(4) The use of unipolar recording, with only six channels is not justified. In practice, the distribution of focal e.e.g. gradients are such that the steep potential gradients are localized in the region of one electrode only. In these circumstances, the averaging potentials that occur in the unipolar recording method, only produce confusion, and bipolar recordings are much more satisfactory for approximate localization by the observations of voltage maxima.

Conclusion

Since the completion of this instrument and while the clinical trials were progressing, work has been proceeding on other approaches and techniques.

The replacement of the thermionic valves in the preamplifier and channel modulators by junction transistors would reduce the need for both highly stabilized h.t. supplies and for a stabilized heater supply. Junction diodes have been shown to be at least as good as thermionic diodes in the channel slicer circuits. The new r.f. junction transistors are being tried for the first stages of the slicer amplifier.

As an alternative approach to the pulse multiplex system, a ring stepping register system using ferrite saturable magnetic cores is being tried.

An alternative display system is being developed using electromagnetic deflexion radar p.p.i. tubes which can economically be of larger diameter than the electrostatic deflexion c.r. tubes. A linear current sawtooth, synchronized with the echelon waveform, deflects the c.r.t. spot in a vertical line. The output waveform from the pulse amplitude detector is transformed into a train of short pulses time modulated by the amplitude modulation of the previous output pulses. These short pulses are applied to the grid of the c.r.t. producing spots whose vertical position reproduces the vertical arrangement of the spots of the electrostatically deflected tube. A slow current sawtooth deflects the whole system horizontally to produce the same patterns as before. A recorder, which uses this principle to produce a permanent record on Teledeltos paper is showing considerable promise.

A New Bistable Element Suitable for Use in Digital Computers

(Part 1)

By C. D. Florida*, A.M.I.E.E.

The article describes the development of a trigger circuit which uses both pnp and npn types of transistor. The circuit has fast (0·2μsec) switch-on and switch-off times but is characterized primarily by its current handling capacity and hence by its ability to drive several other similar circuits. The transient response of this circuit to a voltage ramp input at either the turn-on or the turn-off terminals is dealt with in two mathematical appendices. The method is to divide each operation into a number of discrete stages, each of which is terminated by a significant change in the circuit configuration (e.g. a diode current falling to zero, thereby isolating the diode from the circuit). The initial conditions for any stage are derived from the equations for the preceding stage, and are inserted into the equations at the same time that the force functions are replaced by their Laplace transforms. The inverse Laplace transform then gives the required response for that stage as a function of time.

Results of the mathematical treatment are plotted, assuming certain circuit values and also assuming that the trigger circuit under analysis is driving five other similar circuits. Experimental results are also plotted and it is seen that the agreement is good.

A TRANSISTOR operated digital computer is being built at the laboratories of the Defence Research Board in Canada. For the purpose of the present article it is sufficient to state that this computer uses trigger circuits inter-connected by passive gates. With this type of gate it is desirable that a trigger circuit should combine fast switching speeds, high current handling capacity and a low output impedance. Table 1 shows a specification for a trigger circuit which will meet the requirements of this computer, and the development of such a circuit is described in this article.

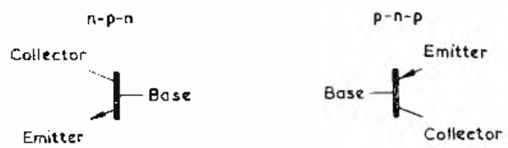


Fig. 1. The symbols used

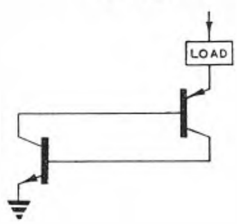


Fig. 2. Basis of fast switching circuit

The circuit uses both pnp and npn types of transistor and it is consequently necessary to distinguish between these types in diagrams. Although they are not considered ideal, the symbols specified by B.S.I. are used and these are shown in Fig. 1.

TABLE 1
Specification

Switching time	< 0·2μsec
Output fall time when driving 5 similar stages	< 0·2μsec
Output rise time when driving 5 similar stages	< 0·5μsec
Output load current	> 30mA
Output impedance	< 20Ω
Resolving time	< 1·5μsec

* *Defense Research Telecommunication Establishment, Canada.*

Development of the Circuit

If two transistors, one an npn, the other a pnp are joined together as shown in Fig. 2, there exists the basis of a fast switching circuit as the regeneration is the highest possible for any way of connecting together two transistors. This rudimentary circuit has stability in the on state and the ability to handle high currents, limited only by the wattage dissipation of the transistors.

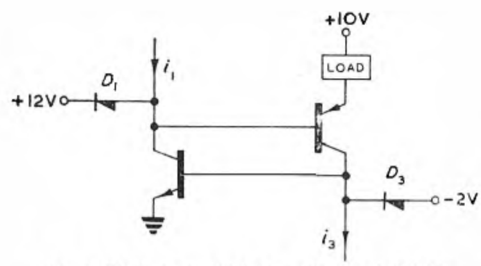


Fig. 3. Method of obtaining 'off state' stability

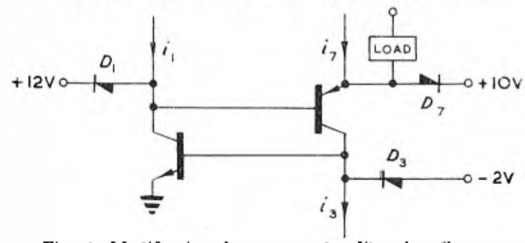


Fig. 4. Modification for current handling in off state

Stability in the off state can be gained by the addition of two current sources and two diodes, one to each base, as shown in Fig. 3. A reverse bias of 2V is applied to each transistor when in the off state, and it is easy to ensure that the current sources are large enough to take care of any possible value of collector leakage current.

The circuit in this form has stability in both the on and the off states and can handle high currents when in the on state. In the off state, however, it cannot handle any current. This may be corrected by the addition of another current source and diode as shown in Fig. 4. Here a current i_7 flows into the diode D_7. Any current up to this value may be withdrawn from the structure without change of potential, while the current flowing into the structure is limited only by considerations of diode wattage.

Switching the circuit from off to on may be accomplished by any means which injects current into the regenerative loop. One effective method is shown in Fig. 5. Diode D_4 is inserted and a negative-going voltage waveform is applied to the capacitor *C*.

Switching Speed

The development of the trigger circuit has thus reached the point where there are stable off and on states, high current handling capacity in both states, low output impedance, and a means for switching from off to on. The switching speed is analysed in fair detail in Appendix A. There, it is shown that the switching process can be divided into several discrete stages, and an equation is given for each stage from which the time taken for that stage can be derived. The results of this Appendix have been applied

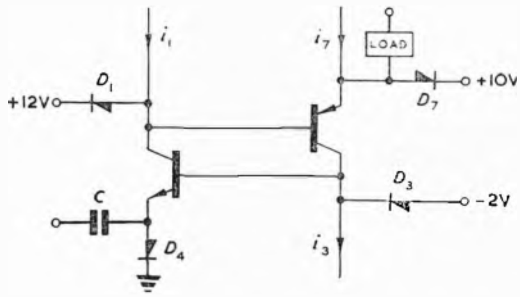


Fig. 5. Method of switching

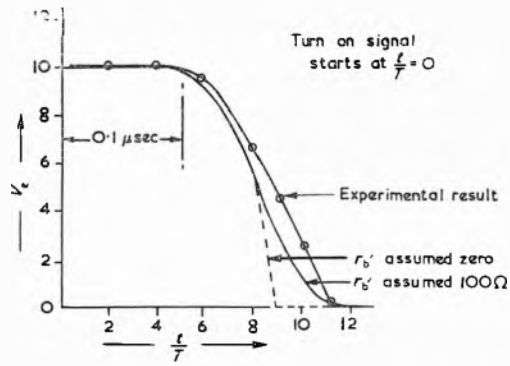


Fig. 6. Potential at output terminal against time-constants

to a circuit with the following parameters:

i_1 = 0·25mA

i_3 = 3mA

i_7 = 10mA

Transistor α cut-off frequency = 8Mc/s ($T = (1/2\pi f_\alpha) = 2 \times 10^{-8}$sec)

Transistor base/collector capacitance 10pF (k = 0·2)

Input capacitor = 100pF

Input voltage fall = 10V in 0·2µsec ($a = 5{\cdot}10^7$V/sec)

Output capacitor = 500pF (n = 5)

The value chosen for the output capacitor is equivalent to a loading of five trigger circuits identical with the one being discussed.

Fig. 6 shows computed and experimental results for the potential at the output terminal V_o, plotted against transistor time-constants. It will be seen that the turn-on cycle can be broken down into a delay of about 0·1µsec and a fall time of about the same duration. For the early stages in the switching process, where the transistor currents are low, the agreement between the computed and experimental values is good, but during the rapid portion of the fall the agreement is only fair. This is not unexpected since the analysis is based on the assumption that transistor parameters are invariant with voltage and current, and a fall at the rate of that shown by a dashed line corresponds with transistor currents of up to 300mA. Even so, it is thought that the analysis will permit overall times to be computed with an accuracy not worse than ±10 per cent if a non-zero value is assumed for base spreading resistance during the times of rapid voltage change.

Further Development

In the circuit of Fig. 5 the fall of the output electrode continues until both transistors saturate. This would cause the circuit to be difficult to turn off but can be avoided by the use of two catching diodes, as in Fig. 7, which ensures a base-collector separation of 3V in the on state.

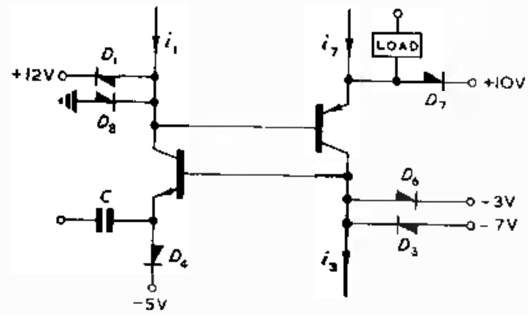


Fig. 7. Addition of catching diodes

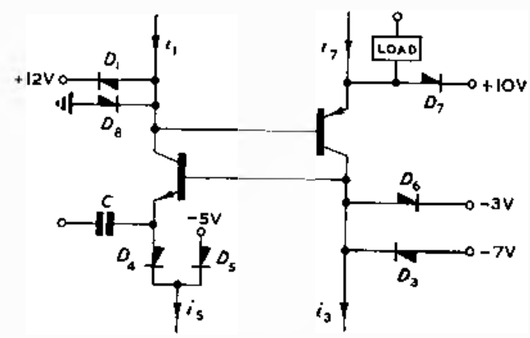


Fig. 8. Modification of Fig. 7

In Fig. 7 a path exists for virtually unlimited current to flow in the npn transistor (via D_8 and D_4). This can be prevented by feeding this transistor with a constant current i_5 of, say, 5mA as in Fig. 8. Any loss of fall time resulting from the restriction on current can be compensated by attaching a capacitor at the junction of D_4 and D_5 to provide a momentary source of high current.

When switching on has been completed, nearly 5mA is available, either to D_8 or to the base of the pnp transistor, which acts as an emitter-follower. If this transistor has a current gain of 20 then up to 100mA can be extracted from the load for a change in potential of only a few tenths of a volt—an output impedance of around 10Ω.

The circuit in this form has fast switching speed, high current-handling capacity and low impedance in both the on and the off states, and the ability to drive at least five similar circuits. It is clamped in the off state at +10V and at ground potential in the on state. It remains to turn it off.

Turn off can be effected in many ways, but in order to equalize turn-on and turn-off sensitivities, and to make turn-off sensibly independent of load, it is preferable to use an extra transistor connected as in Fig. 9.

This transistor normally has 1V of reverse bias. Application of a negative-going turn-off signal will cause base current to flow, thereby inducing an amplified current in the collector circuit. This current as it increases, first turns off D_8 and then feeds the output transistor with reverse base current so that its emitter current is eventually reduced to zero. The output transistor will thus turn off, and the surplus current from the turn-off transistor will charge up the base-collector capacitances so that the potential at the pnp base rises until D_1 switches on.

When D_1 switches on further positive excursion is stopped. This prevents the turn-off transistor from saturating and also causes the current previously flowing in the base-collector capacitances to cease abruptly. Cessation of this current causes D_6 to cease conduction so that the base of the npn transistor is free to fall in potential, so switching off that transistor.

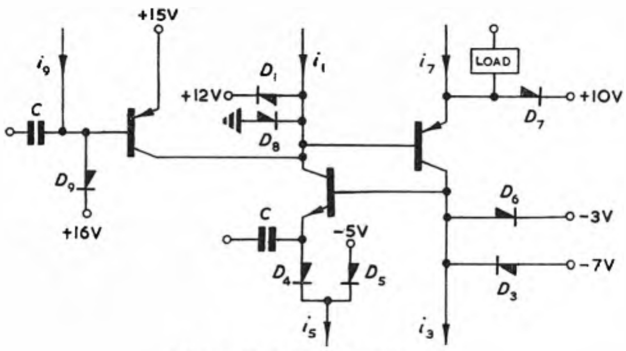


Fig. 9. Method of 'turning-off'

Appendix A

Turn On

Fig. A.1 is the same as Fig. 5 of the article, with the addition of collector-base and emitter-base capacitances. If a negative-going voltage ramp is applied to the capacitor C, as shown, the turn-on cycle has a number of stages, as follows:

(1) The d.c. starting potential at the emitter of X_1 is indeterminate as it is governed by the reverse resistances of D_3, X_1 base-emitter diode, and D_4. No current can flow in X_1 until this potential has been reduced to that on D_3 (-2V). Stage 1 is the time taken for this to occur.

(2) During stage 2 emitter current flows in X_1, whose collector current starts to rise at a rate depending on the α cut-off of the transistor. When this collector current reaches a value i_1, D_1 cuts off.

(3) In stage 3 X_1 collector current continues to rise, the surplus over i_1 being drawn from C_e (pnp) and $2C_c$. This stage is terminated when X_2 commences to draw current.

(4) Once X_2 draws current its emitter and base are held at $+10$V. X_2 collector current builds up until stage 4 is ended by D_3 switching off.

(5) The base of X_1 is now free to rise from -2V to ground potential. The time taken for this to occur comprises stage 5. During this stage D_7 may cease conduction owing to X_2 emitter current exceeding i_7. If this happens, stage 5 is terminated at this point.

(6) If stage 5 ends because of the release of D_7, stage 6 covers the remaining time before the potential of X_1 base reaches ground potential.

(7) During stage 7 the output electrode falls in potential to ground level.

These stages are all calculable and results can be computed using measured parameter values. However, two stages contain parameters which may change with time or from circuit to circuit, or which require excessive measurement time. These stages are (1) because of the dependence of starting potential on reverse resistances and also because of the difficulty of predicting from measurements the effective value of C_e as the transistor base-emitter potential

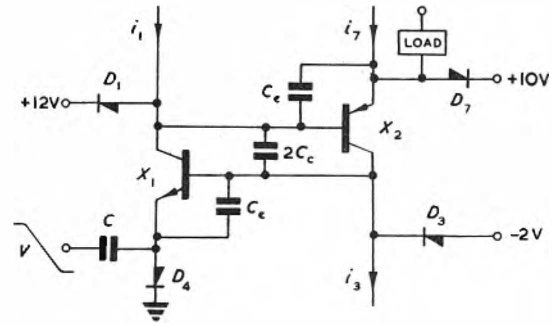


Fig. A1. Addition of collector-base and emitter-base capacitances to Fig. 5

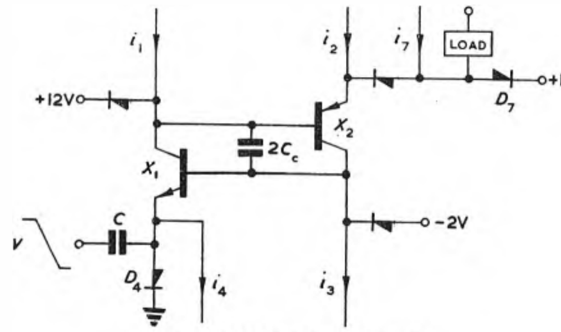


Fig. A2. Modification of Fig. A2

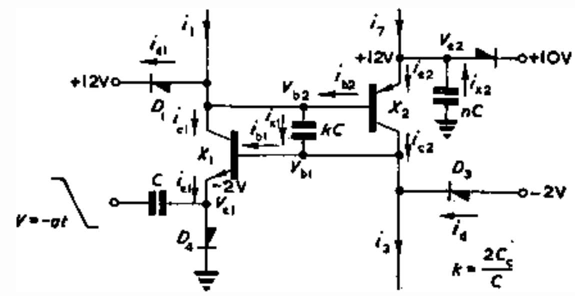


Fig. A3. Circuit for analysis

nears zero, and (3) again because of C_e. These difficulties can be overcome, both for purposes of analysis and also in practice by the provision of 'wetting' currents to both transistors. C_e then becomes merged with the diffusion capacitance which is represented in the operational form for current gain, while the emitter of X_1 is held at -2V by diode action when the circuit is in the off state. Fig. A.2 shows the modified circuit. The only necessary tolerance conditions are

$$i_4 < i_1 - 2\,I_{co}$$
$$i_2 < i_3 - 2\,I_{co}$$

In the analysis it will be assumed that i_2 and i_4 are so small that they do not influence the circuit in any ways other than those just described.

It will be assumed that the load consists of a number, n, of stages in parallel, each similar to the one being analysed. To a close approximation this load is given by n capacitors each of value C.

As X_1 and X_2 bases are nearly always fed from high impedance sources, very little error will occur if it is assumed that the base spreading resistance is zero.

The circuit for analysis now appears as in Fig. A.3. The presence of transistor wetting currents is implied by writing the emitter potentials near the emitters. In the analysis the stage numbers will be altered from those previously described, because the former stage 1 has now been deleted.

Stage 1

We have $i_{e1} = -pCV$

$$V = -at$$

$$i_{c1} = \alpha i_{e1}$$

$$\alpha = \frac{\alpha_0/T}{p+(1/T)}$$

$$T = \frac{1}{2\pi f_{\alpha}}$$

$$i_{d1} = i_1 - i_{e1}$$

Combining these equations,

$$i_{d1}(p+(1/T)) - i_1(p+(1/T)) + p(\alpha_0/T)aCT = 0$$

Initially $i_{d1} = i_1$, $at = 0$

$$\therefore\ \bar{i}_{d1}(p+(1/T)) - \bar{i}_1(p+(1/T)) + p(\alpha_0/T)\overline{aCt} = i_1 - i_1 = 0$$

$$\bar{i}_1 = (i_1/p),\ \overline{aCt} = (aC/p^2)$$

$$\therefore\ \bar{i}_{d1}\,p(p+(1/T)) - i_1(p+(1/T)) + (\alpha_0/T)aC = 0$$

$$\therefore\ \bar{i}_{d1} = \frac{i_1 - \alpha_0 aC}{p} + \frac{\alpha_0 aC}{p+(1/T)}$$

$$\therefore\ i_{d1} = i_1 - \alpha_0 aC(1 - e^{-t/T}) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (1)$$

This equation defines the duration of stage 1. Also of interest during this stage are i_{c1} and i_{d3}.

$$i_{c1} = i_1 - i_{d1} = \alpha_0 aC(1 - e^{-t/T}) \quad \ldots\ldots\ldots\ (2)$$

$$i_{d3} = i_3 + i_{b1} = i_3 + i_{e1} - i_{c1} = i_3 + aC - \alpha_0 aC(1-e^{-t/T})$$

$$= i_3 + aC[1 - \alpha_0(1 - e^{-t/T})] \quad \ldots\ldots\ldots\ldots\ldots\ (3)$$

Stage 2

Stage 2 is defined by the time taken for V_{b2} to fall by 2V (from +12 at end of stage 1 as D_1 cuts off to +10V as X_2 emitter-base diode action inhibits further fall).

We have $V_{b2} = -i_{x1}/pkC$

$$i_{x1} = i_{c1} - i_1$$

$$i_{c1} = \alpha i_{e1} = \alpha aC$$

$$\alpha = \frac{\alpha_0/T}{p+(1/T)}$$

Combining,

$$V_{b2}\,p(p+(1/T)) + (\alpha_0 aC/kCT) - (i_1/kC)(p+(1/T)) = 0$$

Initially $V_{b2} = 0$, $(dV_{b2}/dt) = 0$.

$$\therefore\ \bar{V}_{b2}\,p(p+(1/T)) + (\alpha_0 \bar{a}C/kCT) - (\bar{i}_1/kC)(p+(1/T)) = -(i_1/kC)$$

$$\bar{a} = (a/p),\ \bar{i}_1 = (i_1/p)$$

$$\therefore\ \bar{V}_{b2}\,p^2(p+(1/T)) + \alpha_0 aC/kCT - (i_1/kCT) = 0$$

$$\therefore\ \bar{V}_{b2} = (T/kC)(i_1 - \alpha_0 aC)[(1/Tp^2) - (1/p) + (1/p) + (1/T))]$$

$$\therefore\ V_{b2} = (T/kC)(i_1 - \alpha_0 aC)[(t/T) - 1 + e^{-t/T}] \quad \ldots\ (4)$$

This is the governing equation for this stage, but one is also interested in i_{c1} and i_{d3}

$$i_{c1} = i_1 - kC(dV_{b2}/dt) = \alpha_0 aC - [\alpha_0 aC - i_1]e^{-t/T} \quad \ldots\ (5)$$

$$i_{d3} = i_3 - kC(dV_{b2}/dt) + aC - i_{c1}$$

$$= i_3 + aC - i_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (6)$$

Stage 3

At the end of stage 2, X_2 commences to draw current. This current will build up until D_3 cuts off. This time interval is termed stage 3.

We have $i_{d3} = i_3 + i_{b1} - i_{e2}$

$$i_{b1} = i_{e1} - i_{c1}$$

$$i_{c1} = \alpha i_{e1}$$

$$i_{e2} = \frac{\alpha}{1-\alpha} i_{b2} = \frac{\alpha}{1-\alpha}(i_{c1} - i_1)$$

$$= \frac{\alpha^2}{1-\alpha} i_{e1} - \frac{\alpha}{1-\alpha} i_1$$

$$\therefore\ i_{d3} = i_3 + i_{e1}\frac{1-2\alpha}{1-\alpha} + \frac{\alpha i_1}{1-\alpha}$$

Writing $\alpha = \frac{\alpha_0/T}{p+(1/T)}$ and re-arranging,

$$i_{d3}\left(p + \frac{1-\alpha_0}{T}\right) - i_3\left(p + \frac{1-\alpha_0}{T}\right) - i_{e1}\left(p + \frac{1-2\alpha_0}{T}\right) - i_1(\alpha_0/T) = 0.$$

Initially $i_{d3} = i_{e1} + i_3 - i_{c1}(t_2)$ where $i_{c1}(t_2)$ represents the value of i_{c1} at the end of stage 2,

$$\therefore\ \bar{i}_{d3}\left(p + \frac{1-\alpha_0}{T}\right) - \bar{i}_3\left(p + \frac{1-\alpha_0}{T}\right) - \bar{i}_{e1}\left(p + \frac{1-2\alpha_0}{T}\right) - \bar{i}_1(\alpha_0/T) = -i_{c1}(t_2)$$

$$\bar{i}_3 = (i_3/p),\ \bar{i}_{e1} = (aC/p),\ \bar{i}_1 = (i_1/p)$$

$$\therefore\ \bar{i}_{d3}\,p\left(p + \frac{1-\alpha_0}{T}\right) - i_3\left(p + \frac{1-\alpha_0}{T}\right) - aC\left(p + \frac{1-2\alpha_0}{T}\right) - i_1(\alpha_0/T) = -p\,i_{c1}(t_2)$$

$$\text{or } \bar{i}_{d3} = \frac{i_3 + aC\dfrac{1-2\alpha_0}{1-\alpha_0} + i_1\dfrac{\alpha_0}{1-\alpha_0}}{p} + \frac{(aC - i_1)\dfrac{\alpha_0}{1-\alpha_0} - i_{c1}(t_2)}{p + [(1-\alpha_0)/T]}$$

$$\therefore\ i_{d3} = i_3 + aC\frac{1-2\alpha_0}{1-\alpha_0} + i_1\frac{\alpha_0}{1-\alpha_0}$$

$$+ \left[(aC - i_1)\frac{\alpha_0}{1-\alpha_0} - i_{c1}(t_2)\right]\exp\left[-\frac{1-\alpha_0}{T}t\right] \quad \ldots\ (7)$$

By analogy with equation (5), i_{c1} during stage 3 is given by

$$i_{c1} = \alpha_0 aC - [\alpha_0 aC - i_{c1}(t_2)]e^{-t/T} \quad \ldots\ldots\ (8)$$

Also,

$$i_{e2} = i_{e1} + i_3 - i_1 - i_{d3}$$

$$= aC\frac{\alpha_0}{1-\alpha_0} - i_1$$

$$- \left[(aC - i_1)\frac{\alpha_0}{1-\alpha_0} - i_{c1}(t_2)\right]\exp\left[-\frac{1-\alpha_0}{T}t\right] \quad \ldots\ (9)$$

and $i_{c2} = i_3 + i_{b1} - i_{d3} = i_3 + i_{e1} - i_{c1} - i_{d3}$

$$= \frac{\alpha_0}{1-\alpha_0}(\alpha_0 aC - i_1) + [\alpha_0 aC - i_{c1}(t_2)]e^{-t/T} -$$

$$\left[(aC - i_1)\frac{\alpha_0}{1-\alpha_0} - i_{c1}(t_2)\right]\exp\left[-\frac{1-\alpha_0}{T}t\right] \ldots\ (10)$$

Stage 4

When the current in D_3 has been reduced to zero this diode cuts off. There are then no diodes in circuit except D_7. V_{b1} is free to rise in potential and the total current through the structure can increase rapidly as heavy regeneration is present. With values of supply currents as normally used the increase in structure current causes D_7 to switch off ($i_{e2}=i_7$) before V_{b1} rises to zero volts causing D_4 to conduct. Stage 4 therefore comprises the time taken for i_{e2} to increase from its value at the end of stage 3 (which must be $aC + i_3 - i_1$) to i_7.

We have $i_{e2} = i_{e1} + i_3 - i_1$

$$i_{e1} = pC(V_{b1} - V)$$
$$V_{b1} = (i_{x1}/pkC)$$
$$i_{x1} = i_{c2} - i_3 - i_{b1}$$
$$i_{c2} = \alpha i_{e2}$$
$$i_{b1} = (1 - \alpha)i_{e1}, \text{ from which}$$
$$i_{e2} = \frac{i_3(k - \alpha) - VpkC - i_1(k + 1 - \alpha)}{k + 1 - \alpha}$$

Writing $\alpha = \dfrac{\alpha_0/T}{p + (1/T)}$, and re-arranging,

$$i_{e2} = \left(p + \frac{k + 1 - 2\alpha_0}{(k + 1)T}\right) - \frac{i_3}{k + 1}\left(p + \frac{k - \alpha_0}{T}\right) + \left(\frac{VpkC}{k + 1}(p + (1/T)) + i_1\left(p + \frac{k + 1 - \alpha_0}{(k + 1)T}\right)\right) = 0$$

Initially $i_{e2} = aC + i_3 - i_1$, $V = 0$, $dV/dt = -a$,

$$\therefore \bar{i}_{e2}\left(p + \frac{k + 1 - 2\alpha_0}{(k + 1)T}\right) - \frac{\bar{i}_3}{k + 1}\left(p + \frac{k - \alpha_0}{T}\right) + \left(\frac{\bar{V}pkC}{k + 1}(p + (1/T)) + \bar{i}_1\left(p + \frac{k + 1 - \alpha_0}{(k + 1)T}\right)\right) = \frac{aC + ki_3}{k + 1}$$

$$\bar{i}_3 = (i_3/p), \quad V = (a/p^2), \quad \bar{i}_1 = (i_1/p)$$

$$\therefore \bar{i}_{e2}\left(p + \frac{k + 1 - 2\alpha_0}{(k + 1)T}\right) - \frac{i_3}{k + 1}\left(p + \frac{k - \alpha_0}{T}\right) - \left(\frac{akC}{k + 1}(p + (1/T)) + i_1\left(p + \frac{k + 1 - \alpha_0}{(k + 1)T}\right)\right) = p\,\frac{aC + ki_3}{k + 1}$$

$$\therefore \bar{i}_{e2} = \frac{kaC + i_3(k - \alpha_0) - i_1(k + 1 - \alpha_0)}{p(k + 1 - 2\alpha_0)} + \frac{aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1}{(k + 1 - 2\alpha_0)\left(p + \dfrac{k + 1 - 2\alpha_0}{(k - 1)T}\right)}$$

$\therefore i_{e2} =$

$$\frac{kaC + i_3(k - \alpha_0) - i_1(k + 1 - \alpha_0)}{k + 1 - 2\alpha_0} + \frac{aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1}{k + 1 - 2\alpha_0} \exp\left[-\frac{k + 1 - 2\alpha_0}{k + 1} \cdot t/T\right] \quad \ldots\ldots\ldots (11)$$

As $\alpha_0 \to 1$ this expression simplifies to

$$i_{e2} = \frac{aC - i_1}{1 - k}\left(\exp\left[\frac{1 - k}{1 - k}\, t/T\right] - k\right) + i_3 \quad \ldots\ldots (12)$$

which reveals the positive exponential in the current build-up to be expected in a trigger circuit.

During the next stage we shall need as starting conditions both the potential and the rate of change of potential of V_{b1} at the end of the present stage.

From $i_{e2} = i_{e1} + i_3 - i_1$

$i_{e1} = pC(V_{b1} - V)$, and equation (11)

we get

$$V_{b1} = (1/C)\int_0^t \frac{aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1}{k + 1 - 2\alpha_0}\left(\exp\left[-\frac{k + 1 - 2\alpha_0}{k + 1}t/T\right] - 1\right)$$

$$= \frac{aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1}{C(k + 1 - 2\alpha_0)}\left[-\frac{(k + 1)T}{k + 1 - 2\alpha_0}\exp\left(-\frac{k + 1 - 2\alpha_0}{k + 1}t/T\right) - t\right]_0^t$$

$$= \frac{T[aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1]}{C(k + 1 - 2\alpha_0)}\left[\frac{k + 1}{k + 1 - 2\alpha_0}\left(1 - \exp\left[-\frac{k + 1 - 2\alpha_0}{k + 1} - (t/T)\right]\right) t/T\right] (13)$$

As $\alpha_0 \to 1$ this expression simplifies to

$$V_{b1} \to \frac{T(aC - i_1)}{C(1 - k)}\left[\frac{1 + k}{1 - k}\left(\exp\left[-\frac{1 - k}{1 + k}t/T\right] - 1\right) - (t/T)\right]. \quad (14)$$

From equation (13),

$$dV_{b1}/dt = \frac{aC(1 - 2\alpha_0) + i_3(1 - \alpha_0) + \alpha_0 i_1}{C(k + 1 - 2\alpha_0)}\left(\exp\left[-\frac{k + 1 - 2\alpha_0}{k + 1}t/T\right] - 1\right) \quad (15)$$

which reduces to

$$dV_{b1}/dt \to \frac{aC - i_1}{C(1 - k)}\left(\exp\left[\frac{1 - k}{1 + k}\,t/T\right] - 1\right) \ldots\ldots (16)$$

as $\alpha_0 \to 1$.

Stage 5

When, at the end of stage 4, D_7 cuts off, the load nC is exposed and the emitter of X_2 is free to fall in potential. As this emitter is falling, V_{b1} is rising. Stage 5 is defined by the time between the release of D_7 and the turn on of D_4 as V_{b1} reaches ground potential.

We have $V_{b1} = (i_{x1}/pkC) + V_{b2}$

$$V_{b2} = (i_{x2}/pnC)$$
$$i_{x2} = i_{e1} + i_3 - i_1 - i_7$$
$$i_{e1} = pC(V_{b1} - V)$$
$$i_{x1} = i_{c2} - i_3 - i_{b1} = \alpha i_{e2} - i_3 - (1 - \alpha)i_{e1}$$
$$i_{e2} = i_7 + i_{x2}$$

from which can be derived

$$V_{b1} = \frac{pCV[n(1 - 2\alpha) + k] - i_3[n(1 - \alpha) + k] - i_1(n\alpha - k) + ki_7}{pC\,[k(n + 1) + n(1 - 2\alpha)]}$$

Upon substitution for α and re-arranging,

$$V_{b1}pC\left[p[k(n + 1) + n] - \frac{k(n + 1) + n(1 - 2\alpha_0)}{T}\right] - pCV\left[p(n + k) + \frac{n(1 - 2\alpha_0) + k}{T}\right] + i_3\left[p(n + k) + \frac{n(1 - \alpha_0) + k}{T}\right] - i_1\left[pk - \frac{n\alpha_0 - k}{T}\right] - ki_7\,(p + (1/T)) = 0$$

Initially, $V_{b1} =$, say, V_1 ; $dV_{b1}/dt =$, say, b — These values can be computed from equations (11), (13) and (15)

$V = 0$, $dV/dt = -a$,

the right-hand side of the equation thus becomes

$$C[k(n + 1) + n](pV_1 + b) + \frac{C[k(n + 1) + n(1 - 2\alpha_0)]V_1}{T} - aC(n + k) + i_3(n + k) - k(i_1 + i_7)$$

Now $\bar{V} = -(a/p^2)$, $\bar{i}_3 = (i_3/p)$, $\bar{i}_1 = (i_1/p)$, and $\bar{i}_7 = (i_7/p)$, so

$$\overline{V}_{b1}p^2C\left[p[k(n+1)+n] + \frac{k(n+1)+n(1-2\alpha_0)}{T}\right] + aC\frac{[n(1-2\alpha_0)+k]}{T}$$

$$+ i_3\frac{[n(1-\alpha_0)+k]}{T} + i_1\frac{n\alpha_0 - k}{T} - (ki_7/T) =$$

$$pC\,[k(n+1)+n]\,(pV_1+b) + \frac{pCV_1[k(n+1)+n](1-2\alpha_0)}{T}$$

which can be written $\overline{V}_{b1} = \dfrac{p^2V_1 + pd + e}{p^2(p+f)}$

where $d = b + \dfrac{V_1[k(n+1)+n(1-2\alpha_0)]}{[k(n+1)+n]T}$

$$e = \frac{-aC[n(1-2\alpha_0)+k] - i_3[n(1-\alpha_0)+k] - i_1(n\alpha_0 - k) + ki_7}{CT[k(n+1)+n]}$$

and $f = \dfrac{k(n+1)+n(1-2\alpha_0)}{[k(n+1)+n]T}$

$$\therefore \overline{V}_{b1} = (e/p^2f) + \frac{df-e}{pf^2} + \left(V_1 - \frac{df-e}{f^2}\right)\frac{1}{p+f}$$

and

$$V_{b1} = (e/f)\,t + \frac{df-e}{f^2} + \left(V_1 - \frac{df-e}{f^2}\right)e^{-ft} \quad \ldots . (17)$$

As $\alpha_0 \to 1$ this expression can be written

$$V_{b1} \to V_1 + \frac{[(aC-i_1)(n-k)+k(i_7-i_3)]T}{C[k(n+1)-n]}\cdot(t/T) +$$

$$+ \frac{T[k(n+1)+n]}{k(n+1)-n}\left[b - \frac{(aC-i_1)(n-k)+k(i_7-i_3)}{k(n+1)-n}\right]$$

$$\left(1 - \exp\left[-\frac{k(n+1)-n}{k(n+1)+n}\cdot(t/T)\right]\right) \ldots (18)$$

When calculating the next stage the values of V_{b2} and dV_{b2}/dt at the end of stage 5 will be needed. The value of i_{e2} throughout the switching period is also of interest as it may rise to a very high value.

For V_{b2}, we have

$$V_{b2} = -(i_{x2}/pnC)$$
$$i_{x2} = i_{e1} + i_3 - i_1 - i_7$$
$$i_{e1} = pC(V_{b1} - V)$$
$$V_{b2} = -\frac{pC(V_{b1} - V) + i_3 - i_1 - i_7}{pnC}$$
$$= -(V_{b1}/n) + (V/n) + (1/nC)\int_0^t i_7 + i_1 - i_3$$

$$\therefore V_{b2} = -(V_{b1}/n) + (T/nC)\,(i_7+i_1-i_3-aC)\,t/T \quad \ldots . (19)$$

where V_{b1} may be obtained from equation (17).

For dV_{b2}/dt, we have, differentiating equation (19)

$$(dV_{b2}/dt) = -(dV_{b1}/ndt) + (1/nC)\,(i_7+i_1-i_3-aC)$$

From equation (17),

$$(dV_{b1}/dt) = (e/f) - f\left(V_1 - \frac{df-e}{f^2}\right)e^{-ft}$$

$$\therefore dV_{b2}/dt = f\left(V_1 - \frac{df-e}{f^2}\right)$$
$$e^{-ft} - (e/f) + (1/nC)(i_7+i_1-i_3-aC) \quad \ldots . (20)$$

Now $V_{b2} = -(i_{x2}/pnC)$

$\therefore i_{x2} = -nC\,(dV_{b2}/dt)$

Also, $i_{e2} = i_7 + i_{x2}$

$$\therefore i_{e2} = i_7 - nCf\left(V_1 - \frac{df-e}{f^2}\right)e^{-ft} + (nCe/f) - i_7 - i_1 + i_3 + aC$$

$$= (nCe/f) - nCf\left(V_1 - \frac{df-e}{f^2}\right)e^{-ft} - i_1 + i_3 + aC \quad (21)$$

Stage 6

At the end of stage 5, D_4 conducts, thereby placing an effective short-circuit between X_1 emitter and ground and also arresting the rise of V_{b1}. At this time switching is completed since the input waveform can be removed without harm. It remains for the potential V_{b2} to fall from that given by equation (19) at time t_5 to ground potential.

We have

$V_{b2} = -(i_{x2}/pnC)$	:	$i_{x2} = i_{e2} - i_7$
$i_{e2} = (i_{b2}/(1-\alpha))$	:	$i_{b2} = i_{c1} - i_1 - i_{x1}$
$i_{c1} = (\alpha/(1-\alpha))\,i_{b}$	:	$i_{b1} = i_{c2} - i_3 - i_{x1}$
$i_{c2} = \alpha i_{e2}$	:	$i_{x1} = -pkCV_{b2}$

Combining these expressions,

$$V_{b2} = \frac{i_7(1-2\alpha) + \alpha i_3 + i_1(1-\alpha)}{pC[n(1-2\alpha)+k]}$$

Upon substituting for α, this becomes

$$V_{b2}p\left[p + \frac{n(1-2\alpha_0)+k}{(n+k)T}\right] - \frac{i_7}{C(n+k)}\left(p + \frac{1-2\alpha_0}{T}\right) -$$

$$\frac{\alpha_0 i_3}{CT(n+k)} - \frac{i_1}{C(n+k)}\left(p + \frac{1-\alpha_0}{T}\right) = 0$$

Initially $V_{b2} =$, say, V_2, from equations (17) and (19)

$dV_{b2}/dt =$, say, v_2, from equation (20)

Substituting these initial conditions,

$$\overline{V}_{b2}p\left[p + \frac{n(1-2\alpha_0)+k}{(n+k)T}\right] - p\frac{\overline{i}_7+\overline{i}_1}{C(n+k)} - \frac{\overline{i}_7(1-2\alpha_0)+\alpha_0\overline{i}_3+\overline{i}_1(1-\alpha_0)}{CT(n+k)} =$$

$$pV_2 + v_2 + \frac{n(1-2\alpha_0)+k}{(n+k)T}V_2 - \frac{i_7+i_1}{C(n+k)}$$

Now $\overline{i}_7 = (i_7/p)$, $\overline{i}_1 = (i_1/p)$, $\overline{i}_3 = (i_3/p)$

$$\text{So } \overline{V}_{b2}p^2\left[p + \frac{n(1-2\alpha_0)+k}{(n+k)T}\right] - \frac{i_7(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)}{CT(n+k)} =$$

$$p^2V_2 + p\left(v_2 + \frac{n(1-2\alpha_0)+k}{(n+k)T}V_2\right)$$

whence

$$\overline{V}_{b2} = (V_2/p) + \frac{T(n+k)}{n(1-2\alpha_0)+k}\left[v_2 - \frac{i_7(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)}{C[n(1-2\alpha_0)+k]}\right]$$

$$\left[1/p - \frac{1}{p + \dfrac{n(1-2\alpha_0)+k}{(n+k)T}}\right] + \frac{i_7(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)}{p^2C[n(1-2\alpha_0)+k]}$$

the transform of which is

$$V_{b2} = V_2 + \frac{T(n+k)}{n(1-2\alpha_0)+k}\left[v_2 - \frac{i_7(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)}{C[n(1-2\alpha_0)+k]}\right] \times$$

$$\left[1 - \exp\left\{-\frac{n(1-2\alpha_0)+k}{(n+k)T}\,t/T\right\}\right] +$$

$$\frac{[i_1(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)]T}{C[n(1-2\alpha_0)+k]}\,t/T \quad \ldots\ldots\ldots\ldots (22)$$

which, as $\alpha_0 \rightarrow 1$, reduces to

$$V_{b2} = V_2 - \frac{T(n+k)}{n-k}\left[v_2 - \frac{i_1-i_3}{C(n-k)}\right]\left[1 - \exp\left\{-\frac{n-k}{n+k}t/T\right\}\right] + \frac{i_1-i_3}{C(n-k)}\,t/T$$

The emitter current of the pnp transistor during this stage is given by $i_{e2} = i_1 - nC\,(dV_b/dt)$ which, from equation (22) becomes

$$i_{e2} = i_1 - n\left[\frac{i_1(1-2\alpha_0)+\alpha_0 i_3+i_1(1-\alpha_0)}{n(1-2\alpha_0)+k}\left(1 - \exp\left\{-\frac{n(1-2\alpha_0)+k}{n+k}\,t/T\right\}\right) + v_2 C\exp\left\{-\frac{n(1-2\alpha_0)+k}{n+k}\,t/T\right\}\right] \quad \ldots\ldots\ldots\ldots (23)$$

which, as $\alpha_0 \rightarrow 1$, reduces to

$$i_{e2} = i_1 - n\left[\frac{i_1-i_3}{n-k}\left(1 - \exp\left\{-\frac{n-k}{n+k}t/T\right\}\right) + v_2 C\exp\left\{\frac{n-k}{n+k}t/T\right\}\right]$$

If equations (22) and (23) are computed using typical values for the various parameters it is found that the current through the structure can be as high as several hundreds of milliamperes. The presence of base spreading resistance will cause bottoming of one or both transistors to ensue. If the value of base spreading resistance is known the potential at which bottoming will commence can be predicted.

Once bottoming occurs the operation of the circuit is modified. Several approaches are possible to calculate the ensuing potentials and currents, but one only will be treated here. This is the easiest case: it assumes both transistors bottom simultaneously when the potential across the structure is at V_3 volts.

The circuit then degenerates to a capacitor with stored charge nCV_3 across which is a resistance of value $r_b'/2$. (This assumes the voltage drop $\frac{(i_1 - i_3)r_b'}{2}$ can be neglected).

The potentials V_{b2} then falls as

$$V_{b2} = V_3 \exp[-(2t/nCr_b')] \quad \ldots\ldots\ldots\ldots (24)$$

and the current through the emitter of the pnp transistor is given by

$$i_{e2} = i_1 + (2V_3/r_b')\exp[-(2t/nCr_b')] \quad \ldots\ldots (25)$$

(To be continued)

Vehicle Speed Measuring Equipment

An electronic vehicle speed measuring equipment incorporating fifty-four transistors has been developed by Venner Electronics Limited, and is at present receiving consideration by the Metropolitan Police who are testing the device on outer London roads.

The equipment was primarily developed for the Road Research Laboratories of the D.S.I.R. and is basically a device for measuring small intervals of time to a high degree of accuracy. Attached to the electronic elapsed time indicator are two rubber tubes long enough to span the road, each of which is connected to a pressure switch. The two tubes are layed across the road at right-angles to the traffic flow and spaced a little under six feet apart (70·4in).

To measure the speed of an approaching vehicle all that is required is to set the equipment to 'Ready'. Immediately the front wheels of the vehicle cross the first tube the pressure switch starts the timer. As soon as the wheels cross the second tube the timer stops. The elapsed time is recorded by the meters. The reading obtained remains on the meters until the equipment is manually reset.

It is expected that the rubber tubes may be laid semi-permanently at points known to be black spots and the equipment plugged in from time to time which would discourage motorists from exceeding speed limits and so contribute to road safety.

The equipment incorporates a crystal oscillator operating at 10kc/s as the basic time element. The output of this unit is passed to two binary dividers to give a 2·5kc/s signal which is in turn fed to a gating circuit. The gate is opened by the passing of a vehicle over the first strip and closed by the vehicle passing over the second strip. The number of pulses passing through the gate while it is opened are counted by three decade counting stages with digital presentation on meters calibrated from 0-9. If, for example, the first meter indicates 3, the second 4 and the third 5, then this is the interval of time in 0·4msec steps which elapsed while the vehicle traversed the measured distance of 70·4in. The frequency and distance are chosen so that the resultant indication may be divided into 10 000 giving the speed of the vehicle. Thus, in the case given the speed corresponds to 29 mile/h. The accuracy is ± 1 count for any figure recorded and thus at 30 mile/h (indication 333) the accuracy is to approximately $\pm\frac{1}{3}$ per cent. At 100 mile/h (indication 100) the accuracy is to within ± 1 per cent. The equipment is completely portable, batteries being included in the case which measures 13in by 9in by 7in.

The speed measuring equipment in use.

An Electronic System for Bus Running Control

An electronic system designed to aid London's buses to overcome the effects of traffic hold-ups has recently been demonstrated by London Transport.

The device is known as "BESI" (Bus Electronic Scanning Indicator) and will enable each bus on a route to transmit an identification number.

The overall view of the buses gained by the electronic method will enable the controllers to take quick remedial action when a route's regularity is upset through traffic.

The main difficulty in operating a regular bus service through the congested streets of a city is the lack of knowledge where individual buses are at any instant of time, and whether any acute congestion has built up in any particular area, but to give a complete picture of bus operation on London Transport entails the tracing of some 7 000 buses operating on some 500 routes.

A variety of schemes have been tried out, including magnetic scanning and local radio intercommunication, but bearing in mind the cost involved to cover the whole of the London Transport area, a simple optical system has finally been chosen.

The scheme which is now in process of development consists of a number of check points along each route which, by electronic equipment, reads the number of each bus as it passes the point and transmits it over telephone wires to a central location, where it is displayed on a panel and finally will be recorded on a chart. When each bus passes the next check point the number will be recorded on another panel and cancelled on the first, so that in effect each panel will contain the running number of each bus between the two boundary check points.

Numerically, the largest item will be the equipment necessary on each of the 7 000 buses, and for this reason this item has been designed so as to be as cheap as possible; moreover, equipment on a moving bus is difficult to maintain so that it was decided only a form of reflecting plate would be suitable.

Some of the other more important factors that had to be considered in the design of the system were as follows:

(a) The equipment designed to facilitate quick and easy changes of route and running number.

(b) The bus equipment in no circumstances to interfere with the ordinary functioning of the bus, or its crew.

(c) The road-side equipment easily installed and maintained.

(d) The System capable of being operated by day and by night, over a range of bus speeds from one to thirty miles per hour.

General Description

The pilot system is now being tried out in service on a bus route which has been divided into four sections each with a scanning equipment for reading in either direction.

Each bus running number has been coded into a binary number and this has been reproduced on a plate containing reflectors for each digit, and a corresponding blank for each zero. The plate has been placed on the nearside of the bus towards the front, above the driver's cab, as shown in Fig. 1.

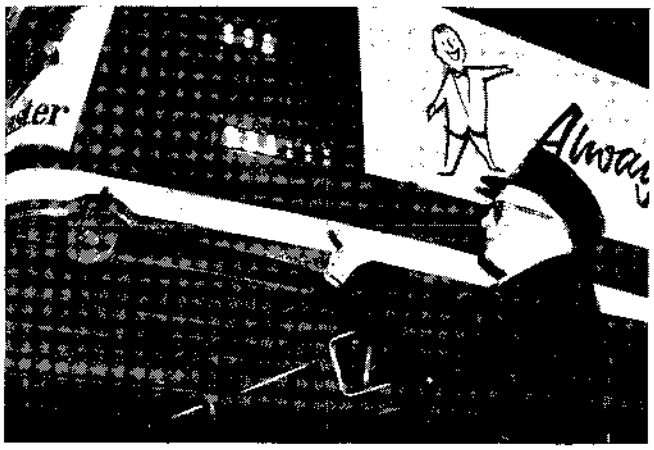


Fig. 1. The reflectors mounted above the driver's cab

The scanning equipment has been erected at the same height as the bus plate and as near as possible to the roadway, either on a post at the pavement edge, or where the side-walk is narrow, on a convenient point on the building line.

The equipment consists of a source of modulated light focused on to the plates of the buses, and a further optical system to receive back the reflected light. As each reflector passes the beam of light, the reflected light falls on to a photo-electric cell and the resultant pulses are amplified shaped and transmitted by line to the control point.

At the control point, the incoming pulses operate relays which will build up the binary number of the bus, and when completed, will close the circuit, causing the actual running number to be illuminated on the check point display panel.

The panels of each check point are so wired that as each number comes up on one panel, the number on the previous panel is cancelled. The general layout is shown diagrammatically in Fig. 2.

Bus Equipment

As stated above, the equipment mounted on the bus consists of a plate containing reflectors representing the running number of the bus in binary form.

There are several problems to be overcome in designing the plate. Firstly, there is the difficulty of recording buses travelling at any speed between one and thirty miles per hour; secondly, there is the variation in the height of the reflectors, due to the loading of the bus, and, thirdly, there is the variation of angle of the reflectors from the vertical due to the position of the bus on the camber of the road as it passes the scanner. Added to the above is the requirement that the plates must be of reasonable size, not only for handling purposes—as they are changed every time a bus alters its running number—but also because there is little space on the bus not already occupied by destination blinds, advertising spaces and windows. Moreover, the plates must be placed at a height, well above the level of pedestrians on the side walk and, maybe, cars parked at the kerb.

The most suitable form of reflector has been found to be the type used for road signs, generally referred to as 'cats eyes'. These are $\frac{5}{8}$in in diameter and can be designed for a very narrow reflecting field and negligible scatter.

To compensate for varying speeds of the bus, the plate is produced in two parts, the upper half being a complete row of reflectors to form a timing base—plus one additional one at the end to act as a clear down unit. The lower half is placed 10in lower, and exactly in line vertically is the binary code plate. The horizontal spacing between each reflector is 2in, which has been found satisfactory for timing purposes up to at least forty miles per hour, leaving a reasonable margin for normal bus operation.

To compensate for both loading and camber, the reflectors are mounted in pairs, one above the other at 1in centres.

As previously stated, these plates are fitted towards the front of the bus on the nearside, above the driving cab, and are fitted into slots, and in this position will not interfere with the mechanical washing equipment nor with normal operation.

Scanning Equipment

At each check point a scanning equipment must be installed at each side of the road to register the flow of buses. The scanner unit is mounted so that the buses pass

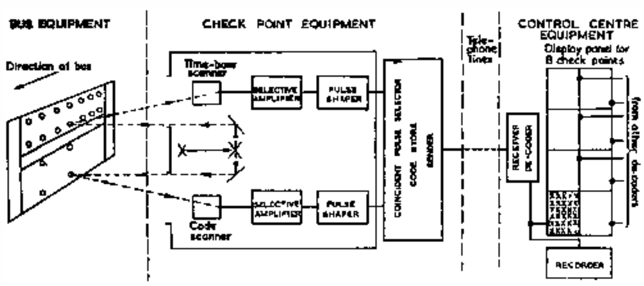


Fig. 2. General arrangement of the equipment

within about 12ft of it, as above this distance the angular displacement of the light becomes increasingly difficult to accommodate. In addition, if the spacing from the kerb becomes too great, there is the danger of a parked lorry completely masking the scanner. Sites, therefore, are selected where the road is narrow, or its width restricted due to a pedestrian island. As the scanner unit itself may need to be pole mounted, this is made as small and light as possible. The relays and other circuit equipment are mounted at suitable positions as near as possible, either by space within a building or, if necessary, in a street box mounted on the pavement.

The scanner is designed to reject all signals other than those contained in the reflectors, particularly the gold lettering and other reflecting vertical lines often seen on trade vehicles.

It must also work in bright sunlight as well as on dark nights—it is also hoped that it will operate so long as the buses themselves can be operated in a fog.

The scanner (shown in Fig. 3) consists of a light source consisting of a 12V, 36W lamp, which is focused by lenses and passed through slots on a rotating disk, the number of slots and the speed of its rotation results in the light being modulated at 3kc/s. These light pulses are passed to two mirrors set at right-angles, which divide the beam into two, and by further mirrors deflected one on to the time-base plate and the other on to the coding plate on the bus itself.

The light reflected from these plates is passed into the scanner once more and focused on to two photo-sensitive transistors which transform the light pulses into electrical pulses which are then passed to resonant amplifiers. These reject the effect of all extraneous incoming light not associated with that sent out by the instrument's lamp at the controlled frequency, and therefore not part of the bus identification.

Neither the 3kc/s modulation nor the waveform of the resulting signals would be suitable for operating the telephone-type circuits which follow, so that they are passed into pulse shaping transformers where a square wave shaped pulse is produced from each signal from the reflectors.

The pulses from the time-base plate are now paired with those of the code plate by feeding them to cold-cathode valves, and when both time-base and code valves fire together, relays are operated, thus storing the code so formed.

As at least two scanners, and maybe more in busy traffic centres, will share the telephone line to the control centre, the code is stored by the relays mentioned above.

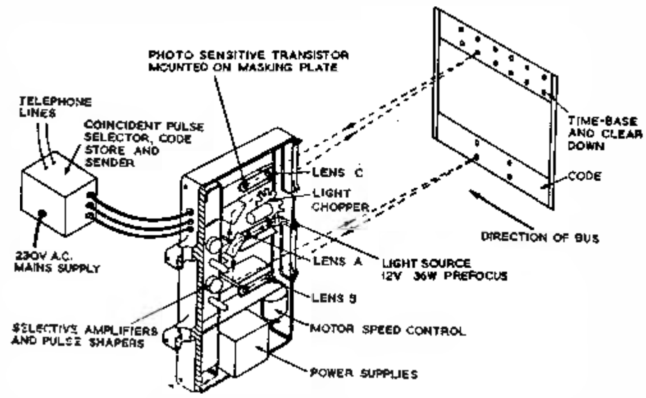


Fig. 3. Arrangement of the scanner

The stored binary number at the scanning point is fed into the line, as soon as a clear channel is available by normal pulsing procedure, and the pulses are again stored on relays at the centre, and when built up, energize the lamp with the running number corresponding to the code.

Control Centre Equipment

The control centre will have a series of panels, one for each scanner position. These panels will have a display of lamps, each lamp having printed on it the running number of a bus operating on the route being scanned. As each bus progresses from scanning point to scanning point, so the lamps will light, signifying between which pair of scanners the bus is located at any instant of time.

In addition to lighting the indicating lamp, a recording pen mark can be made on a chart moving forward on a scale of time, which forms a permanent record of the time each bus enters the section, and hence the regularity of the service; alternatively, the information can be recorded in punched tape form.

The characteristic of the final relay illuminating the lamp is designed so that if two lamps are in circuit at the same time, as when a bus passes from one scanning zone to another, the current demand causes the voltage to drop and the previously energized relay drops off, leaving the subsequent relay in charge and only the single up-to-date indication left on the appropriate panel.

The equipment is in its primary development stage and there will no doubt be many improvements to be made over the next few months, but it is expected that it will at least enable some improvement in service regularity to be achieved in spite of the extremely difficult conditions which have arisen as a result of street congestion.

The device is covered by provisional patent application No. 33758/57. The whole of the equipment has been designed and developed at the Executive's Electrical Testing Section at Wood Lane, under the direction of Mr. T. S. Pick, the Chief Electrical Engineer.

A New Electronic Assembly System

By H. C. Bertoya*, A.M.Brit. I.R.E.

MANY new mass production techniques have been evolved for electronic and allied equipments over the last few years; but the method of construction of apparatus in laboratories has hardly altered. The usual method is, of course, to take a sheet of aluminium and to make an intelligent guess at the number and position of holes likely to be required. The metal work is then carried out and the sheet is bent to form either a box or a channel; the sides of the box serving to protect the components when the box is laid on the bench. Usually more holes have to be made after the main components have already been attached. If this is impossible, components have to be removed so that the metal work may be carried out. The chassis is, however, unusable in its normal horizontal position since the components are not accessible. Because of this the chassis is often stood on end, or propped up against a convenient text-book, with the attendant danger to valve top-caps, etc., when the chassis slips over. If a circuit spreads over the confines of the area available, 'Christmas-trees' of components appear which are mechanically very unstable. If the device is fairly complex, either a great number of functional units must be built on one chassis (which tends to make confusion worse confounded) or numbers of small chassis are arrayed on the bench top. In general the normal methods are unsatisfactory and a great deal of design time is spent, not in improving or testing the electrical circuit, but in carrying out makeshift repairs to a ramshackle set-up. The obvious waste of time in the writer's own laboratory caused him to reconsider constructional techniques and to design a method which would aid the construction of circuits and complete electronic systems. It is now being marketed under the name of 'Rapikon.'

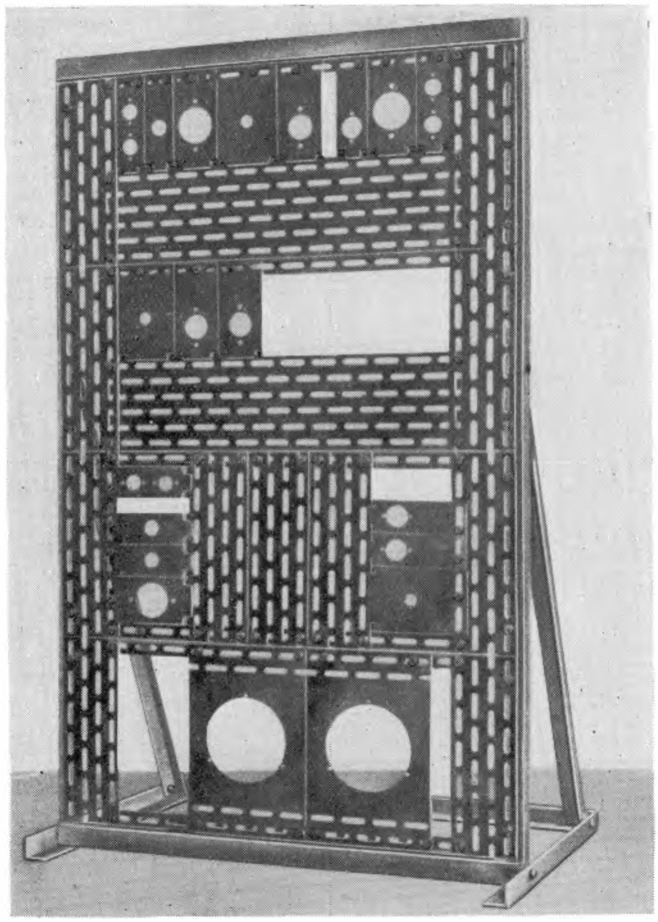
Existing Methods of Unit Construction

An aid to rapid construction is the 'unit' system. Some systems have been described[1,2] which have admirable features. Their 'box-chassis/front-panel' approach was, however, rejected. The fundamental problem is not that of constructing a conventional chassis by means of standard parts. The problem is, rather that of finding the best means of supporting circuit components, having regard to the purpose which that circuit has to serve.

As the design proceeded it was soon apparent that many compromises would have to be made between the conflicting requirements of temporary and permanent equipment. It was, therefore, decided to concentrate solely on facilitating the experimental construction and proving of circuits and systems, although this does not prevent the construction of permanent equipment if it is thought necessary.

In fact, there is no great virtue in saving a day's construction time in a 'one-off' equipment if it is to serve for a couple of years; and the price that is paid for a unit or 'modular' system (expense, space and lack of complete flexibility) is unnecessarily high for production models. In any case, the important thing is to develop a circuit which functions well. Problems of mechanical construction are more properly the province of the production engineer and can follow on only after the successful completion of the electrical development stage.

* *Shandon Electronics Ltd.*



A typical 'chassis' assembly.

Requirements of a Desirable Method

After experimenting with several methods of layout and construction, the most important requirements were found to be as follows:—

(1) Circuits and their supports should be capable of modification and growth exactly in accordance with the evolution of the design in the engineer's mind.
(2) Complete systems as well as single circuits should be accommodated, preferably as 'block diagrams'.
(3) The working surfaces should be disposed in such a manner as to provide maximum accessibility and to allow the engineer to work comfortably.
(4) Metal-work should not have to be done.
(5) The form of mechanical layout should be such as to encourage neat and orderly construction. This is an important—though not often considered—point, and one which is less likely to produce wrong answers for irrelevant electrical reasons (bad or mistaken connexions, incorrect components, unnecessarily long leads, etc.)

The New Method of Construction

The method of construction finally developed consists of the following component parts:—

THE HOLDER PLATES

These are specially punched to hold potentiometers, valveholders, switches, terminals, plugs and sockets, meters, etc. They are made from 20 s.w.g. plated mild steel. Each holder plate is movable in the plane of the chassis and has a pronged end for sliding on to a 'slotted plate'. The other end is fixed with a nut and bolt to ensure good electrical connexion. The pronged end means that direct access to the point of fixing is not required: holder plates may be removed and replaced without disturbing overlapping wiring or tag panels, etc. Aluminium holder plates are also provided for special components for which the punched plates are not available.

THE SLOTTED PLATES AND GIRDERS

These may be assembled in a variety of ways, in general forming a rectangular frame 14½in wide. The plates are also used to take components for which the special holder plates are not required, e.g. tag panels, terminal blocks, etc., etc. The slotted plates and girders are assembled in such a way as to leave constant-width gap in a convenient position. Into this gap are inserted the holder plates.

The largest slotted plate is 3in wide; a large quantity of tag panels and components were measured and it was found that this width would accommodate the majority of them. The fixing of a plate at right-angles to the main cross-piece enables power inputs to be conveniently connected to terminal blocks.

THE RACK

This is a heavy mild steel rack about 25in high and 15in wide. It has tapped holes in the uprights and it can be screwed to the bench if necessary.

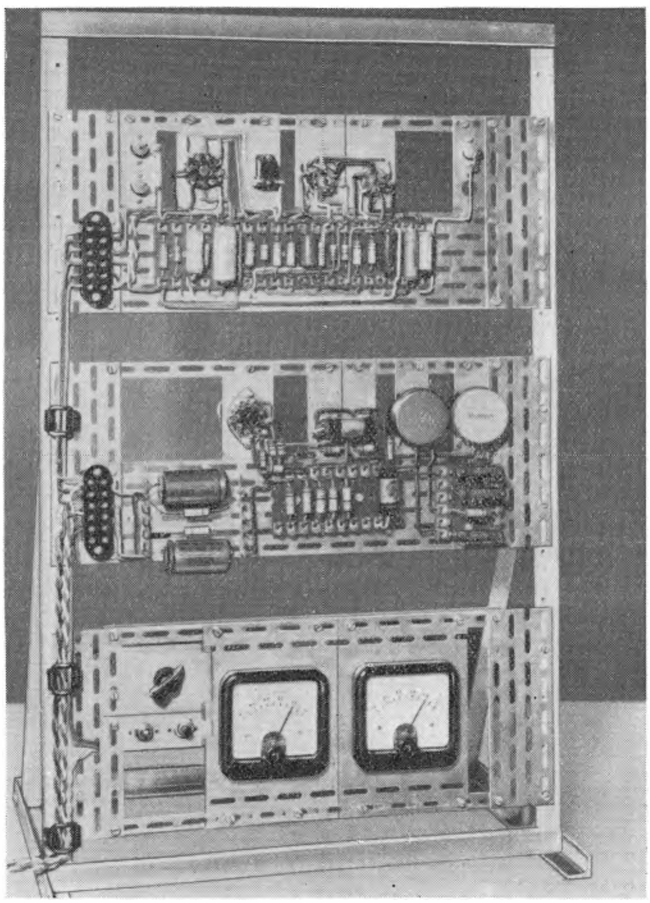
ASSEMBLY

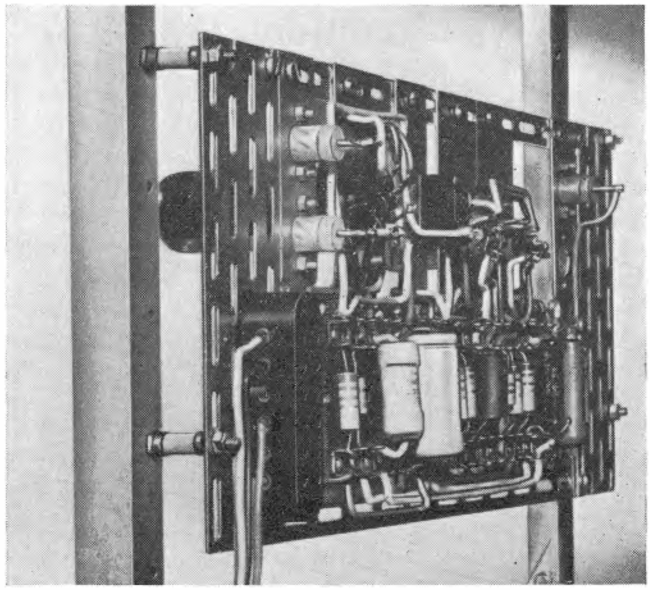
A complete chassis consists of a frame made up from slotted plates and girders, on which are mounted the holder plates. The complete chassis is bolted to the rack.

Conclusion

It is thought that the system conforms fairly closely to the ideal one described under the heading 'Requirements of a desirable method'. Using the same paragraph numbers, the way in which the requirements are satisfied is discussed, as are also some additional advantages which result.

A complete unit.





A unit mounted with components and wiring points facing the front. Since this particular chassis was at an elevated potential, it was mounted on insulators.

(1) Modifications are simple since the 'holes' (i.e. the holder plates) are movable. Chassis may be increased in size by simply altering the size of members or by bolting on extensions. New holder and other plates are to be made as the need arises.

(2) The 14½in width of the basic chassis was the result of experiment. It is the optimum size for most circuits without being too large to prevent the construction of single functional units, e.g. *RC* oscillators, time-bases, gates, etc. Each chassis can form a block of the system block diagram and may be quickly removed and replaced. An improved version of a particular block may be constructed separately and inserted into the system, thus causing minimum disruption of work on the system as a whole.

(3) The working surfaces are held in the vertical plane by the rack. The rack is heavy enough to allow work to be done without shifting. Chassis are held firmly and compactly—the vertical plane means a saving in bench space since the chassis are stacked—and flying leads between chassis are short and do not become displaced (one chassis cannot move in relation to another). When a circuit has been developed and found to be satisfactory there is no cause further to refer to the wiring points which face the experimenter. The chassis may then be reversed on the rack so as to bring all controls to the front.

(4) Metal work is eliminated; in addition there is no scrap since the items may be re-used.

(5) The standard of 'mock-up' construction has certainly improved since the introduction of the system.

Although specifically designed for experimental work, semi-permanent laboratory test gear has been constructed on Rapikon. When not in use it is stored by sliding into special holders. The system has been in use for some time and has proved extremely satisfactory.

Acknowledgment

The writer thanks Messrs. Shandon Electronics Ltd for permission to publish this article.

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A 1kc/s Junction Transistor T-Parameter Measurement Set

By R. A. Hall*, B.Sc.

A set has been constructed for measurement of the parameters of the simple T-equivalent circuit of a junction transistor at a frequency of 1kc/s. A description of the set is given, together with the experimental procedure.

The simple equivalent circuit at 1kc/s is discussed. Experiments so far indicate that serious frequency-dependent errors in measurement of $r_c(1-\alpha)$, $C_c/(1-\alpha)$, and, to a lesser extent, r_e are likely to occur in the case of many low frequency transistors, but not generally in the case of transistors with f_{α} in the megacycle region. These errors are caused by omission of certain reactive elements from the equivalent circuit and, in the case of r_e, the fact that the inner end of capacitor C_c is connected not to the internal node as in the simple circuit, but to some point along the base resistance. The errors can be avoided by reduction of the measurement frequency to about 200c/s.

THE set, together with certain additional equipment, consists essentially of three circuits which are shown in Figs. 1, 2 and 3 with bias supplies connected as for a pnp transistor and will be referred to as Circuits 1a, 1b and 2 respectively. In Figs. 1 and 2 the transistor is represented by the appropriate form of the simple T-equivalent circuit whose parameters are measured and a discussion is given later on the validity of this equivalent circuit.

The measurements which can be made using each of these circuits are as follows:

Circuit 1a r_c, C_c, r_b.

Circuit 1b $r_c(1-\alpha)$, $C_c/(1-\alpha)$, r_e where $\alpha = \alpha_{ce}$

Circuit 2 α_{cb} or $\alpha/(1-\alpha)$

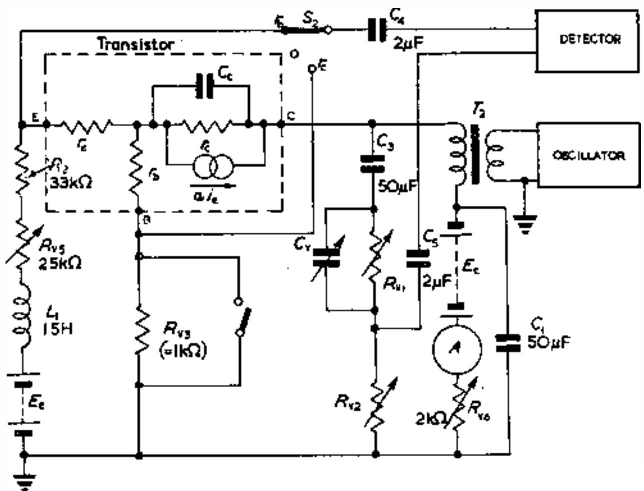

Fig. 1. Circuit 1a

A description of the three measurement circuits is given together with the experimental procedure in each case and this is followed by a description of the complete circuit of the measurement set.

CIRCUIT 1A.

The transistor is placed in a bridge circuit and the impedance in the emitter circuit is very high so that when an alternating voltage is applied in series with the base-collector bias voltage the effect of the internal current generator αi_e due to feedback current in the emitter circuit may be neglected. It is estimated that the reduction in output impedance due to this effect should be less than 5 per cent for most transistors so that the emitter may be considered to be on open-circuit.

The procedure is to set switch S_2 to r_c, R_{v3} to 1kΩ and R_{v2} to 10Ω, then a balance is obtained by adjustment of R_{v1} and C_v. The basic circuit is then as shown in Fig. 4 for which, neglecting r_b in comparison with r_c and the reactance of C_c, the balance conditions are

$$r_c/1000 = (R_{v1}/10) \text{ or } r_c = 100 \,.\, R_{v1} \quad \ldots\ldots \text{(1)}$$

and similarly

$$C_c = (1/100) \,.\, C_v \quad \ldots\ldots\ldots\ldots\ldots \text{(2)}$$

but the variable capacitor must be calibrated against known values of capacitance in place of a transistor in order to allow for residual capacitances.

Next R_{v3} is reduced to zero and switch S_2 set to r_b and a second balance obtained by adjustment of R_{v2} and C_v but leaving R_{v1} at its previous setting. The basic circuit is then as shown in Fig. 5, for which the balance condition is

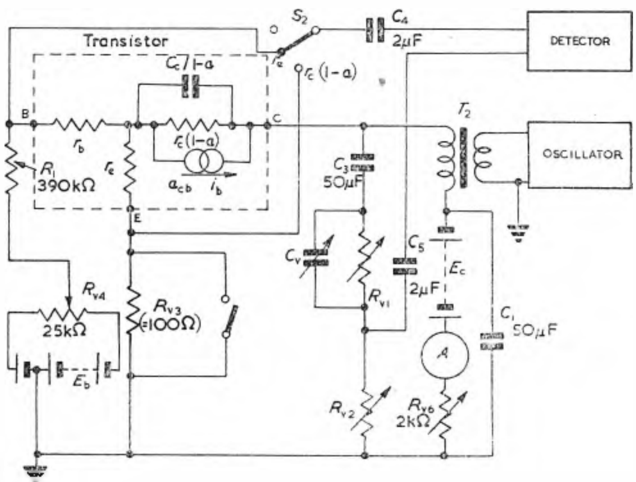

Fig. 2. Circuit 1b

$$r_c/r_b = (R_{v1}/R_{v2})$$

and by substituting for R_{v1} from equation (1)

$$r_b = 100 \,.\, R_{v2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(3)}$$

CIRCUIT 1B

The transistor is placed in a bridge circuit similar to Circuit 1a except that the alternating voltage is applied in series with an emitter-collector bias voltage and the impedance in the base circuit is very high so that the effect of the internal current generator $\alpha_{cb} i_b$ due to feedback current in the base circuit may be neglected. It is estimated that the increase in output impedance due to this effect should not exceed 5 per cent for most transistors so that the base may be considered to be on open-circuit.

The procedure is to set switch S_2 to $r_c(1-\alpha)$, R_{v3} to 100Ω and R_{v2} to 100Ω then a balance is obtained by adjustment of R_{v1} and C_v. The basic circuit is then as shown in Fig. 6 for which, neglecting r_e in comparison with

* *Standard Telecommunications Laboratories Ltd.*

$r_c(1-\alpha)$ and the reactance of $C_c/(1-\alpha)$ the balance conditions are

$$\frac{r_c(1-\alpha)}{100} = R_{v1}/100 \text{ or } r_c(1-\alpha) = R_{v1} \quad \ldots\ldots (4)$$

and similarly

$$C_c/(1-\alpha) = C_v \quad \ldots\ldots\ldots\ldots\ldots (5)$$

but again allowance must be made for residual capacitances.

Next R_{v3} is reduced to zero (taking care to re-adjust base bias current by means of R_{v4}) and switch S_2 set to r_e and a second balance obtained by adjustment of R_{v2} and C_v, but leaving R_{v1} at its previous setting. The basic circuit is then

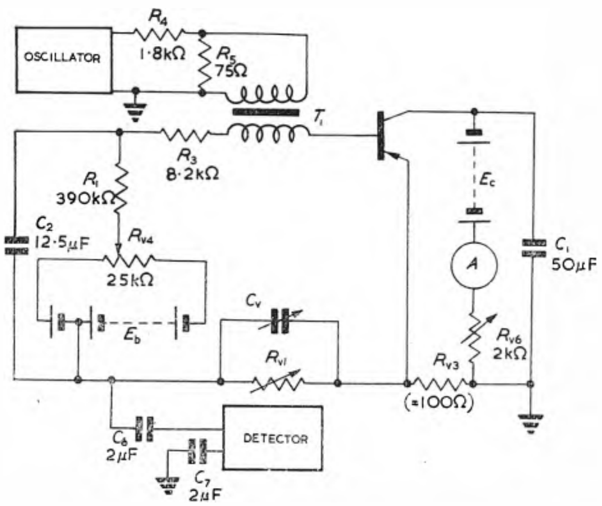

Fig. 3. Circuit 2

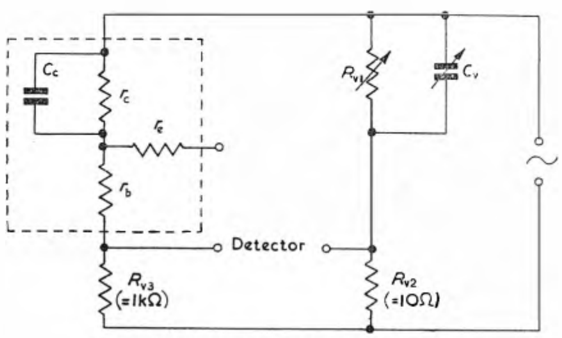

Fig. 4. Basic circuit of Fig. 1 (first balance)

as shown in Fig. 7 for which the balance condition is

$$\frac{r_c(1-\alpha)}{r_c} = R_{v1}/R_{v2}$$

and by substituting for R_{v1} from equation (4)

$$r_e = R_{v2} \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

Circuit 2

The transistor is placed in a common-emitter circuit in which an alternating component is added to the base bias current and a collector load of 100Ω is inserted in order to measure the alternating component of collector current under approximately short-circuit conditions. Due to the approximate 180° phase difference between base current and collector current a null can be obtained in the detector by adjustment of R_{v1} and C_v (taking care to re-adjust base bias current after each large change in the value of R_{v1}). The null condition is

$$\alpha_{cb} = \frac{\text{impedance of combination of } R_{v1} \text{ and } C_v}{100} \quad \ldots (7)$$

In general the phase of α_{cb} is so small at 1kc/s that the effect of the reactance of C_v may be neglected, the exceptions being mainly low frequency transistors with high current gains such as the Brimar TJ3 whose parameters are discussed later.

Complete Circuit

The complete circuit diagram is shown in Fig. 8. The measurement set panel contains Post Office type jacks for

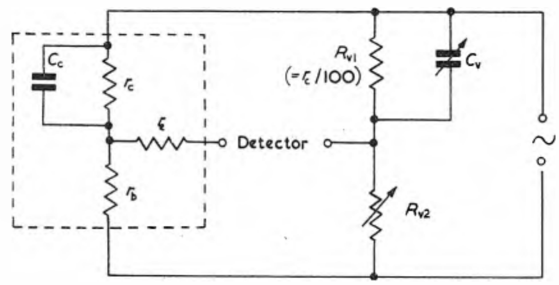

Fig. 5. Basic circuit of Fig. 1 (second balance)

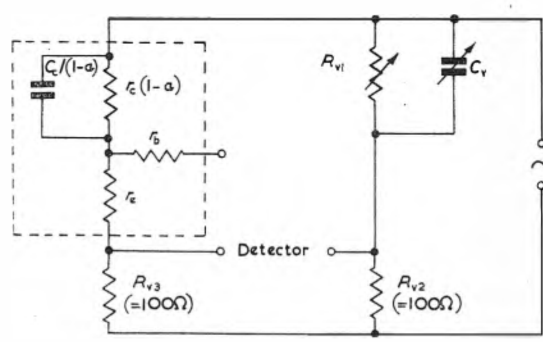

Fig. 6. Basic circuit of Fig. 2 (first balance)

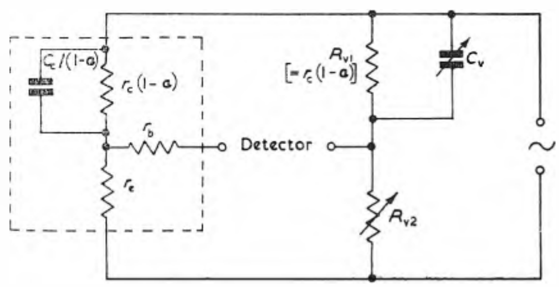

Fig. 7. Basic circuit of Fig. 2 (second balance)

R_{v1}, C_v, generator (GEN) and detector (DET) and also terminals for R_{v2}, external position of R_{v3} switch, bias supplies, collector current meter and the transistor under test.

In order to obtain one of the three measurement circuits it is necessary to insert the R_{v1}, C_v, DET and GEN plugs in the appropriate jacks marked (1) for Circuits 1a and 1b and in jacks marked (2) for Circuit 2. The transistor plug must be inserted in jack J_1, J_2 or J_3 for Circuit 2, 1b, 1a respectively. Switch S_3 is provided as a means of reversing bias supply and meter connexions so that measurements may be made on both pnp and npn transistors and the bias supplies are switched on by setting switch S_1 to the appropriate position.

Adjustment of collector bias current is made by means of R_{v5} in Circuit 1a and R_{v4} in Circuits 1b and 2, and fine adjustment of collector bias voltage is made by means of R_{v6}. The collector bias voltage should be measured with a high impedance voltmeter between base and collector in Circuit 1a and between emitter and collector in Circuits 1b and 2 and in order to ensure that conditions in each circuit are identical it is necessary to measure also the voltage between emitter and base which may be appreciable in some cases.

The additional equipment used is as follows:

R_{v1} 0 to 1MΩ in 10Ω steps.

R_{v2} 0 to 1kΩ in 0·01Ω steps (Sullivan dual-dial resistance box)

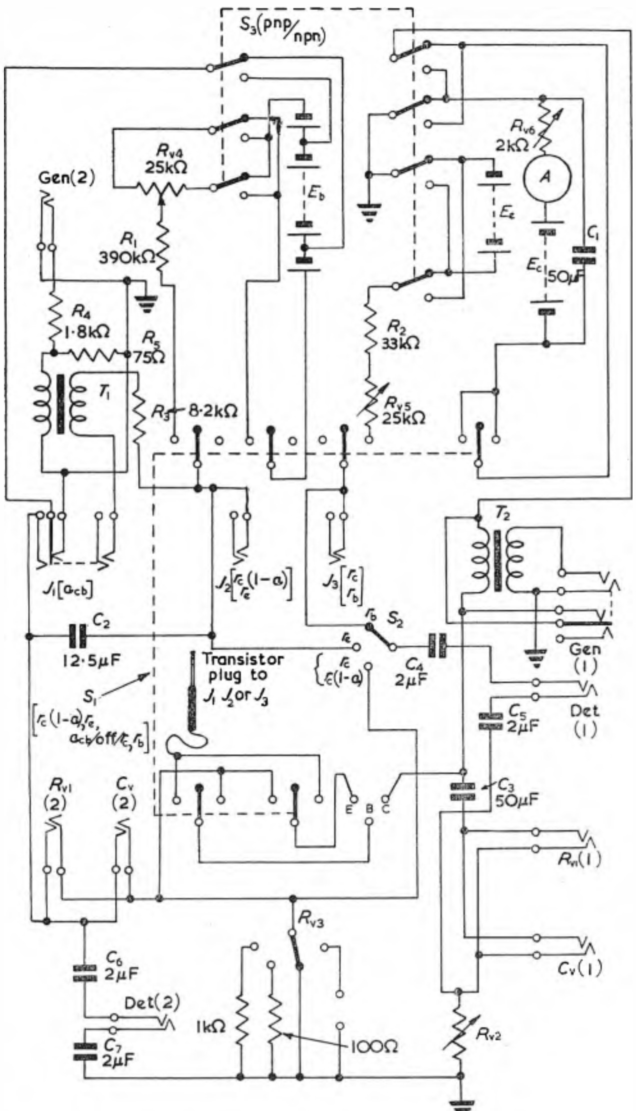

Fig. 8. The complete circuit

C_v Range of 50pF to 0·1μF covered continuously. 50 to 500pF (Variable) (+ about 30pF residual capacitance in plug and leads)

100 to 1 500pF (Variable) }*
0 to 0·1μF in 1 000pF steps (Decade Box) }

* (+ about 100pF residual capacitance with two connected in parallel).

GENERATOR

Ediswan l.f. oscillator having at maximum output an e.m.f. of about 20V peak-to-peak with an output impedance of 10kΩ.

DETECTOR

Tuned transistor differential amplifier as described elsewhere[1].

Transformers T_1 and T_2 are each Fortiphone Type EX45 having a primary inductance of 15H and a turns ratio of 7·1/1 from primary to secondary. Each transformer is connected for voltage step-down from oscillator to bridge so that when the oscillator output is set to maximum the voltage across the bridge arms is about 2·5V peak-to-peak and the base current alternating component in Circuit 2 is about 2μA peak-to-peak.

It is realized that functionally the measurement set could be greatly simplified by replacement of separate plugs and switches by a multi-gang switch. However the set was designed initially not for use by unskilled operators, but for laboratory work in which it is important that measurement equipment should be flexible and easy to modify and it was considered that functional simplifications of that type would not serve these interests.

Errors Due to Incomplete Transistor Equivalent Circuit

At very low frequencies where the frequency dependence of the internal generator of a transistor may be neglected one obtains, for the output admittance in the common emitter connexion with open-circuit base,

$$Y_{out} = \frac{1}{1-\alpha}\,[(1/r_c) + j\omega C_c] \quad \ldots\ldots\ldots (8)$$

neglecting r_e in comparison with $r_c(1-\alpha)$ and the reactance of $C_c/(1-\alpha)$.

For most transistors $f_{c\alpha}$ is not less than 100kc/s, so that the variation of α in magnitude and phase with frequency up to 1kc/s may be represented to a good approximation by

$$\alpha = \frac{\alpha_0}{1 + j\omega/\omega_{c\alpha}} \quad \ldots\ldots\ldots\ldots\ldots\ldots (9)$$

where $\omega_{c\alpha} = 2\pi f_{c\alpha}$

This is the same current which one would obtain by feeding from the current generator through a single RC stage to the collector circuit.

Substituting for α in equation (8)

$$Y_{out} = \left\{\frac{(1 - \alpha_0 + \omega^2/\omega_{c\alpha}{}^2) - j\alpha_0(\omega/\omega_{c\alpha})}{(1-\alpha_0)^2 + (\omega/\omega_{c\alpha})^2}\right\}.\,[(1/r_c) + j\omega C]$$

Hence, neglecting $\omega^2/\omega_{c\alpha}{}^2$ in comparison with $(1-\alpha_0)$

$$Y_{out} = \left\{\frac{(1-\alpha_0) - j\alpha_0(\omega/\omega_{c\alpha})}{(1-\alpha_0)^2 + (\omega/\omega_{c\alpha})^2}\right\}.\,[(1/r_c) + j\omega C] \quad \ldots . (10)$$

The main effect due to the complex nature of α is therefore to cause a reduction of output capacitance and resistance below the very low frequency values given by equation (8), the percentage changes in output capacitance and resistance depending on the values of $\omega/\omega_{c\alpha}(1-\alpha_0)$ and $\omega C_c r_c$ at the frequency of measurement. The measurements in Table 1 made on two Brimar junction transistors indicate the error experienced in practice, although the TJ3 is considered to have a fairly extreme value of $f_{c\alpha}(1-\alpha_0)$.

In Table 2 are given the calculated values of output impedance in Circuit 1b using equation (10) based on measured values of output impedance in Circuit 1a, α_{cb} in Circuit 2 (assumed $-\alpha_0$) and $f_{c\alpha}$.

The error in calculated values of r_{out} and C_{out} is observed to be not greater than about 10 per cent which is considered satisfactory.

In order to verify that the suggested equivalent circuit shown in Figs. 1 and 2 is independent of frequency up to 1kc/s, taking into account the frequency dependence of α as discussed above, measurements were made using a wideband detector at 250c/s and 1kc/s on a group of transistors consisting of the two Brimar types mentioned above and two American types having $f_{c\alpha}$ in the region 5 to 10Mcs. The indication from these few measurements is that over this frequency range the resistive parameters are indepen-

TABLE I

TRANSISTOR	CIRCUIT 1A			CIRCUIT 1B					CIRCUIT 2		
	r_c (MΩ)	C_c (pF)	Time Constant (10^{-6} sec)	r_{out} (KΩ)	C_{out} (pF)	Time Constant (10^{-6} sec)	Apparent α_{cb} (*)	α_{cb}	Phase of α_{cb}	$f_{c\alpha}$ (kc/s)	$f_{c\alpha}(1-\alpha)$ (kc/s)
TJ 1 ..	2·56	35	92	115	650	75	21	21	6°	330	15
TJ 3 ..	9·6	34	330	74	2900	217	129	104	15°	650	6

* Obtained from ratio of r_c/r_{out} assuming $r_{out} = r_c\,(1-\alpha)$

TABLE 2

TRANSISTOR	r_{out} (kΩ)	C_{out} (pF)	Time Constant (10^{-6} sec)
TJ1 ..	110	700	77
TJ3 ..	72	3200	230

dent of frequency to within less than 5 per cent except in the value of r_e for the Brimar TJ3 which shows a fall of 10 per cent with reduction of frequency. The margin in the case of C_c and $C_c/(1-\alpha)$ is also 10 per cent, but this is due to reduction in sharpness of reactive balance at the lower frequency. This is considered sufficient, but there is also a general tendency for r_e in the low frequency transistors to decrease with frequency and it seems likely that if measurements were made on a large number of such transistors then variations in r_e of greater than 10 per cent may be encountered.

Another point for comment is that if reference is made to the measurement procedure in Circuits 1a and 1b it will be seen that if the transistor equivalent circuit were exactly as shown then the balancing capacitor C_v would not require re-adjustment in the 2nd balance in either circuit. In practice this is not found to be the case and it is attributed to the fact that the inner end of capacitor C_c is connected to some point along the base resistance as shown in Fig. 9, the explanation being as follows.

If the balance conditions are calculated in terms of this transistor equivalent circuit then it is found that the conditions as given by equations (1) to (10) are not affected, provided that for r_b one substitutes $(r_{b1} + r_{b2})$, and also if the values of C_v for the 1st and 2nd balances in either circuit are C_v and C_v' then one obtains the additional conditions for the 2nd balances in each circuit as follows:

$$\textit{Circuit 1a} \quad r_{b2} = (C_v'/C_v)\,.\,R_{v2} \quad \ldots\ldots\ldots\ldots (11)$$

$$\textit{Circuit 1b} \quad r_{b1}' + r_e = (C_v'/C_v)\,.\,R_{v2} \quad \ldots\ldots (12)$$

where $r_{b1}' = r_{b1}\ (1-\alpha)$ and a correction to the value of $(1-\alpha)$ is required unless the frequency dependence of α may be neglected.

Hence, if the equivalent circuit of Fig. 9 is correct then r_{b1} and r_{b2} may be measured separately and their sum compared with the measured value of $(r_{b1} + r_{b2})$ given by equation (3).

Measurements made on the small group of transistors mentioned above show that good agreement is obtained between the sum of r_{b1} and r_{b2} and the measured value of r_b at 1kc/s in the case of the h.f. transistors, but there is poor agreement in the case of the l.f. transistors. Further, when the frequency is reduced to 250c/s, although errors are caused by reduction in sharpness of reactive balance there is a general reduction in the value of r_{b2} of at least 20 per cent and is as high as 50 per cent in the case of the Brimar TJ3. From this it may be concluded, therefore, that although the equivalent circuit of Fig. 9 is more correct than that shown in Fig. 5 there are still certain reactive elements missing which cause a variation with frequency of some of the parameters and this variation is greatest in the case of l.f. transistors, but may be negligible in the case of h.f. transistors.

However, measurements of input and output impedances at higher frequencies suggest that the main reactive elements missing from Fig. 9, excluding the reactance associated with the internal generator, are shunt capacitances associated with r_{b1} and r_e and it is believed that the capacitance in shunt with r_{b1} is of the order 0·001μF to 0·01μF. The balance conditions based on such an equivalent circuit show that at low frequencies where the error in measurement of $(r_{b1} + r_{b2})$ is negligible the error in measurement of r_{b2} is independent of frequency and may be very large depending on the parameters of the transistor. It is clear, therefore, that a negligible variation with frequency in the measurement of r_{b2} does not necessarily imply freedom from error caused by neglect of reactive elements in the transistor equivalent circuit. And since r_{b1} usually appears to form the greater part of $(r_{b1} + r_{b2})$ and its value is subject to considerable error in calculation it follows that large errors of any kind in measurement of r_{b2} may not be detectable.

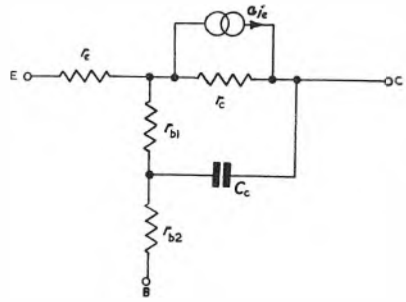

Fig. 9. Illustrating the need for re-adjustment of C_v

Hence the above equivalent circuit is regarded only as a partial explanation of the observed facts and the method is not considered to be generally reliable for measurement of r_{b1} and r_{b2}. However if calculations are confined to equation (1) to (10), excluding equations (4) and (5), it is concluded that the equivalent circuit parameters will be, for most transistors, substantially independent of frequency up to 1kc/s although increases in r_e of greater than 10 per cent for an increase in frequency from 250c/s to 1kc/s may occur in the case of certain low frequency transistors. If, using only equations (1) to (7), frequency dependent errors in r_c and r_c $(1-\alpha)$ are to be completely avoided then it would be necessary to reduce the measurement frequency to about 200c/s or less but this improvement would be obtained at the expense of reduced accuracy in output capacitance measurements.

Acknowledgment

The author is indebted to Messrs. Standard Telecommunication Laboratories Ltd for permission to publish this article and to Mr. T. H. Walker under whose supervision the work was carried out.

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1. HALL, R. A. A 1kc/s Transistor High Gain Tuned Amplifier. *Electronic Engng. 30.* (To be published) (1958).

The Reactance Valve at Audio Frequencies

By B. J. Alcock, B.Sc., Ph.D.

The reactance valve circuit is analysed to derive circuits producing an inductive impedance at audio frequencies with particular reference to obtaining a large Q factor.

THE reactance valve circuit consists basically of a feedback amplifier in which the feedback path has a transmission factor which introduces a phase shift of approximately 90°. The output impedance of the amplifier can be arranged to be predominantly reactive and either positive or negative. The magnitude of the impedance varies with the gain of the amplifier, and is sometimes used to vary the frequency of an LC oscillator. Another feature is that, using only physical capacitive reactances, an effective inductive reactance may be produced. Most of the available theory on reactance valve circuits contains numerous approximations, and its application is limited to radio frequencies. This article attempts to provide a more exact analysis so as to

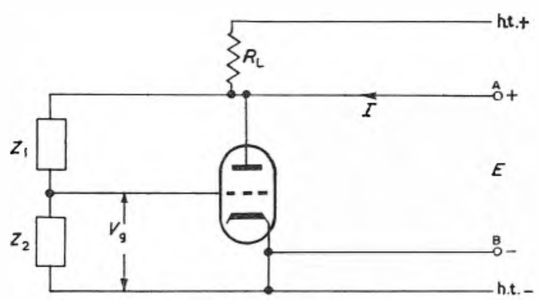

Fig. 1. The basic circuit

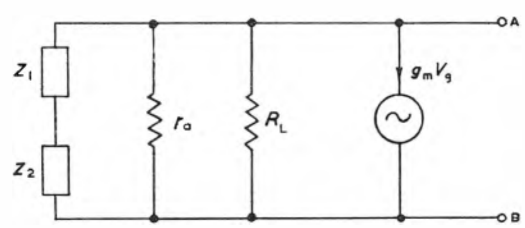

Fig. 2. The equivalent circuit for Fig. 1

deduce the effects of the circuit values upon the effective Q factor of the reactive impedance produced. Calculations are made of the optimum values and circuit arrangements, and the highest Q factors reasonably attainable. The analysis has been carried out for a frequency of 1kc/s.

Analysis of the Basic Circuit

The incremental impedance Z_{AB} seen between the terminals A and B can be found using the equivalent circuit of Fig. 2, where g_mV_g is a constant current generator, from which the ratio E/I for incremental changes and assumed quasi-linear conditions can be calculated as follows.

Let

$$\beta = \frac{Z_2}{Z_1 + Z_2} \quad \text{.................. (1)}$$

Then from Fig. 1

$$V_g = \beta E \quad \text{.................. (2)}$$

Thus the equivalent circuit for the impedance Z_{AB} is as shown in Fig. 3.

When Z_1 is a resistance R and Z_2 is a capacitance C then

$$\beta = \frac{1}{1 + j\omega CR}$$

Thus

$$1/\beta g_m = (1/g_m) + j\omega(CR/g_m) \quad \text{........ (3)}$$

One may now write the circuit for impedance Z_{AB} as shown in Fig. 4.

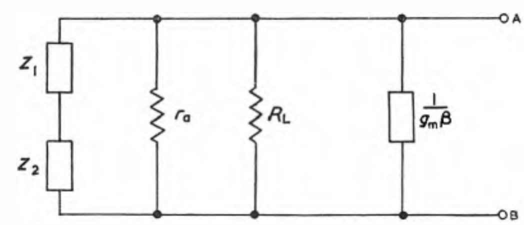

Fig. 3. The output impedance Z_{AB}

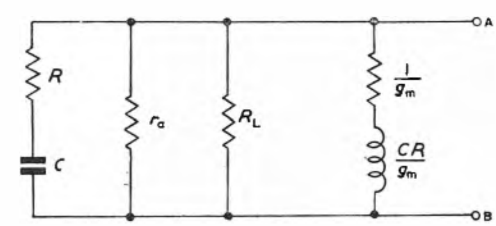

Fig. 4. The impedance Z_{AB} when an RC feedback network is used

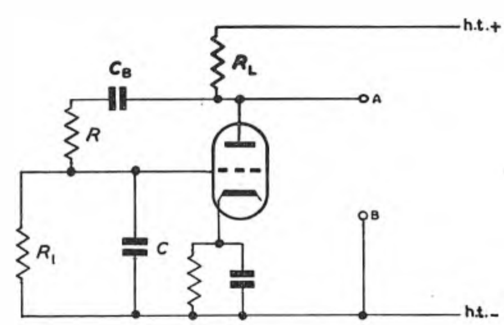

Fig. 5. A practical circuit

In order to arrange for the correct d.c. operation of the valve the practical circuit will take the form shown in Fig. 5 in which C_B is a large blocking capacitor, and R_1 is the necessary grid to earth resistor.

In considering the optimum values for the circuit elements so as to give an inductive impedance with the maximum Q, the following points may be noted.

R_L must be as large as possible, preferably at least as large as r_a which must also be large. Also as the inductance and series resistance are shunted by resistances then for the maximum possible Q, g_m must be as large as possible. The conditions for r_a and g_m can be best satisfied by a high slope pentode, and in the experimental work an SP61 was used.

A convenient method of making R_L large, without having to resort to very large values of h.t. is to use a triode with current feedback as the anode load of the reactance valve. The incremental impedance Z_{XY} seen between the points X and Y in Fig. 6 is given by

$$Z_{XY} = r_a + (\mu + 1)R_B \quad \ldots\ldots\ldots\ldots (4)$$

assuming that $R_g \gg (1/j\omega C_g)$ and that the cathode bias resistor R_X is small compared to the cathode load resistor R_B.

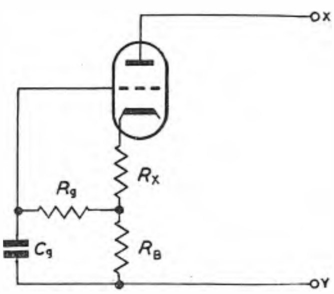

Fig. 6. A load network with a high incremental impedance

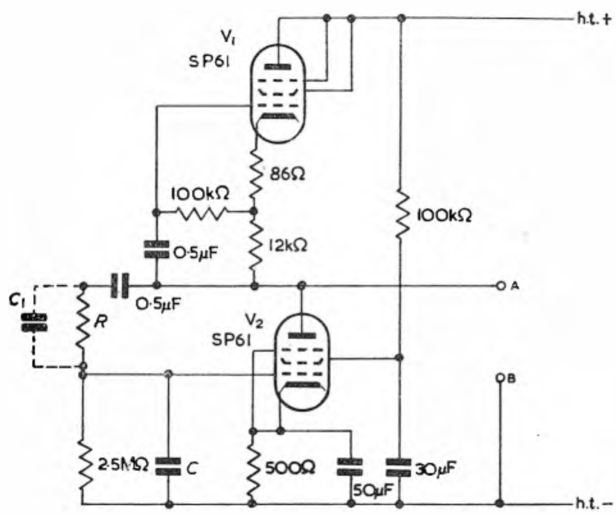

Fig. 7. A practical circuit incorporating a high impedance load network

A suitable circuit is shown in Fig. 7 using SP61's for both valves, the upper one being strapped as a triode. Using the values shown the a.c. impedance of the triode with current feedback is about 0·7MΩ, and this is approximately equal to r_a of the pentode.

In this form, however, the Q of the inductive impedance is limited by several factors. Firstly R and C appear in series across the terminals AB, and if R is made large the stray capacitances across it affect the phase shift and lower the Q. Secondly as R and thus $1/j\omega C$ must be kept large the grid resistor also reduces the maximum Q obtainable. The expression for the Q for this circuit is given below, which shows the effect of R_1 and the stray capacitance C_1 across R, assuming the blocking capacitor C_B is large enough to be neglected.

$$Q = \frac{1}{(1/\omega CR) + (1/\omega CR_1) + \omega C_1 R + (\omega CR/g_m R_a) + (\omega C/g_m)} \quad \ldots\ldots\ldots\ldots (5)$$

where R_a is the anode impedance of V_2 in parallel with its anode load impedance. The optimum values of R and C are obtained by partially differentiating equation (5) with respect to C and equating to zero, then repeating this with respect to R, and solving the two resulting equations for C and R. The final equation in R is unfortunately of the fifth order and its solution can be laborious, especially if complex roots are present. For the circuit of Fig. 7 C_1 was

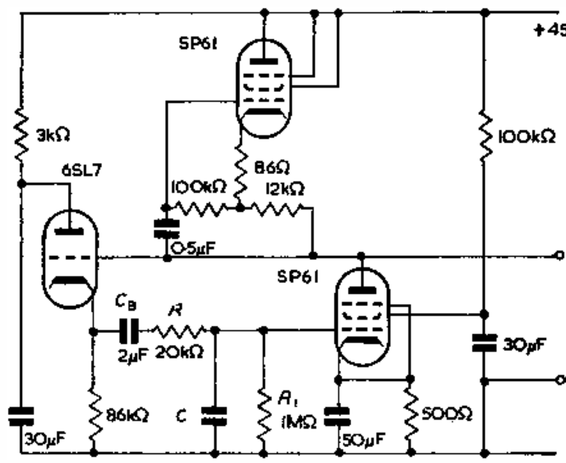

Fig. 8. An improved circuit with cathode-follower buffering

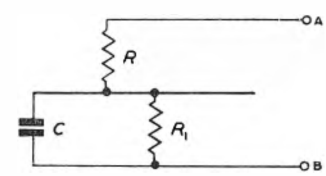

Fig. 9. The equivalent feedback network of Fig. 8

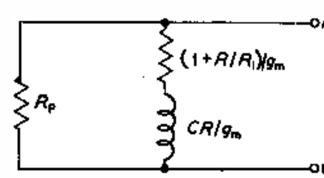

Fig. 10. The impedance Z_{AB} for Fig. 8

approximately 7pF and the optimum values of R and C calculated for a frequency of 1kc/s using the above method were 270kΩ and 0·03μF respectively. This gives a calculated value for Q of 14 and the measured value was 13·2.

An Improved Circuit

Fig. 8 shows a circuit with a cathode-follower in the feedback path. This enables R to be reduced in value so that C_1 may be neglected. Also R and C in series do not appear across AB, and the direct coupled cathode-follower input impedance is high enough to be ignored.

The values of the feedback network must now be chosen so as not to adversely affect the cathode-follower gain, and also to take into account the criteria mentioned above. The loss in the cathode-follower may be taken into account in the calculations by multiplying the g_m of the reactance valve by the gain of the cathode-follower.

Using the circuit of Fig. 8 in which R has been made as small as possible, and R_1 as large as possible, and C_B large

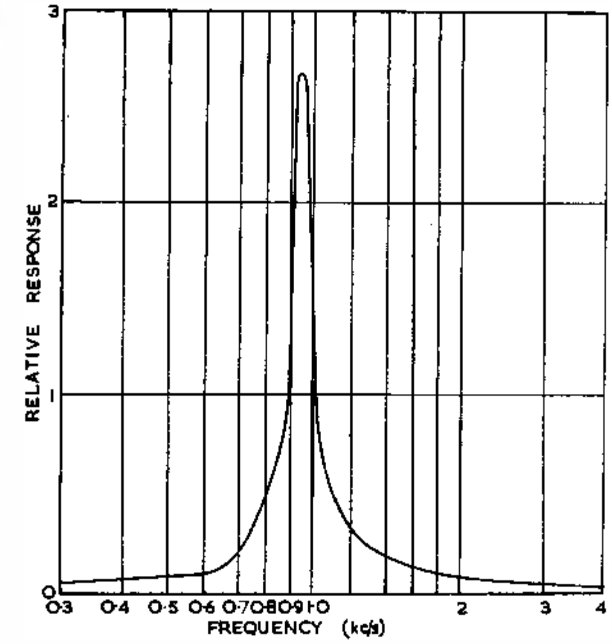

Fig. 11. The graph of the relative response of the tuned circuit consisting of Z_{AB} in parallel with a capacitance

enough to be neglected, the feedback network is as shown in Fig. 9.

Now

$$\beta = \frac{1}{j\omega CR + R/R_1 + 1} \quad \ldots\ldots\ldots\ldots (6)$$

Thus the equivalent circuit for Z_{AB} is now as shown in Fig. 10.

Now Q_o for the network excluding R_a is given by

$$Q_o = \frac{\omega CR}{1 + R/R_1}$$

And Q for the whole network is therefore

$$Q = \frac{1}{1/\omega CR + 1/\omega CR_1 + \omega CR/g_m R_a} \quad \ldots\ldots (7)$$

Considering only the denominator of equation (7).

$$d(1/Q)/dC = -(1/\omega C^2) - (1/\omega C^2 R_1) + (R/g_m R_a)$$

Equating to zero and taking the value to give a minimum for $1/Q$.

$$C = \sqrt{\left(\frac{R + R_1}{R_1} g_m R_a\right)}$$

Substituting the numerical values. $R_1 = 10^6\Omega$,

$R = 20 \times 10^3\Omega$, $g_m = 9{\cdot}3 \times 10^{-3} \times 0{\cdot}94$mA/V,

$\omega = 2\pi \times 10^3$, $C = 0{\cdot}42\mu$F which gives

$Q = 26{\cdot}0$

and $L = 1{\cdot}05$H

The values measured on a circuit were

$Q = 24{\cdot}5$

and $L = 0{\cdot}98$H.

A response curve of the tuned circuit formed by placing a capacitor across AB to resonate at 1kc/s is shown in Fig. 11.

This appears to be the optimum performance obtainable without further complication, and in this form can be used to provide effective inductances with a reasonable Q. The inclusion of an amplifier (with zero phase shift) in the feedback path will raise the Q obtainable, the limit being one of stability.

A Conductivity Storage Transistor Pulse Width Modulator

By J. C. Price*, B.Sc.

Transistors of both point-contact and junction types exhibit in certain pulse applications, prolonged conduction generally ascribed to 'hole storage'. This can be controlled in a circuit to provide a simple means of pulse width modulation for time division multiplex. The technique is described for use with selected point-contact transistors.

A VERY important field of application for transistors is in low power pulse circuits such as are used in digital computers and time division multiplex systems. The small physical size and modest power requirements of transistors make elaborate electronic systems practicable without the apparatus becoming very bulky. Since there appears to be nothing in a transistor to wear out[1] it should be possible ultimately to achieve a very high standard of reliability with these devices. These considerations point to the importance of transistor pulse circuits particularly those requiring a minimum of associated components.

The transistor pulse width modulator described here had its origin in work carried out to achieve improved reliability and independence of individual transistors in pulse circuits. In a particular pulse circuit it was found that, with the samples of point contact transistors then being produced, a wider pulse occurred than was predicted by simple theory. A way was available to prevent the unwanted effect in the particular application, and further, the possibility was considered of exploiting the effect to some useful purpose. From this emerged the pulse width modulator described here.

Mechanisms Influencing Turn-Off Time in Transistor Pulse Circuits

Consider a transistor circuit fully turned on. Making a direct appeal to elementary physical theory of transistors, it would appear that the collector current might be limited by two possible mechanisms:

(1) Under the operating conditions at that point, the emitter is unable to supply sufficient holes.

* *Standard Telecommunication Laboratories Ltd.*

(1) The emitter is emitting holes freely, but the collector potential is insufficient to enable these to be collected.

Thus one might expect collector current pulses of equal duration with the emitter current to occur in condition (1), and widened pulses in the case of (2) due to the excess of holes injected during the emitter pulse but not used.

A tentative correlation of this with the load line diagram of the pulse circuit is as follows:

Fig. 1 is a sketch showing the load line of some pulse generator circuit on the output characteristics of a point contact transistor. The load line is shown straight for illustration. In practice it may have a rather complicated form due to the effect of feedback or non-linear loads. In the case of a pulse generator of the type shown in Fig. 2, if the means of feedback is an open-circuited delay line, initially the load approximates to the characteristic impedance of the line in shunt with R_2. The transistor characteristics are shown in two parts by whole and broken lines. It may be supposed that the region of full line presentation corresponds to condition (1) and that the region of broken line presentation corresponds to condition (2).

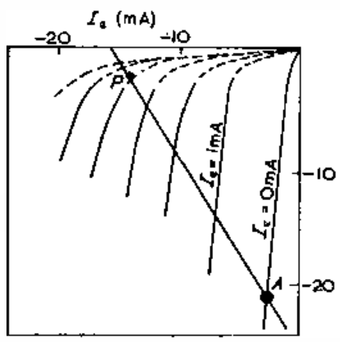

Fig. 1. Load line of pulse generator where P falls in region (2)

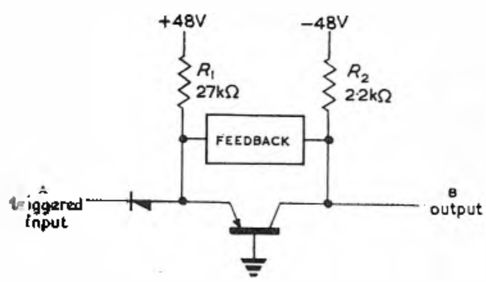

Fig. 2. Basic type of trigger circuit

Point A represents the 'off' state of the pulse generator and point P the 'on' state. Where point P occurs will be a function of emitter current and the current gain of the transistor for given circuit parameters. Fig. 1 illustrates the case where P falls in region (2). Fig. 3 shows the case where P occurs in region (1). Fig. 4 shows the case where the load line changes direction at Q and the 'on' position occurs at R. Such an effect could be obtained by a catching diode. This was suggested by E. H. Cooke Yarborough[2] for preventing widened pulses. The larger collector potential in the 'on' state secured in this way assists the attraction of holes to the collector; consequently the collector current is somewhat greater and fewer holes are stored in the body of the transistor.

From the above remarks it will be seen that if a circuit is devised in which a modulating signal controls the extent to which a pulse generator in the 'on' condition is driven into condition (2) width modulation of the pulse will occur.

Theory of Operation

A basic type of transistor trigger circuit[3] is shown in Fig. 2. The transistor is a point contact type. Typical biasing conditions are indicated and the feedback from collector to emitter may take the form of a capacitor, series tuned circuit, or open-circuit delay line.

With point A held at a small negative potential to earth, the emitter will produce no holes and the collector will present a high impedance. The potential of the collector is of the order of -40V. If the potential applied to A is raised sufficiently to cut off the input diode, the current flowing through R_1 will be directed to the emitter and holes will be injected into the germanium. Some of these are attracted to the collector causing increased collector current. This will raise the potential of the collector. With a feedback circuit such as any of the types mentioned, the action will be cumulative while the feedback continues, and the pulse generator is driven to the 'on' position. After a period of time which is a function of the feedback circuit the current fed back to the emitter ceases preventing further emission of holes. In the absence of appreciable hole storage and transit time effects, the collector current and hence collector potential falls, and the pulse generator reverts to the 'off' position.

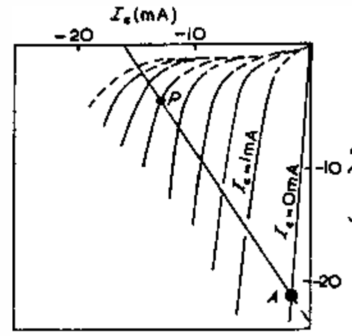

Fig. 3. P falling in region (1)

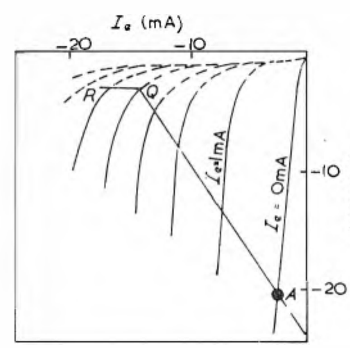

Fig. 4. Load line changing direction of Q

The description so far is a brief sketch of the action of a normal transistor pulse generator. If, however, during the emitter pulse, holes are stored in the germanium, the collector current will continue at a high level until these holes are used up as considered above. Such conditions of hole storage are obtained with selected transistors of high current gain and by use of a fairly high collector resistance. A diode is used to control the collector potential, and hence hole storage, during pulses giving direct width modulation of the normal generator output[4].

The circuit of a complete pulse width modulator is shown in Fig. 5. X_1 is a selected point contact transistor. The pulse generator is triggered by a positive-going pulse to the emitter via MR_1. In the conducting conditions the collector potential is clamped by MR_2 connected to the modulating audio source. A load of 680Ω is used shunted by a 0·04μF

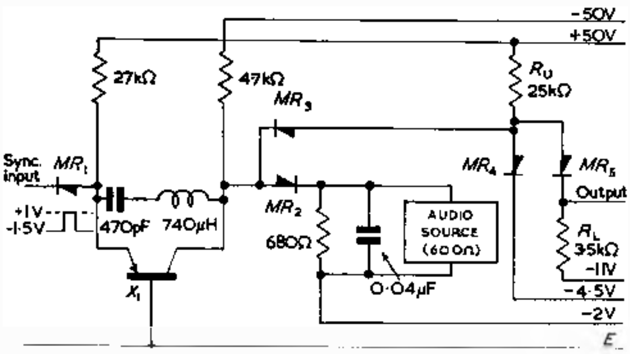

Fig. 5. Pulse generator for width modulator

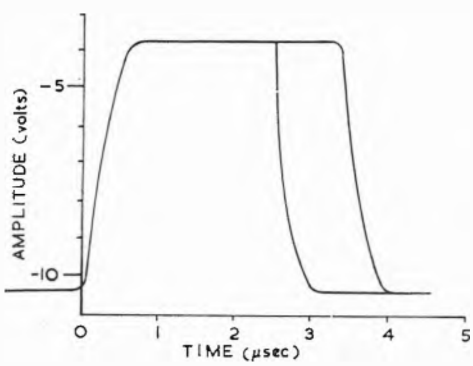

Fig. 6. Width modulation of output pulse

capacitor for storing energy between pulses. The diodes MR_3, MR_4 and MR_5 provide the necessary clipping of unwanted portions of the waveforms. A typical output waveform is shown in Fig. 6.

Performance Details and Selection of Transistors

Fig. 7 is a curve showing the relation between delay of the back edge of the pulse, measured from an arbitrary origin and the controlling voltage on diode MR_2. It will be seen that a linear relation holds over the region, approximately $-2 + 0{\cdot}75$V. The only limitations on modulating frequency are those inherent in the recovery time of the pulse generator. Since the pulse generator used in the tests was operated at about 20kc/s there is clearly no restriction on the audio range that can be modulated in this way.

An advantage of this method of p.w.m. is that it produces a width modulated pulse of reasonably high power. Hence for example p.t.m. (pulse time modulation) can be readily obtained using a differentiating transformer.

It is found that most production point contact transistors will give some storage effect of the type required, e.g. Bell A1698 and Standard 3X/100, but the effect is generally too small to be useful. This is particularly so since manufacturers of point contact transistors for pulse circuits now generally aim to reduce hole storage as much as possible.

For practical modulation the large storage effect obtained with a few selected samples is required. Suitable transistors were most plentiful in a batch made with 4 to 6Ω-cm germanium with antimony only as the impurity.

In the course of selecting transistors two effects have been noted which tend to be associated with this storage effect. The first of these is high initial values of current gain α. It is required that the mean α for $0 < I_e < 0{\cdot}5$mA and $V_c = -20$V must exceed five, where I_e = emitter current and V_c = collector voltage. The other associated effect is a falling off in rectification efficiency at a comparatively low frequency, approximately 50kc/s, using the collector and base as a diode.

The basic cause underlying these three related effects is not completely understood. Attempts have been made to interpret them in terms of various physical theories of the point contact transistor which have been advanced publicly[5,6] and in private discussion, but no single theory has been decisively established.

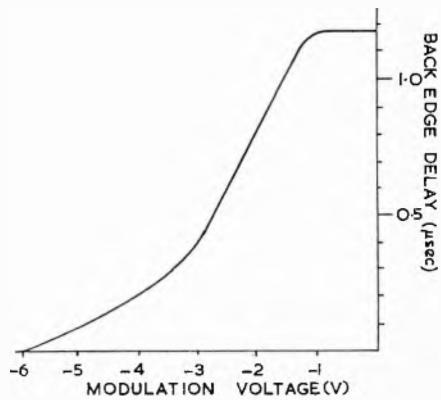

Fig. 7. Delay back edge of pulse as a function of modulation potential

Conclusions

The degree of selection of transistors now necessary makes the present pulse width modulator purely experimental. It was first thought that it might be possible to obtain suitable storage by some minor change in the manufacture of present point contact transistors, but this has not proved so easy. With improvements in frequency response of junction transistors these have gained in favour for pulse circuits both on account of their greater gain and reliability. It may well be that in the future these will be used for conductivity storage. Nor are the possibilities limited to modulation of collector potential, for other arrangements in which storage is controlled by modulating potentials, currents or impedances are possible.

Moreover, if effective control of conductivity storage in semi-conductor devices can be achieved, a much wider use of semi-conductors in multiplexing and automatic switching can be envisaged in which conductivity storage in a dice of germanium replaces energy storage in relatively bulky capacitors and inductors.

Acknowledgments

The author is indebted to Mr. K. W. Cattermole for the point contact pulse generator employed in the modulator circuit described, and thanks Standard Telecommunication Laboratories Ltd. for permission to publish this article.

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A FUNCTION GENERATOR

By N. Hambley*, B.Sc.

When it is desired to control an operation according to a complex variable function, it is convenient to obtain control directly from a graphical function. The apparatus described here uses a photocell and cathode-ray tube to convert graphical data into a suitable electrical form. The conversion accuracy is better than one per cent.

IN order to study the behaviour of metal structures and specimens at various temperatures, heating equipment is required and it is necessary that the temperature is accurately controlled. During the investigations it may be necessary to change the desired rate of temperature rise between wide limits. The heating equipment is required to follow different functions for the various specimens, and the curves may be complex. To cover all possibilities an automatic graph reading device was considered necessary, so that the temperature control can be derived quickly from a graphical function. This article describes the design and construction of apparatus for converting a graphical function into an electrical form which is used to control the temperature of a specimen.

Principle of Operation

Referring to Fig. 1, it will be seen that the conversion from a graphical function to a proportional voltage is achieved by means of a closed loop feedback system involving the cathode-ray tube, photocell and Y amplifier. The vertical position of the spot on the cathode-ray tube is controlled by the Y amplifier. With the feedback loop closed, as shown, the Y amplifier is controlled by the photocell which in turn, is actuated by the cathode-ray tube spot. If the graphical function, with its upper portion blacked out, is interposed between photocell and cathode-ray tube, the spot will adjust itself on to the edge of the black portion. The sense of the feedback is arranged such that any deviation from the common black-white boundary causes the Y amplifier to exert a greater pull back again. This ensures that the spot remains on the black-white boundary all the time when the feedback loop is closed. The output voltage function is derived from the Y amplifier, which gives a voltage proportional to the vertical position of the spot. The cathode-follower and level control reduce the function into the necessary form for operating the temperature controlling system.

It is, of course, essential that the photocell receives light only from the cathode-ray tube spot. Accordingly the cathode-ray tube and photocell are housed in a light proof enclosure. The horizontal spot position is controlled by the time-base. The latter can be set to scan the graph in any integral number of minutes from one to thirty inclusive. Stabilized power supplies are necessary, to keep the conversion within the specified accuracy of one per cent.

In order to exclude external light, the photocell and cathode-ray tube are housed in a light proof enclosure. This assembly is painted with a matt black finish to minimize the effects of reflection. The graphical function is carried on a frame which is mounted between the photocell and the face of the cathode-ray tube. The front portion of the light proof enclosure is hinged to allow access for interposing the function between the tube and photocell. A typical graphical function is shown in Fig. 2. Temperature is the ordinate and the upper portion of the curve is blacked out. This may be done by cutting the curves from black card, or by blacking out one portion of a photographic plate.

* *A. V. Roe & Co. Ltd.*

The cathode-ray tube, (type DB 13-2) has a post deflexion accelerator which enables a fine spot to be obtained. The electrode potentials are tapped from a resistor chain connected between +2kV and −2kV, so that the focusing assembly voltage is approximately +250V. This is neces-

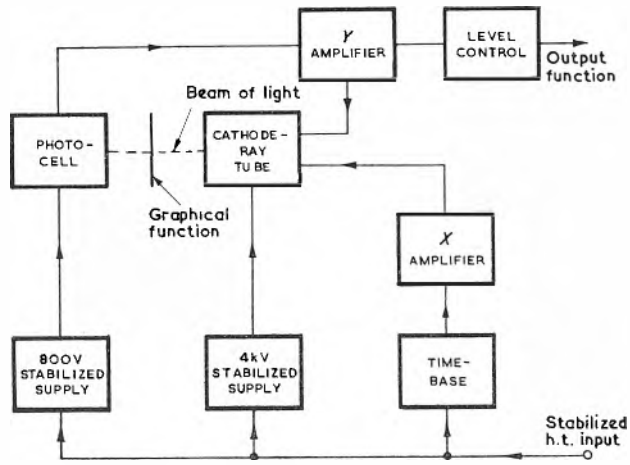

Fig. 1. Block diagram of function generator

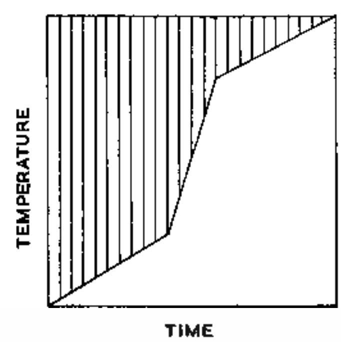

Fig. 2. Typical graphical function

sary to prevent defocusing due to a difference of voltage between the focus anodes and the mean deflector plate voltages. Since these voltages are directly coupled to the X and Y amplifiers they necessarily have a mean voltage of +250V. Symmetrical push-pull deflexion systems are used on both X and Y amplifiers. A Mumetal screen is placed round the tube to minimize the effect of stray magnetic fields.

The photocell is a photo-multiplier, type 27M2, and requires a potential chain of 800V to operate with a current gain of 250 000. The current caused by full view of the cathode-ray tube spot is about 1mA and the dark current is less than 1μA. This current variation demands a high degree of stabilization in the 800V power supply. As the photo-multiplier is most sensitive to blue light, this colour is chosen for the cathode-ray tube face.

Stabilized 800V Supply

The anode of the photo-multiplier is directly coupled to the input of the Y amplifier and so the positive end of the

800V is at approximately earth potential. A radio frequency oscillator unit[1] is chosen as the source of e.h.t. voltage because of its inherent advantages over the conventional mains method. There is a reduction in size and weight of the transformer and capacitors. In addition, an r.f. supply is non-lethal and an accidental overload quenches the oscillation.

Fig. 3 shows an audio pentode is used as oscillator. Transformer coupling is arranged between anode and grid to provide the oscillation. The e.h.t. and rectifier heater

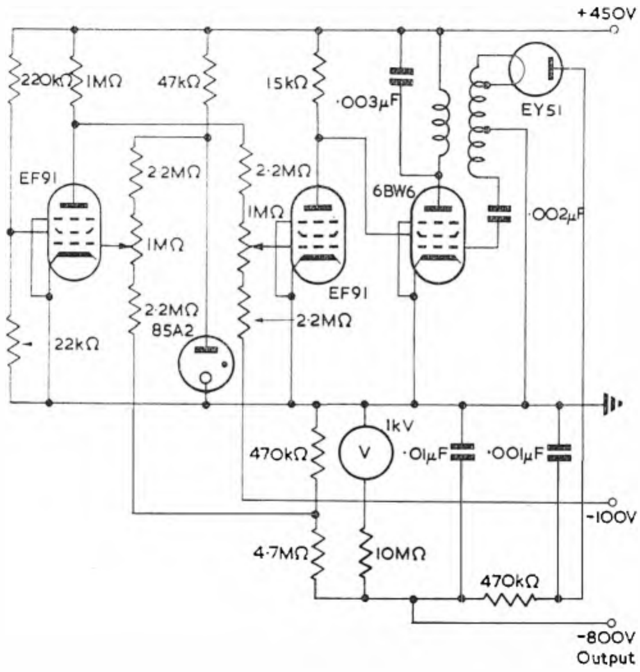

Fig. 3. 800V power supply

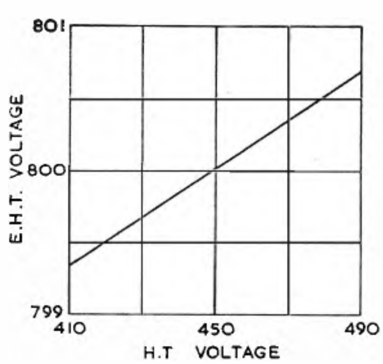

Fig. 4. E.H.T. regulation

supply are derived from coils wound on the same transformer. The e.h.t. transformer was originally wound on an air-cored former, but later a ferromagnetic core was used and found to give better regulation and simpler construction. The material used is Mullard FX1031 ferroxcube.

Stabilization[2] is achieved by controlling the screen grid voltage of the oscillator valve. A portion of the e.h.t. is fed back, after amplification, on to the oscillator screen grid. It is necessary to use a two valve system in order that the feedback is in the correct sense. The amplifier is d.c. coupled, the first stage being operated under 'starvation' conditions. The second stage acts as an invertor with comparatively low gain. The overall gain of the amplifier is about 5 000 and the losses within the loop amount to about 100. This indicates a stabilization ratio of 50 to 1. The regulation curves are shown in Fig. 4. The level of ripple is 0·05V, at 10kc/s. The output voltage is monitored on a moving-coil meter with a 100μA movement and a 10MΩ series resistance. The h.t. power is drawn from a stabilized power supply giving +450V and −100V.

The 4kV Supply

The cathode-ray tube e.h.t. is similar in principle to the 800V supply. One difference is, however, that the 4kV potential chain is earthed at its centre point so that the focus anode voltages may be at about +250V. This makes the final anode at +2kV and the grid at −2kV. Also it is found to be necessary to use two audio pentodes for the oscillator, because the tube takes a beam current of 2mA at full brilliance. The stabilization of this power supply is achieved by a two-stage d.c. amplifier. The percentage regulation is similar to that shown in Fig. 4. One of these units has been in continuous use for a period of over 6 months, and during this time has not required any adjustment. H.T. again comes from a stabilized power pack.

X Amplifier and Y Amplifier

The amplifiers are shown in Fig. 5. A twin triode is used as a long-tailed pair to provide push-pull deflexion voltages to the X plates. The anodes are connected to the X plates. One of the grids is connected to the X shift control and the other grid is connected to the time-base voltage generator. The Y amplifier has a similar arrangement, except that the second grid is connected to the photo-multiplier anode in order to close the feedback loop.

Feedback Loop

To follow the action of the feedback loop it is convenient to start at the cathode-ray tube spot. When the spot is partially hidden behind the blacked out portion of the curve, the photo-multiplier receives a certain amount of light, which causes a potential drop across its anode resistor. This voltage is directly coupled to the Y amplifier grid, which controls the vertical position of the spot. This adjusts itself until the portion visible to the photocell is sufficient to balance its vertical position. It will be seen, therefore, that as the spot follows the function more of the spot is visible at one end of the curve than the other. This effect is shown diagrammatically in Fig. 6, where the spot is shown in three positions. The spot size is magnified over 30 times for the purpose of illustration. The dotted line shows the deviation from the function, which is an inherent feature of the d.c. feedback loop. In practice the spot has a very small diameter, and the variation of the visible amount of the spot is about one-tenth of its full size. This limits the maximum deviation to one-tenth of 1 per cent of full scale.

Time-Base

A slightly modified Miller rundown circuit is used to provide the necessary linear voltage change for applying to the X amplifier, shown in Fig. 5. The spot is thus made to cross the cathode-ray tube face at a steady speed. The time taken is selected by varying the resistor in the Miller *RC* network. This can be set to give full scan of a function in any integral number of minutes between one and thirty inclusive. The time-base is started by applying a positive step function to the suppressor grid of the Miller valve. The linearity of the time-base is better than one-tenth of one per cent. The h.t. power for the time-base is derived from a +450V stabilized line and a −100V stabilized line.

Output and Level Control

The output function is in the form of a voltage, the

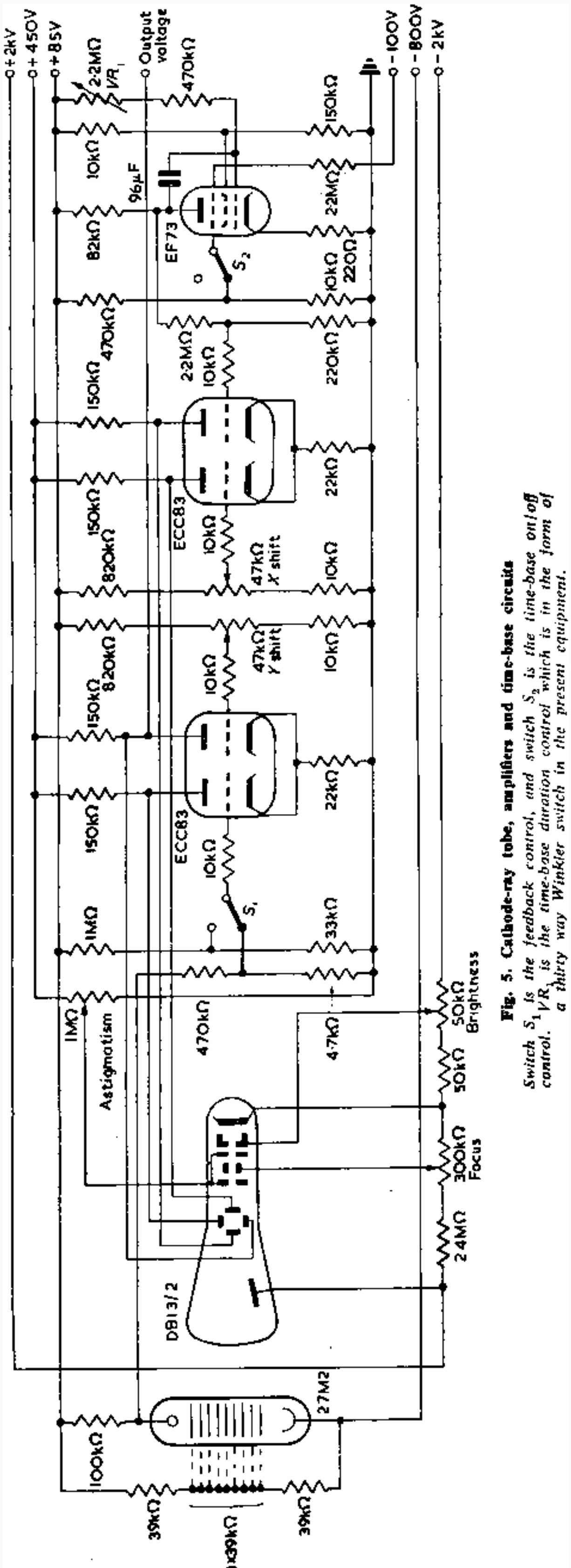

Fig. 5. Cathode-ray tube, amplifiers and time-base circuits

Switch S_1 is the feedback control, and switch S_2 is the time-base on/off control. VR_1 is the time-base duration control which is in the form of a thirty way Winkler switch in the present equipment.

amplitude of which is proportional to the ordinate of the graphical function at any time. The voltage function is derived from the Y plates of the cathode-ray tube and is attenuated to give a range of zero to 8V, which corresponds to the range of the ordinate on the graph. The function is fed through a cathode-follower to the output terminals, see Fig. 7. Two safety circuits are incorporated which open-circuit the output in the event of the voltage function leaving the range zero to 8V. The safety circuits

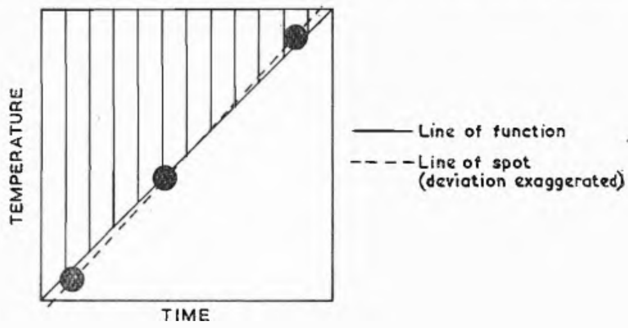

Fig. 6. Showing deviation of spot

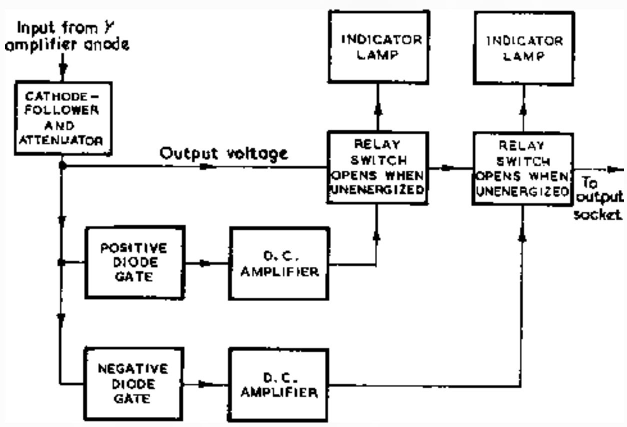

Fig. 7. Level control system

When the output voltage reaches +8V the positive diode gate opens, after which 0·01V applied to the d.c amplifier is sufficient to make the d.c. amplifier open the first relay switch. This open-circuits the output, and the switch remains open until the voltage returns to below +8V. The negative diode gate operates similarly if the output voltage becomes negative. The indicator lamps are arranged to show when their associated relays are energized.

are fail-safe devices, and one or the other of the circuits will operate when any valve or component failure occurs. In conjunction with the safety circuits, warning lamps are arranged to indicate the type of fault.

Constructional Details

The equipment consists of two units, each of which is mounted on a standard Post Office rack panel. One unit holds the light proof enclosure and the other unit contains the amplifiers, power packs, time-base and the controls.

Operating Procedure and Characteristics

Since the equipment is designed for use by unskilled personnel, the number of controls is reduced to a minimum. When the power packs have been switched on there are only four controls which are used to generate a function. Two of these controls are used to set the abscissa and ordinate of the function to correspond with units chosen for temperature and time on the graph. Of the remaining two controls, one is used to close the feedback loop and the other is used to start the time-base, which begins the generation of the function, At the end of the time period selected by setting the abscissa, the function switches off automatically. The output voltage function is monitored by a meter on the front panel.

Conclusions

The equipment follows a curve within an error of one per cent. At present it is being used to convert graphical data into an electrical form for controlling temperature. It is felt that it may have many more industrial applications as an alternative to the more expensive methods of control by digital computation.

Acknowledgments

The apparatus described in this article was developed as part of the research programme of Messrs. A. V. Roe & Co., Ltd, and the author would like to express his thanks to the Directors for permission to publish this article.

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A Three-Phase Three-Valve Multivibrator

By W. F. Lovering*, M.Sc., A.M.I.E.E.

The two valve multivibrator is a well-known source of rectangular voltage waves. In this article a three-valve, three-phase version of this circuit is described.

THE two valve multivibrator is well known as a convenient source of rectangular voltage waves; it may be regarded as a development of the bistable Kipp relay circuit. The Kipp relay has two possible stable states with one valve cut-off and the other conducting: the multivibrator changes from one state to the other as the charge on the appropriate coupling capacitor leaks away.

Multi-stable circuits may be made by connecting three or more valves with d.c. couplings; each anode being coupled to every other grid. A three valve circuit connected in this fashion has eight possible stable states, of which two, namely all valves conducting and all valves cut-off, are unimportant.

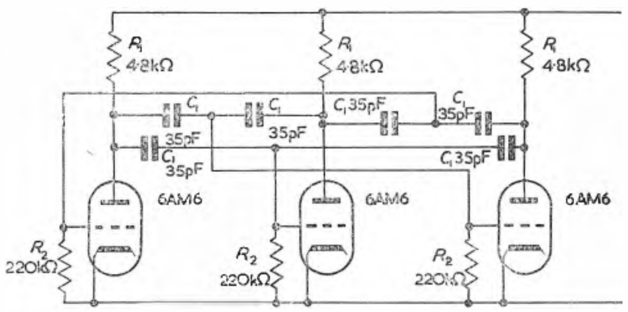

Fig. 1. Circuit of three-phase multivibrator

With screens at +250V, repetition frequency $\simeq$ 13kc/s. The grid time-constant is $2C_1R_2$

If three valves are so coupled and a small negative grid bias voltage is applied all three valves conduct. If the negative grid voltage is gradually increased a state is reached at which one valve ceases to conduct. The anode voltage of this valve is then high which causes the other two grids to be driven positive so that grid current flows. A voltage pulse applied to the grids causes the non-conducting valve to conduct once more while one of the other valves cuts off. The circuit in this condition has three stable states with one valve cut off and the others conducting. If the negative grid bias voltage is increased further another condition is reached in which the circuit has three stable states but with one valve conducting and the others cut off.

The three valve circuit thus has two possible modes of operation each of which has three stable states. In either mode of operation the circuit may be triggered from one stable state to another by the application of an appropriate pulse to all three grids or anodes. If the circuit is quite symmetrical the states will succeed one another in a random fashion but if an additional unidirectional coupling is introduced by small capacitors connected in sequence a_1 to g_2, a_2 to g_3, a_3 to g_1, the states will succeed one another in this order.

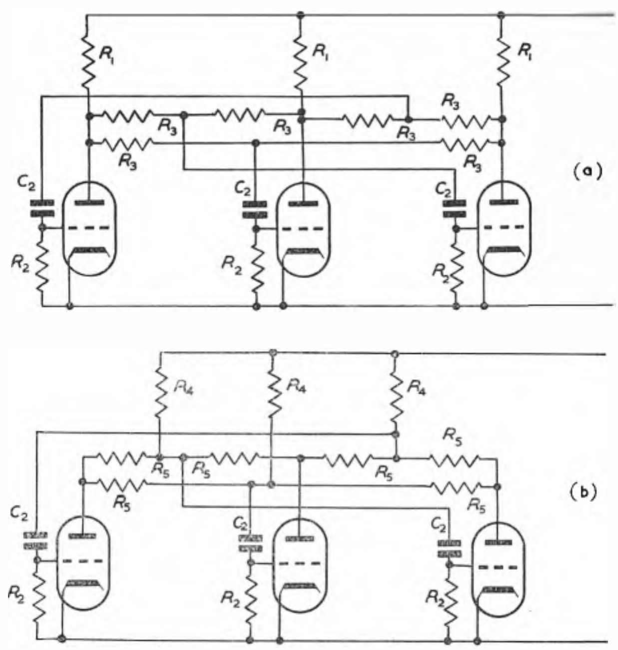

Fig. 2. Alternative circuit connexions

If the three valves are coupled by capacitances only (Fig. 1) the circuit acts as a three-phase multivibrator and generates rectangular anode voltage waveforms having a cycle of two thirds low voltage, one third high; or two thirds high, one third low, depending upon which of the two possible modes occurs.

Alternative connexions are shown in Fig. 2. The opera-

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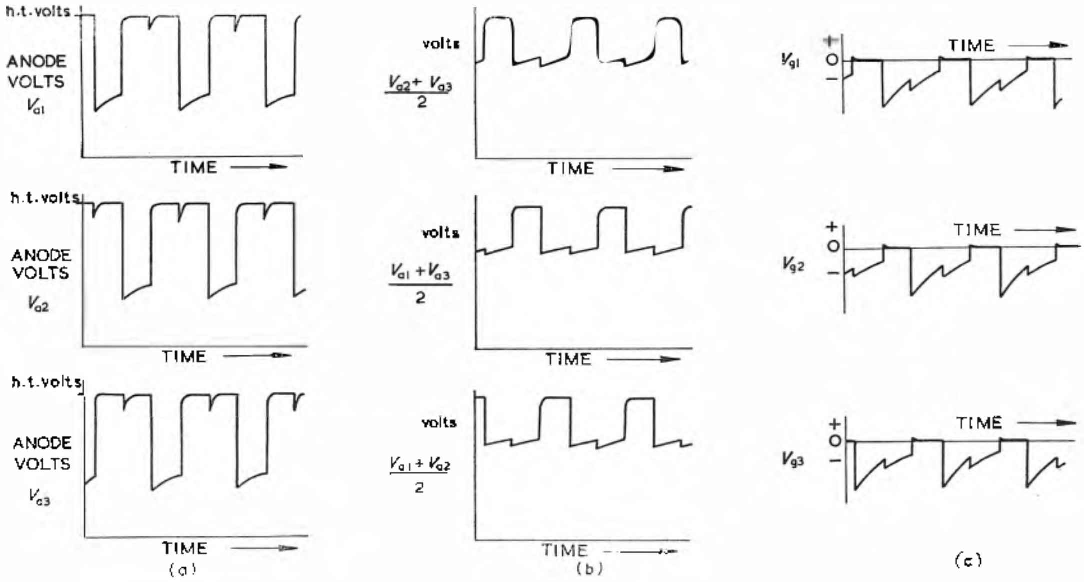


Fig. 3. Voltage waveforms

(a) Anode voltages
(b) Voltages mid-way in potential between anodes
(c) Grid voltages

tion of all three circuits is much the same but the circuit of Fig. 2(a) is simplest to describe. The two-off, one-on mode is assumed.

Each grid is supplied through a capacitor from a point which is midway in potential between the anodes of the other two valves. If the anode potential of V_1 remains constant, a rise in the anode potential of V_2 causes a rise of half this amount to be communicated to the grid of V_3. The anode potential of V_3 then falls and half this fall is applied to the grid of V_2. V_2 and V_3 therefore act as a multivibrator and 'turn over'. The anode potential of V_2 rises and that of V_3 falls and there is little change in potential of the mid-point. The grid potential of V_1 is therefore unchanged, as is its anode potential.

The anode and grid voltage waveforms of an oscillator working in the two-off, one-on mode are illustrated in Fig. 3. Each pair of valves acts in turn as a two-valve multivibrator and one is switched off and the other switched on.

The order of conduction, V_1, V_2, V_3 or V_1, V_3, V_2 is determined originally by any asymmetry in the circuit but once switching has commenced in a particular order, this order is maintained. Turn-over takes place when the grid capacitor of a valve which has been cut-off discharges to the point at which anode current restarts. The grid capacitor which has the greater time to discharge is, of course, that which is connected to the grid of the valve which has been cut off for the greater time. This valve is therefore the next to start conducting, when it turns over with the valve which has been conducting.

Oscillation in the 'two-on, one-off' mode is possible but is difficult to initiate. If the grid resistors are returned to a potentiometer connected across the h.t. supply and a positive bias applied while the circuit is oscillating it is possible to initiate this mode. However, if the h.t. supply is then switched off and restored, the oscillation does not recommence until the potentiometer is returned to the zero bias position. Oscillation then restarts in the two-off, one-on mode.

The two-on, one-off mode provides positive anode pulses lasting one third of a cycle. This is the type of waveform required for operating a three valve gate circuit using suppressor grid switching; each valve is switched on for one third of a cycle.

The two-off, one-on mode is more economical of h.t. power and provides positive anode voltage pulses of two-thirds of a cycle in duration. However, the circuit may also be used to supply positive voltage pulses of one-third of a cycle duration but of half the amplitude. This is simply achieved by taking the output from a point mid way in potential between the anodes; the potential at the point is half the sum of the anode potentials (Fig. 3(b)).

When a valve is switched off, its anode potential rises rapidly to the h.t. potential and stays constant at this value until the valve once more conducts. While the valve is conducting its anode potential changes slightly as the grid capacitor charge current changes. The positive half-wave is a close approximation to a rectangle but the negative half-wave is not.

The rise and fall times of the positive pulse are determined by the anode load resistance and the stray valve capacitances. Using miniature r.f. pentodes and anode loads of 5kΩ a rise-time of about 0·5μsec is obtained and a recurrence frequency of 50kc/s (positive pulse duration 6·6μsec) is possible with a good switching waveform.

For switching purposes a positive pulse is applied to a gating valve which is normally biased beyond cut-off; for this purpose only the positive waveform is important.

It has been found that four or more valves may be used in similar circuits to produce polyphase multivibrators but the circuits required are so complicated that their utility is doubtful; a four valve circuit for example requires twelve coupling resistors or capacitors to couple each anode to every other grid. The three valve multivibrator was developed to provide a switching waveform suitable for operating a three channel cathode-ray tube beam switch which could be used for the simultaneous examination of phenomena in each phase of a three-phase circuit. For this purpose the three-phase multivibrator was found to be simple and effective.

Recent Developments in Telephone Equipment

Research and development is being carried out continually by the Post Office, at their research station at Dollis Hill, and by the members of the telecommunications industry who supply the equipments into ways in which the public telephone service can be improved both technically and economically. As a result of this activity a number of new pieces of equipment have recently gone into service and a brief description of some of these is given below.

Group Routing and Charging Equipment (GRACE)

The lower charges for telephone calls which were recently introduced flow from a new system under which exchanges will be grouped and all exchanges in a Group have the same charges. The simplified system will pave the way for automation of the trunk service which will follow later and effect big operational and administrative economies.

For trunk dialling by subscribers, a system of national numbers will be required.

Post Office engineers have devised new equipment to handle most of the automatic operations involved in this innovation, and it is known as GRACE—Group Routing And Charging Equipment. It will interpret numbers dialled, route the call to the exchange required, dial the numbers wanted -often over hundreds of miles— and control the charging of the calls.

General Description

The Group Routing and Charging equipment performs automatically the functions which the trunk operator now does manually in setting up a call. It notes the number dialled by the caller and decides the correct route and charge rate. It then selects a route to the required exchange, and when the called subscriber answers it records the charge on the callers exchange meter.

The equipment has three main parts, Call Chargers, Registers and Translators, the number of these varying with the number of calls to be dealt with.

To make an automatic trunk call the caller dials the national number of the person to whom he wishes to speak. The first digit of all national numbers is 'O' and receipt of this digit by the exchange equipment causes the call to be connected to a Call Charger.

The Call Charger associates itself with a Register and the remaining digits of the number are received and stored in the Register which forms a memory for this information during setting up of the call. The first 1, 2 or 3 digits received by the Register identify the distant group. It is the function of the Translator to inspect these digits and deduce from them the route and charge rate for the call. The Translator is the brain of the system and incorporates a permanent memory giving details of the routes and charge rates for calls from the originating exchange to all other groups in the country.

The digits stored in the Register are referred to the Translator which examines them, identifies the exchange code, selects the appropriate route and charge rate for the call and passes this information back to the Register.

The information is in the form of a charging rate digit and from one to five routing digits. To avoid having to provide storage capacity in the Register for all these digits at once, it is arranged that they are passed to the Register one at a time as required. The Register uses the digit supplied by the Translator to further the setting up of the call and then makes a fresh demand for another digit, indicating to the Translator which one is required. The time taken by a Register to use a digit is far greater than the time the Translator takes to supply it. The Translator is therefore freed between demands for use by any other Register requiring its services.

The first digit returned to the Register from the Translator is used to select the appropriate charging rate in the Call Charger. Subsequent digits are used by the Register to operate switches in the originating and distant exchanges to complete the connexion. When the connexion has been completed the Register is released and made available for use with other Call Chargers in setting up further calls.

The Call Charger remains connected throughout the call and when the called subscriber answers, levies the charge by operating the callers exchange meter periodically at intervals of time depending on the distance between the calling and called subscribers charging groups.

The equipment has been developed in two different forms, one using conventional electro-mechanical components and the other using electronic techniques. The electronic version will be employed in the first installation at Bristol and the following description refers to this type of equipment.

The electronic equipment is based on the use of cold-cathode tubes. The 'voltage transfer' method of using cold cathode tubes has been adopted. With this method the output potential from one tube is sufficient to strike another tube, the potential being controlled by rectifier gates. Rectifier gating permits combinations of tubes to control the striking of another tube without mutual interference. Miniature selenium rectifiers are used to form the gates.

Register

Digits received by the register are in the form of dial pulses. These pulses are counted on the incoming counter which is a chain of cold-cathode tubes. When a pulse train ceases the number of pulses counted is stored on five tubes in 'two out of five' code by striking the appropriate two tubes. The incoming counter is then reset ready to count the next digit and a second counting chain is stepped to keep check of the number of digits received. Up to nine digits may be received and each is stored on a separate group of five tubes, requiring 45 tubes for digit storage. Digits to be transmitted by the register are received from the translator and stored temporarily on the counting chain tubes of the outgoing counter. The register then generates pulses—like those generated by a dial—and as each pulse is transmitted it is counted by the outgoing counter until the correct number has been transmitted. The register then receives the next digit to be transmitted from the translator and as soon as the 'inter train pause' switching period has expired, this digit is transmitted.

Translator

The electronic translator can deal with a register in a few milliseconds. This high speed of operation permits the translator to be connected to a register at regular intervals, whether the register requires the translator or not. The translator acts as a common control. Each time it is connected to a register it can decide what action, if any, is required. This common control arrangement permits a simple design of register since the register is merely a slave acting on instructions from the translator. Registers are connected in sequence and up to 40 may be controlled by one translator, each register being connected once every two-thirds of a second.

It will be appreciated that since a translator is common to as many as 40 registers a faulty translator could seriously disrupt service to subscribers. It is arranged therefore that any mis-operation is detected by check circuits and causes a standby translator to take over service automatically.

Thus in the rare event of a translator developing a fault only one call can be affected.

CONSTRUCTION

The equipment is mounted on standard sized telephone racks approximately 10ft 6in high and 4ft 6in wide. Racks for the electronic equipment are box section and divided into cubicles catering for 20 registers per rack of three translators per rack. Equipment within a cubicle is mounted on 'leaves' which are hinged together at the rear to form a 'book'. Cold-cathode tubes are mounted at the front of the leaves so that their glow can be observed through the red tinted front cover. Access to the equipment for fault location can be obtained by opening the front cover and parting the leaves at any desired position, free space at either side permitting the leaves to be opened to a wide angle. If work has to be carried out on the equipment it may be unplugged and removed from the rack.

INITIAL INSTALLATION

The initial installation at Bristol will comprise 66 registers and three translators. Two of the translators will be in service normally, the third acting as standby for either of the other two. This equipment can handle about 6 000 calls per hour, and can establish connexion to the called subscriber within a few seconds of completion of dialling.

The system has been devised by Post Office engineers and equipment for the initial installation has been designed and developed in co-operation with the General Electric Company Ltd.

Magnetic Drum Director

Known as the Magnetic Drum Director, this electronic telephone controlling equipment has been developed by Automatic Telephone and Electric Company Limited, and is a noteworthy technical improvement on the existing electro-mechanical equipment in use in London and other large cities throughout the world. Its functions are, however, basically the same—it must 'store' the dialled numbers, translate them into routing directions, and ensure that calls are put through in the shortest possible time.

The magnetic drum around which the new electronic director has been constructed consists of a bronze cylinder some twelve inches in diameter and about three inches deep. Its working surface is coated with approximately two ten-thousandths of an inch of nickel which forms the medium for the magnetic records. The records consist of sequences of magnetic 'dots' which can be packed closely enough together to obtain 100 dots to the inch of circumference and at least ten similar circumferential 'tracks' of dots to the inch of depth. The total capacity of the drum for dots is at least 110 000. In operation the drum rotates at about 1 800rev/min and dots may be recorded or reproduced at a rate of about 100 000 dots/sec.

The magnetic dots are coded into machine language to allow numbers and instructions to be built up in binary code, thus four dots are required to record a single digit, and seven sets of four dots to record a seven-digit number such as would be dialled by a caller.

The drum records are used in two ways. Some of the 30 tracks are used for remembering the translations likely to be required—about 700 in all. These records are virtually permanent in the sense that they are changed only when, for administrative reasons, it is necessary to alter a translation. The remainder of the tracks are used as control tracks—some provide the means for synchronizing the operation of the complete equipment—the others provide individual memories for switch control units of which there are 114.

The electronic equipment associated with the drum contains a scanner, driven by the synchronizing tracks, which, among other things, causes each switch control unit to be associated in turn with the particular portion of one of the tracks allotted to it. The switch control memory tracks provide means for keeping a running record of the state of each control unit and by continuously checking this running record the electronic equipment can determine what action is next required in each case.

The running record is kept up to date merely by recording the most recent record in place of the old one. In this electronic equipment and the drum can be time shared over 114 switch control units in as little as 17 milliseconds each without interfering with any other. Moreover, each of the 114 control units can be rescanned every 17msec so that changes of state of up to 60 changes per second are recognized. This permits considerable economies in apparatus and is one of the reasons for developing the trial equipment. The translation tracks are referred to whenever a switch control unit reaches an appropriate condition.

Maintenance Equipment

RECENT DEVELOPMENTS IN TELEPHONE EQUIPMENT

The London automatic telephine network serves some 1.35 million exchange lines, having about 2.25 million telephone instruments, connected to 235 automatic and 56 manual exchanges by nearly 2.5 million miles of pairs of paper-insulated copper wire in the lead-sheathed underground cables. About 35 million dialled telephone calls are made over the network every week.

Three automatic devices have been developed in London to aid the maintenance engineers in reducing the fault liability of the whole system, so reducing the cost of maintenance.

AUTOMATIC TELEPHONE SUBSCRIBER

The possible combinations of lines and exchange equipment that could be used to complete a telephone call from a P.B.X. extension in the City to a subscriber on, say, Harrow exchange amount to several hundred thousand. It is almost impossible to repeat the particular combination that gave rise to a complaint of call failure and so this device has been developed to help in locating the cause of such failures. It makes calls in sequential order to telephone numbers at upwards of 100 other telephone exchanges where the call is received on equipment that simulates the answering condition on a telephone call and also provides means of verifying the called number. Should a call fail for any reason, the calling equipment holds the connexion and draws the attention of the exchange maintenance engineer to the condition. The call is then traced and the defect that caused the failure is located and put right.

LINE INSULATION TESTER

Incipient trouble arising from falling insulation resistance on a telephone line can be detected much more readily by adopting a comparatively high standard of insulation resistance, $2M\Omega$ for subscribers' telephone lines. The defect causing the trouble, generally a small crack or hole in the lead sheath of a telephone cable, can then be located before the insulation resistance has fallen to such a low value, due, for example, to water entering the cable and saturating the insulating paper during a period of wet weather, that the line becomes unserviceable. To achieve this objective, tests must be made once each week and manual testing methods are out of the question. This automatic device tests each telephone line connected to a telephone exchange in numerical order and it will record the telephone number of each line that fails to pass the test. Lines are tested at the rate of about 600 per hour and an analysis of the route taken by those cable pairs serving the lines that failed the test indicates that section of cable in which incipient trouble is developing. It can then be located and cleared, so preventing a cable fault or a complete cable breakdown from occurring.

AUTOMATIC TESTING OF SUBSCRIBERS' TELEPHONE LINES AND INSTRUMENTS

Testing of telephone lines and instruments is at present done manually from a test desk in the exchange. This device enables a faultsman or an installer to make a complete test of a telephone line without requiring any co-operation from the exchange staff. The test is completed in 15 seconds and any faulty condition that may be found is advised to the faultsman by a verbal announcement.

Short News Items

SIMA (Scientific Instrument Manufacturers' Association) announce that their 1958 Convention will be held at the Majestic Hotel, Harrogate, from 6-9 November. Mr. P. J. Ellis, O.B.E., of R. B. Pullin & Co Ltd, the Honorary Treasurer of SIMA, has been nominated Chairman of the Convention Committee whose task is to settle a programme and organize the event. This will be the seventh annual convention but the first to be held in the North, the previous ones having been at Eastbourne.

The Metal Physics Committee of the Institute of Metals is organizing a one-day symposium on 'Metallurgical Aspects of Semi-Conductors,' to be held on Tuesday, 25 February, at the College of Technology, Gosta Green, Birmingham. A number of short papers will be presented but will not be available in advance of the meeting. Visitors will be welcome; tickets are not required. In conjunction with the meeting, an exhibition is being arranged at the College by the Royal Radar Establishment, Great Malvern. This will be open on 25 February and also on the evening of Monday, 24 February.

The Institution of Electrical Engineers announces that at a meeting held on 20 December, Group B—The British Group for Computation and Automatic Control of the British Conference on Automation and Computation—was formally constituted, comprising some twenty-three learned societies as its members. Mr. T. E. Goldup, C.B.E. (President of The Institution of Electrical Engineers), was elected Chairman of the Group, with Messrs. J. F. Coales, O.B.E. (Society of Instrument Technology), and E. M. Renals (Institute of Cost and Works Accountants) as Vice-Chairmen. Mr. J. D. Green (Institute of Chartered Accountants in England and Wales) was elected Honorary Treasurer, and Mr. W. Bamford (The Institution of Electrical Engineers) Honorary Secretary, and the offer of The Institution of Electrical Engineers to provide secretarial services for the Group was accepted.

The International Instrument Show to be held at Caxton Hall during the last week of March will be almost one-third larger in floor area than the 1957 exhibition and will occupy the Great Hall and Court Room. Tickets are available on request from the sponsors, B & K Laboratories Ltd, 57 Union Street, London, S.E.1.

An exhibition of Airmec Electronic Instruments will be held at the Napier Hall, Vincent Street, London, S.W.1, from 24-27 March. The complete range of Airmec electronic equipment, including several entirely new instruments, will be shown. Tickets for admission will be forwarded on request to Airmec Ltd, High Wycombe, Bucks.

The European Broadcasting Union has announced the introduction of a periodical entitled The E.B.U. Review. The new publication, which replaces the E.B.U. Bulletin, will appear in two parts, Part A Technical and Part B General and Legal. Separate English and French editions will be published. Part A is to be issued in January, March, May, July, September and November, and Part B in February, April, June, August, October and December. In addition to authoritative articles dealing with particular aspects of broadcasting engineering, Part A will give up-to-date information about the situation in all of the frequency bands allocated to sound and television broadcasting, and also about plans and achievements in these fields in Europe and elsewhere. A specimen number of Part A is available on request to the Technical Centre of the European Broadcasting Union, 4 rue de la Vallée, Brussels. Single numbers of Part A cost 30 Belgian francs, the annual subscription being 150 Belgian francs. The combined subscription to Parts A and B is 300 Belgian francs, all the foregoing prices including postage by surface mail.

The BBC has placed orders for thirty television camera channels for the new Television Centre now being built at White City, London. Fifteen channels will be supplied by E.M.I. Electronics Ltd and fifteen by Marconi's Wireless Telegraph Co Ltd. The cameras will use 4½in tubes of the image orthicon type. This type of camera tube has been decided upon after extended trials under the conditions of operation that will be encountered. It is hoped that three of the studios at the new Television Centre will be brought into operation during 1961, and a fourth early in 1962. Of the thirty camera channels ordered, four working channels and one spare channel will be installed in each of the two larger of these studios, and three working channels and one spare channel in each of the other two studios. Others will be used as spares, and the remainder will be used for equipping additional studios as required.

Southern Instruments Computer Division have recently shipped a Film Analyzer to the Canadian Defence Research Board. The Film Analyzer is an automatic analogue computer capable of various computations on data supplied in the form of continuous lines on film or paper.

Chapman & Hall have been appointed publishers of books to the Institute of Physics, with the exception of those books already published by Edward Arnold Ltd. The Institute will continue to publish The Journal of Scientific Instruments, The British Journal of Applied Physics, and all supplements thereto.

Camp Bird Industries Ltd announce the formation of a new company in the audio field. This will be Electronic Reproducers Ltd, and it will be primarily concerned with the manufacture of crystal cartridges. The works will be sited at Bletchley, Bucks, and the company will be under the Chairmanship of Mr. P. J. N. Collaro. Dr. F. E. Templeman will be Technical Director.

The Institute of Physics at the University of Lund, South Sweden, is at present engaged in assembling an apparatus that is believed to be the largest of its kind in the world. It is an electron synchrotron for a tension of 1 200 000 000V and is to be used for studying heavy mesons and hyperons.

Mr. J. O. C. Whelans of the Mullard Radio Valve Co Ltd was awarded the M.B.E. in the New Year Honours List. Mr. Whelans is assistant to the Manager responsible for production activities of receiving-valve manufacture in the company's Southern factories, including those at Mitcham, Whyteleafe and Hove. He has been with the company for thirty-two years and is one of the leading technologists in receiving-valve manufacture.

Lieutenant-Colonel E. N. Elford, Manager of the Radar Division at Marconi's Wireless Telegraph Co, was awarded the O.B.E. in the New Year Honours List. This honour has been conferred in recognition of his work in connexion with the radar defences of Great Britain.

Vice-Admiral J. W. S. Dorling, C.B., M.I.E.E., is to retire from his appointment as Director of the Radio Industry Council, on medical advice, but not with effect until 31 October 1958. He was the Council's first Director, being appointed in 1946, immediately after the formation of the Council as the co-ordinating body of four associations in the industry.

Dr. H. W. Melville, Secretary of the Department of Scientific and Industrial Research, has been appointed to succeed Sir Ben Lockspeiser, K.C.B., F.R.S., as one of their delegates to the Council of the European Organization for Nuclear Research (C.E.R.N.).

LETTERS TO THE EDITOR

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

Electronic Filtering Circuit

DEAR SIR,—During the design of equipment containing a considerable number of pulse circuits it was found that conventional power supply filtering and decoupling was insufficient to prevent interaction between different circuits. However, d.c. stabilization was not required and the extra expense of regulated power supplies not warranted. Since the existing supplies had ample reserve of current it was possible to incorporate an electronic filtering circuit.

When measured, the effective a.c. impedance of the circuit was under 0·5Ω, and it reduced the ripple to less than 50mV in a high tension system using three conventional paralleled power supplies of 250V, 125mA each. This agrees with the calculated performance. Effective by-passing of R_k can further improve the performance if necessary. The value R_k is made adjustable so that the cathode current of the shunt valve can be made large enough to allow fluctuations in the external circuit to be cancelled. The current drawn should not exceed 70mA unless a further shunt valve is added.

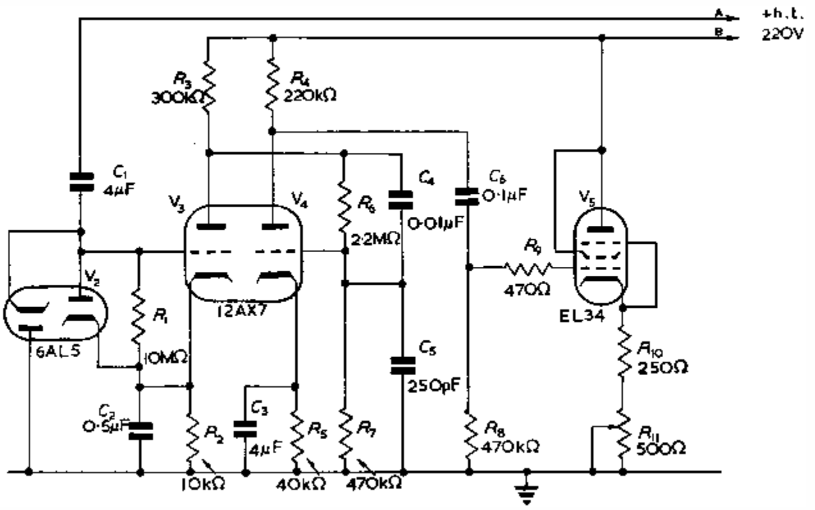


Fig. 1. The electronic filter circuit

The circuit (Fig. 1) consists of a two stage amplifier driving a shunt valve which presents an effective a.c. impedance of $(1 + g_m R_k)/A g_m$ ohms, where A is the gain of the amplifier and g_m and R_k apply to the shunt valve. Since there is feedback, phase relationships are critical. Direct coupling is used in one stage to reduce the phase-shift and the time-constants of the other coupling circuits chosen to avoid low frequency instability. The input is taken directly from the h.t. supply line at the point where the most complete filtering is required: i.e. at the common h.t. distribution point. For this reason two h.t. leads are shown; a potential lead marked A, and a current lead marked B. This arrangement increases the effectiveness of the filtering many times by avoiding the presence of a common impedance in the h.t. lead. Such an impedance may only be a fraction of an ohm (fuses, etc.) but it is of the same order of magnitude as the effective a.c. output impedance and would unnecessarily add to it. The diodes included at the input prevent line voltage surges or steps of either polarity from biasing the first valve for long periods beyond its grid-base; without, however, restricting the maximum performance.

This circuit has now been in continuous operation for over a year and has given completely satisfactory performance.

Yours faithfully,

R. G. T. BENNETT
C. S. L. KEAY
Physics Department,
University of Canterbury,
New Zealand.

"Dekatron"

DEAR SIR,—In your issue for November 1957 reference is made in an advertisement to dekatron tubes (page 61).

The writer is instructed to point out that "Dekatron" is a Trade Mark which denotes cold-cathode counting tubes manufactured by this company exclusively. We are concerned to maintain the distinctive character of this word, which should only be used in this sense.

Yours faithfully,

G. H. HILLIER,
Patents Department,
Ericsson Telephones Ltd.

Binary Coded Digitizers

DEAR SIR,—Mr. Tiffany in his recent very interesting article on Wind Tunnel Instrumentation, in your November issue, appears more pessimistic over the resolution and performance of Binary Coded Digitizers than development warrants.

In fact, with mechanical digitizers and torques of below 2oz in (in selected cases below 1oz in), disks with a resolution of 1 200 unambiguous coded outputs over 320° are already commercially available. No doubt if disk diameters in excess of 4½in were acceptable, higher resolutions than this would be obtainable.

But apart from the mechanical digitizers, there are non-contact type digitizers, such as the optical, in existence, which have—in the laboratory—given 1 : 10 000 over 360° on 6in disks and 1 : 1 000 over 360° on 2in disks, and although these are not yet commercially available there is great hope that by the end of 1958, by one technique or another, these resolutions will be accomplished. Being non-contact types, torques in the order of a few g.cm should be achieved and, of course, the life of digitizers should be considerably enhanced. These advanced types of digitizers should be capable of competing with the high speed chart recorders.

Yours faithfully,

R. F. COLVILE.
Hilger & Watts Ltd.

The author replies:

DEAR SIR,—I note with interest Commander Colvile's remarks regarding the possibility of higher resolution digitizers, but would refer him to the less compressed discussion of this point in a later article[1]. In the case of mechanical digitizers, diameters above 4½in are decidedly not acceptable and considerable compression would be desirable. The modest improvements in torque and resolution reported are welcome, but far removed from requirements existing now.

The ancillary equipment required to read out digitizers of the optical type is far more involved, and expensive, than many potential applications could tolerate. For multi-channel readout it is essential that the equipment associated with individual channels be cheap, simple and readily replaced. A complex, rapid central scanner or multiplexer may be justifiable if sufficient channels can be scanned and processed at a suitable rate.

I would be the first to welcome a commercially available, British digitizer of high resolution, but the phrase "in the laboratory" has a ring we have come to recognize as ominous. The equipment I have described has been in daily use for almost two years now and at the time of conception nothing at all was available commercially in spite of many projects 'in the laboratory' at that date. The need for these devices is now urgent.

Yours faithfully,

A. TIFFANY,
Aircraft Research Association Ltd.,
Bedford.

1. TIFFANY, A. A Simple Shaft Digitizer and Store. *Electronic Engng.* 29, 568, 1957.

BOOK REVIEWS

Light, Colour and Vision

By Yves Le Grand. (Translated from the French by R. W. G. Hunt, J. W. T. Walsh and F. R. W. Hunt). 512 pp. 125 figs. Demy 8vo. Chapman & Hall Ltd. 1957. Price 63s.

NOWADAYS many electronic engineers are perforce making their first real acquaintance with the difficult field of colorimetry, so that this book makes a particularly well-timed appearance.

Commencing with a comparatively terse and factual survey of the fundamentals of vision and photometry, the author then proceeds to explain at length the experimental observations upon which modern colorimetry is based and the physiological theories which have been advanced to explain these observations. Although initially only an elementary technical knowledge is assumed in the reader the text rapidly advances to a higher standard and is then copiously provided with references to books and articles from which the industrious may benefit. Unfortunately, it often appears that work thus described is considered as being of doubtful validity or that it is in conflict with other work which is also referred to, the reader being left to evaluate the evidence and come to a conclusion, though this could seldom be done by the inexpert.

Though the book is a positive mine of information, it is not an easy one to read. This is very largely because it is sometimes only too apparent that, as is stated by the author in his preface, the translation closely follows the French text. However, the information is there, and thanks to an excellent index it can readily be found. The book should be of particular value to those who are required to devise colorimetric experiments in new fields, for the probable causes of error are all described.

The book contains many valuable tables and diagrams and includes a bibliography of some thousand items. It will be of great value as a work of reference.

C. F. Chapter

Pressure Measurements in Vacuum Systems

By J. H. Leck. 144 pp. 79 figs. Demy 8vo. Published for the Institute of Physics by Chapman & Hall Ltd. 1957. Price 30s.

SOME years ago, the reviewer advocated in these columns the publishing of specialist works in vacuum physics rather than the frequent publication of all embracing texts which were too general to be of value to advanced workers. Thus it was with some sympathy that your reviewer read the opening statement to Mr. Leck's work that "Whilst several books have been written dealing with vacuum technology in general, none has yet given a detailed treatment of the measurement of pressure, nor am I aware of any text devoted exclusively to this subject." Apart from Dushman's detailed consideration of vacuum gauges covering 129 pages in his "Scientific Foundations of Vacuum Technique" which should not be ignored, Mr. Leck's statement is correct, and it is against this background that one judges how well he has used his opportunity.

In all there are six chapters devoted to mechanical manometers; thermal conductivity gauges, ionization gauges, the Knudsen radiometer gauge; surface reaction techniques and gauge calibration.

Considering the text as a whole one finds it difficult to understand why Mr. Leck never included a section on leak detection. It is true that types of leak detection apparatus are used which are not based on pressure measuring devices, but conversely nearly all the known forms of vacuum gauges are used for leak detection purposes either by simply measuring the rate of pressure rise in a system with a manometer or with dynamic procedures involving probe gases and hot wire or ionization gauges. Like the recipe attributed to Mrs. Beeton for jugged hare, one must first obtain a "vacuum" before its pressure can be measured accurately, and the fact that the pressure measuring device often with simple modification can be used to provide a leak free system is of great value.

One would also expect to have found a definition of what is meant by pressure, i.e., other than the useful technical data given on pressure units. Also, some of the relations of the kinetic theory might have been derived in the text instead of the distressing habit of referring the reader to one of three theoretical works for many of the equations quoted. Understandably the author has tried to free his work from extraneous matter, but a few pages devoted to kinetic theory would have been of value.

Some general items which should be corrected in future texts are: the use of the phrase "reflected molecules" on page 34 where the energy exchange between gas molecules and heated surfaces is discussed. It would be better to use the term *restituted* or *re-emitted* since the molecules may linger on the target for a finite time. The photograph in Fig. 3.16 of a Philips gauge head is indistinct. There is no reference in the Chapter on Radiometer Gauges of a useful paper by Steckelmacher (*Vacuum 1*, 266, 1951) in which early work by Spivak is corrected and extended. The simple glow discharge tube which is still used for indicating backing pressure is not dealt with.

Mr. Leck has usefully given the probable accuracy of the pressure indication for each type of instrument based on constructional and electrical factors and attention is also given to the effect on the gauge response of the nature of the gas being measured. However, there are cases, particularly in research, where it is insufficient merely to know the total pressure of the gas in the system however accurately, but one must also know the nature of its components and their partial pressures. Obviously chemical and other reactions occurring at the same total pressure will be quite different if in one case the residual atmosphere is a water vapour/air mixture and in another a hydrocarbon/air mixture, and so on. For example, the reduction of an oxide may critically depend upon the relative values of the partial pressures of hydrogen to water vapour in the residual gas. It is in this respect that the mass spectrometer (particularly the Omegatron type with its small volume) is such a valuable tool for indicating the nature and abundance of the residual gas components, and this might have been discussed briefly.

The chapter on surface reaction techniques is a useful pointer to the problems facing vacuum engineers in measuring "ultra high vacuum." When one achieves a vacuum of the order of 10^{-10}mm Hg collisions between gas molecules in transit have little importance (the mean free path of air molecules in air is 310 miles at this pressure!) but one may still obtain measurable effects from the gas molecules striking internal surfaces. Thus it will take several hours for a monomolecular layer to form on a clean active surface at this pressure. Apart from the unreliability of pressure instruments in this range, the use of pressure units is not essential and possibly high vacuum physicists will finally agree to a unit of measurement based on the monomolecular layer formation time using a given gas –target combination.

Often in industrial processing it is not necessary that the vacuum gauge should read the true pressure of the residual gas providing the gauge gives a repeatable value at which it is known that the process proceeds satisfactorily. As the author states in his Preface, instruments based on ionization processes are really density measuring devices, since they depend on the number of molecules occupying the volume through which the electrons pass. Thus to call an ionization gauge a pressure measuring device is a misnomer, which, however, is forgivable providing the conditions of measurement are constant. Also such gauges are commonly used on electronic apparatus where the process is affected by the total cross-sectional area of residual molecules in the path of high speed particles. Such a relation between the quantity measured and its effect on the process being carried out is obviously desirable.

It follows from the views expressed here that "pressure measurements in vacuum systems" is really a part of a more general subject which could be called "Vacuum Measuring Instruments," i.e., devices for assigning a physical value to a rarified atmosphere in terms of the gas pressure, molecular density or rate of impingement of molecules on surfaces. This division also covers the three most important stages in general experience, e.g. (i) the vacuum used in many industrial processes for drying or metallurgy from 20mm Hg down to 10^{-3}mm Hg., (ii) the long mean free path range of importance in thermonic tubes and particle accelerators (10^{-4}-10^{-7}mm Hg) and (iii) the range in which gas bombardment of surfaces is of greatly reduced interest to those studying electron emission and surface chemistry (10^{-7}-10^{-11}mm Hg).

Mr. Leck has described vacuum gauges which can be used in each of the foregoing ranges, but a place still exists for a more comprehensive work on 'Vacuum Instruments' which is not restrained by consideration of pressure measurement. The opinion of your reviewer must, however, be considered alongside the fact that high vacuum technology has become part of the curriculum of many technical colleges and universities, and Mr. Leck's work will be a useful text within the reach of advanced students. The layout of the book and the drawings are of the high standard one associates with publications of the Institute of Physics.

L. HOLLAND

Britain and Europe

299 pp. 40 figs. Demy 8vo. The Economist Intelligence Unit. 1957. Price 15s.

THE main object of this book is to assess the impact of European Free Trade on British industry—whether the changes will be favourable or unfavourable, big or small. The E.I.U. has examined the prospects in detail for each main manufacturing industry, and it is clear that manufacturers as a whole will be considerably better off if there is a Free Trade Area than if the Common Market alone develops. Of the industries studied, covering some 85 per cent of total manufacturing industry, those likely to be adversely affected by 1970 within a Free Trade Area account for less than 15 per cent of the output.

The industries which will probably be better off within a Free Trade Area include: motor vehicles, chemicals (output estimated at 10-20 per cent more by 1970 than if there is a Common Market only); wool, electrical engineering, general engineering, rubber manufactures (output 5-10 per cent more); steel, hosiery, clothing (output up to 5 per cent more). The following industries should also be gainers: non-ferrous metals, metal manufactures, aircraft, shipbuilding, oil refining, building materials, glass, scientific instruments, sports goods. Cotton, rayon, paper, leather, and watches and clocks are likely to lose from Free Trade; the prospects for china, footwear and toys are also somewhat doubtful. Railway engineering, jute manufactures and furniture should be little affected.

Apart from the greater intro-European trade that will follow from the abolition of tariffs and quotas and from fairer competition, there should be a general quickening of the growth of national incomes and production. The industry studies show that the 'leaders' such as chemicals, motor vehicles and electrical engineering will be considerably stimulated, and in all sectors the efficient, faster growing units of production will benefit.

FBI Register 1958

1 138 pp. Royal 8vo. 30th Edition. Kelly's Directories Ltd. and Iliffe & Sons Ltd. 1957. Price 42s.

IN the following words the President of the Federation of British Industries, Sir Hugh Beaver, introduces this new edition. "I am glad to commend the 30th edition of the FBI Register to its readers in all parts of the world. The object of this annual publication is to spread the widest possible knowledge of the great range of British manufactured goods, so that prospective buyers will not only know what Britain makes but will also have a ready means of getting into contact with British firms supplying the goods or services they are seeking."

This comprehensive guide to a substantial cross-section of British industry lists the products and services of over 7 500 member firms under more than 5 400 alphabetical headings. In addition to the Classified Buyers' Guide there are seven other sections in the Register, giving addresses of companies and firms, and valuable information about Trade Associations, proprietary names, trade marks, etc. As the only authorized directory of the Federation of British Industries, the FBI Register is compiled by the publishers in close collaboration with the Federation.

BBC Handbook 1958

288pp. Demy 8vo. The British Broadcasting Corporation. 1957. Price 5s.

THE first edition of this handbook was published in 1928. The new edition goes a long way towards meeting the public interest in how the BBC is run, what it is doing and aiming to do.

A table of figures shows that the cost of television broadcasting increased from £2 675 per hour in 1956 to £3 256 per hour in 1957. Sound broadcasting costs also went up, from £540 per hour in 1956 to £575 in 1957.

The handbook shows how, subject to the requirements of the BBC Charter, the Corporation enjoys complete independence in the day to day operations of sound and television broadcasting.

One of the most impressive tables in the handbook lists no fewer than 50 countries throughout the world which regularly rebroadcast BBC overseas programmes.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

AUDIO OSCILLATOR

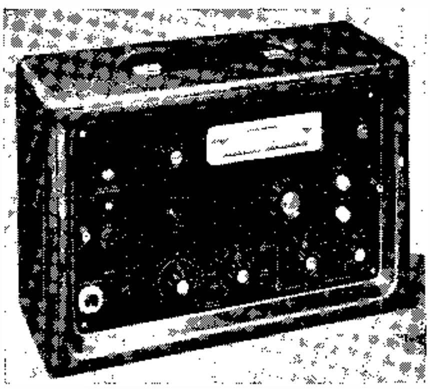
(Illustrated below)

Wayne Kerr Laboratories Ltd, Roebuck Road, Chessington, Surrey

THE wide range audio-frequency oscillator type S.121 is a high grade laboratory instrument providing a stable, controllable signal within the frequency range 10c/s to 120kc/s.

The novel dial display is designed for simple and rapid operation and permits either the selection of major intervals by means of switches, or the continuous fine control of frequency on an open horizontal scale.

Two alternative outputs are provided. An accurate 600Ω output incorporates a step attenuator covering the range +10dB to −70dB in 1dB steps on a reference level of 1mW. The second output provides a continuously variable control of voltage



from 0 to 30V in five decade ranges into an impedance of 3·3kΩ. The stability of the output level is to within 0·2dB over the whole frequency range.

The total harmonic distortion is better than 0·2 per cent above 300c/s.

Despite its high performance the instrument is compact, light and portable.

CRYSTAL OSCILLATOR FREQUENCY DIVIDER

(Illustrated above right)

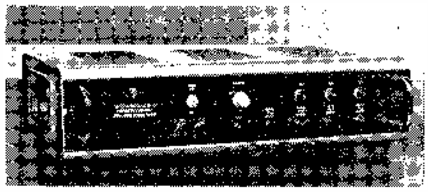
Ericsson Telephones Ltd, Beeston, Nottingham

THIS is a general purpose instrument for laboratory and workshop use which has a wide range of applications such as the calibration of oscilloscopes and the standardization of oscillators.

The crystal oscillator and frequency divider can also be used as a source of time bases for use with any of the Ericsson range of unit construction equipment.

It provides four substandard crystal controlled frequencies, the basic frequency of the 100kc/s crystal oscillator having an accuracy of ·005 per cent. This fact, along with the unique circuit design which incorporates fully automatic amplitude stabilization, ensures a high degree of frequency stability.

Four frequency outputs are available, at front mounted Belling-Lee terminals, and these consist of a 100kc/s sine wave



and three semi-square wave outputs at frequencies of 10kc/s, 1kc/s and 100kc/s. The three semi-square wave frequencies are derived from the 100kc/s crystal oscillator by three multivibrator frequency dividers consisting of one double triode per stage and these select the required sub-harmonics.

Three pre-set potentiometers are provided to adjust the division ratio of the multivibrator circuits. These are accurately set before the instrument leaves the works and under normal circumstances should require no further attention.

The crystal oscillator and frequency divider type 124 can be adjusted for use with an a.c. mains supply of 100 to 125V or 200 to 250V at 40 to 60c/s a.c.

The instrument has its own built-in power supply employing standard full-wave rectification and retains its stability over a mains fluctuation range of ±10 per cent of nominal.

EXTENSIBLE LEADS

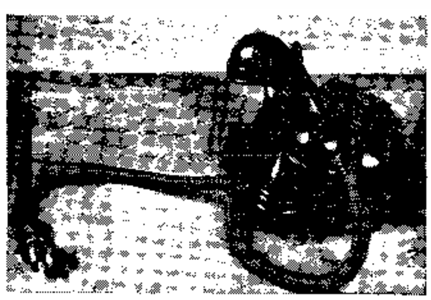
(Illustrated below)

Creators Ltd, Plansel Works, Sheerwater, Woking, Surrey

CREATORS LTD have introduced a new type of extensible lead, extremely small and light, as shown on the photograph.

These leads have an extension ratio of 4 : 1 and up to five conductors can be used in their construction, which has made them acceptable for use in the aircraft and electronic industry, and for use on counting machines etc. where power units can be detached from the main chassis without disconnexion.

They are also used on telephones, microphone hand sets, electric irons, kettles, clippers, etc. A large manufacturer of clocks has found that a 3in length of this cable attached to the back of an electric clock eliminates the problem of controlling slack cable when adjusting the clock.



The outside braiding is available in acetate or nylon, the latter being highly resistant to wear and abrasion.

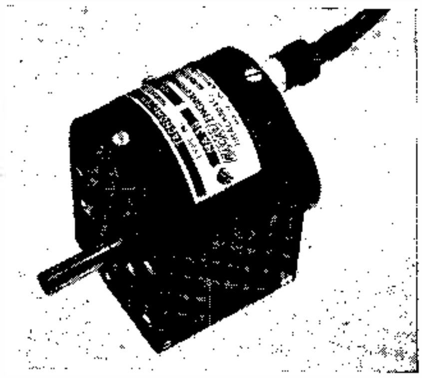
TACHOMETER

(Illustrated below)

Racal Engineering Ltd, Western Road, Bracknell, Berkshire

WHEN used in conjunction with Racal digital frequency meters, type SA.20 or SA21, this tachometer may be used for direct reading in shaft speed measurements, in the range 100/20 000rev/min.

The MA.38 consists of rotor with 60 teeth rotating in a magnetic field, so that an output of 60 pulses is obtained for every complete revolution of the rotor. This output is then fed to the frequency meter which can sample the shaft speed in rev/min.



Apart from the shaft speed measurement as such, the MA.38 may be used for other types of measurement. This may be accomplished by transforming linear or other movement to rotary action, and has uses in flow and process control etc.

The MA.38 may be used in conjunction with other digital frequency meters using a one second time base.

SCANNING UNIT

(Illustrated above right)

Isotope Developments Ltd, Beenham Grange, Aldermaston Wharf, Nr Reading, Berkshire

THE I.D.L. scanning unit type 701 employs a strip chart recorder for automatic plotting of pulse height spectra, in conjunction wth scintillation counting equipment incorporating a single channel pulse analyser and a ratemeter.

The chart recorder drum is direct-coupled to a 355° potentiometer. This varies the discriminating voltage of the pulse analyser, while the pen movement (controlled by the ratemeter) follows the corresponding pulse rate.

The scanning time can be switched to 10min or 1hr, and the chart length per scan is 6in. Each 6in record represents a scan over either the complete voltage range or the upper half of the voltage range, as selected by a 2-position switch.

The servo-type precision 10kΩ potentiometer has a linear accuracy of ±0·5 per cent. Calibration marks can be put on the chart at any required point by means of a switch which makes the recorder read full scale or zero. The exact zero of the potentiometer is indicated by means of a 'witness' mark.

A recording switch permits the upper half of the normal scan to be separately expanded into a full 6in record, thus giving the equivalent of a 12in chart length per full scan. Pre-set potentiometers are provided for accurately defining the voltage scanned.

The instrument is designed for standard Post Office rack mounting with dust cover, but it is also suitable for bench use. Special practical features of the design include a unspillable inkwell for the recorder, and quick-drying ink which does not smudge or grow mould. Mains input plugs at the rear permit connexion to existing equipment without the need for an extra mains supply point. The light-weight door swings completely clear or can be quickly removed to give easy access. A paper tear-off device is fitted, but the chart can hang freely through a slot in the bottom of the door, or can be wound on to a take-up spool.

PRECISION INSTRUMENT SWITCH

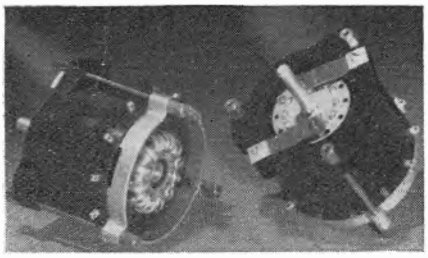
(Illustrated below)

W. G. Pye & Co. Ltd, Granta Works, Cambridge

AN instrument switch of completely new design is undergoing a test to destruction at the factory of W. G. Pye & Co. Ltd. Although 17 million operations have been performed, which is equivalent to some thirty years of service, the performance of the switch remains virtually unchanged—in fact, the stability has improved within the rather narrow range specified.

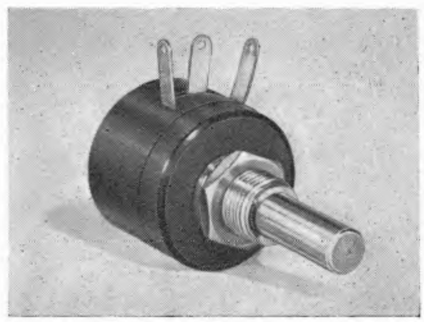
No maintenance was necessary and test conditions have been rather more severe than would be encountered in service.

The overall switch resistance is approximately 0·7mΩ and the variation of contact resistance is $\pm 20\mu\Omega$. The overall resistance drift during the life of the switch is $\pm 50\mu\Omega$. Thermal e.m.f. is 0·008μV/°C with uniform change of temperature. Temperature gradient across the switch produces a maximum of 0·025μV/°C.



The stationary contacts are mounted in a strong insulating frame and the set of brushes makes a bridge contact between pairs of adjacent bars in a truly symmetrical manner. The brush itself is composed of several contact springs, each of which is independent of the others and unrestricted in its movement. Low and consistent contact resistance is achieved without high contact pressure and mechanical wear. The moving contacts are totally enclosed.

The stability of switch resistance and the virtual elimination of thermal e.m.f.s make possible improved designs in potential and resistance measurements, giving greater precision with simplified circuits and operating techniques. Long life without maintenance and robustness of construction also allow greater reliability in instruments, where precision requirements are less critical.



MINIATURE POTENTIOMETER

(Illustrated above)

Reliance Manufacturing Co. (Southwark) Ltd, Sutherland Road, Walthamstow, London, E.17

THIS miniature wire-wound potentiometer has been developed as a standard commercial version of the type RVW13 and 14 as called for in the Joint Services Specification RCL121.

It is available in a resistance range (linear low only) of 5Ω to 50kΩ with a tolerance of ±5 per cent above 100Ω and ±10 per cent below 100Ω It has a rating, with the whole element uniformly loaded, of 1W at 20°C ambient uniformly mum working voltage between spindle and track is 500V d.c.

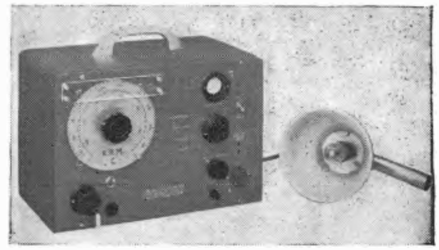
STROBO-TACHOMETER

(Illustrated above right)

C. F. R. Giesler Ltd, Small Hall, River Place, Islington, London, N.1

THIS is a portable instrument for use as an accurate and easily operated stroboscope; and as a tachometer for rapidly measuring the speed of rotating or reciprocating mechanisms.

The control box contains a stable generator of electrical impulses, the frequency of which is under the control of the operator; the impulses are used to trigger the stroboscopic lamp, which will thereupon flash at a rate governed by the impulse generator. The generator pulse-



rate is adjusted by turning a large dial, which is clearly marked in revolutions per minute, and the range of the instrument is further extended by a 'range switch', according to which the readings of the main dial are to be multiplied by 1, 10 or 100.

In addition to the normal lamp supplied with the strobotachometer, a socket is provided to permit of the instrument being used for driving a high powered stroboscope for photographic purposes.

When the instrument is used alone, a shaft-driven contactor unit with phase change can be supplied to drive the stroboscope, and for this purpose an external contactor socket is provided on the panel. This attachment permits for exact synchronization of the stroboscope with the rotating machine under test—thus providing facilities for close examination of the test rotor in any position around the cycle.

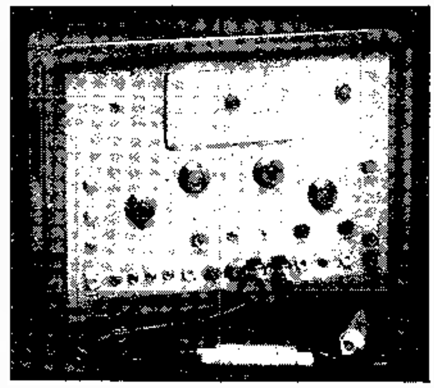
The speed range is 150 to 13 000rev/min. For calibration the instrument is fitted with a 1in cathode-ray tube, which indicates a characteristic wave-form when the instrument is operating at 50c/s or 3 000rev/min. The main rev/min dial is set to 3 000rev/min and a control switch is adjusted until the characteristic pattern is visible on the cathode-ray tube. The instrument is then accurate to ±1 per cent throughout.

SWEEP GENERATOR

(Illustrated below)

Marconi Instruments Ltd, St. Albans, Hertfordshire

MARCONI Instruments Ltd announce the introduction of a new 20Mc/s sweep generator, type TF1099. This high-quality video-frequency sweep generator, in conjunction with a suitable oscilloscope, enables response measurements to be made with a discrimination exceeding 0·01dB. It is primarily designed for the precise alignment of



monochrome or colour-television transmitter and studio equipment, but is equally applicable to radar or television receivers where the highest order of measurement accuracy is required.

The sweep starts at a frequency of less than 100kc/s and extends to an upper frequency that is continuously variable up to 20Mc/s; the lower frequency limit is locked to the start of the sweep in order to ensure a steady display on the c.r.o. The level of the frequency-swept signal is variable from 0·3 to 3V peak-to-peak, and the selected output level is held constant to within −0·1dB—a factor that contributes materially to a high order of measurement accuracy. A 250V time-base waveform, synchronized to the sweep frequency, is provided for connexion to the X-plate of the c.r.o., and a series of marker pips at 1Mc/s intervals is available for superimposing on the display to facilitate accurate frequency identification.

As well as providing facilities for conventional response measurement involving the display of the frequency-swept output of the apparatus under test, the TF1099 includes special circuits for precision differential measurement. For the differential method, the overall gain of the apparatus under test is reduced to unity by means of an external attenuator, and the outputs of the sweep generator and of the apparatus are both sampled, the two signals are then combined and the amplified difference voltage is displayed on the c.r.o.

An alternative version of this instrument, the TF1099/1, is electrically identical, differing only in that it has neither case nor mains lead and is intended for mounting in a standard 19in rack.

LOW VOLTAGE STABILIZED POWER UNITS

(Illustrated above right)

Roband Electronics Ltd, 33 Mountgrave Road, Highbury, London, N.5

THE conventional thermionic valve stabilized supply is not suitable for low voltage high current applications for the following reasons :—

(1) A negative stabilized line is required to operate the feedback amplifier.

(2) A large number of series valves are required to handle the current involved.

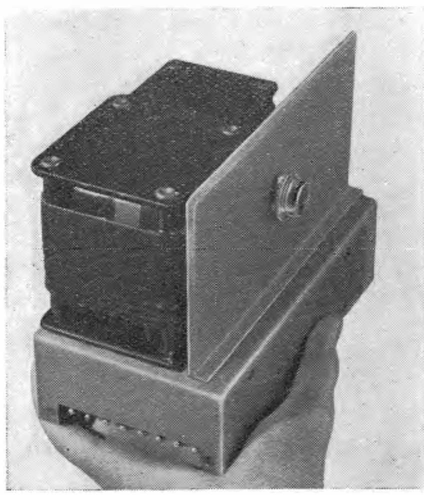
These factors lead to an unnecessarily large and expensive unit which is also wasteful in power efficiency, especially when one considers the heater power which is required for the rectifiers and series valves.

With this in mind, a series of miniature transistorized units have been developed in which no heater supply is required. A fixed output voltage is provided, either rail of which may be earthed, and a current of up to 1·0A d.c. can be drawn.

Briefly, the circuit details are as follows. A bridge rectifier system employing germanium junction rectifiers feeds a transistor circuit, the input stage of which compares and amplifies the difference voltage between a reference source and the output. This is amplified again and then fed into a power transistor in such a way as to reduce the difference voltage to nominally zero. The reference source employs two reference tubes and filtering is such that the stabilized output has less than 1mV ripple content.

The unit may be mounted in either the horizontal or vertical plane by means of brackets which are provided. Special attention has been paid to the component layout and this, coupled with a colour coded wiring system, enables servicing to be effected in the minimum time. A partially enclosed tag strip is provided for the external connexions.

A generously rated C-core transformer and choke, made to RCS215/H2 requirements, ensures that the temperature rise of the unit is kept to a practical minimum. Conservatively rated rectifiers and transistors, and the use of high stability



carbon resistors in place of wirewound types, ensures prolonged trouble-free service.

These units are available for a wide range of output voltages between 1·4V and 24V.

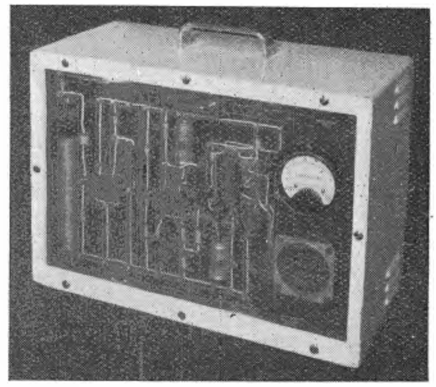
DEMONSTRATION GEIGER COUNTER

(Illustrated above right)

Burndept Ltd, Erith, Kent

THIS instrument has been designed in collaboration with A.E.R.E., Harwell, as a demonstration model for educational, services training and civil defence authorities. The monitor can be used for innumerable instructional experiments to illustrate the operation of gamma radiation monitors.

It is self-contained, mains operated and complete with the usual controls; oral and visual indication of counting rate is also provided. The monitor has a Perspex front panel which is engraved with the multi-coloured counting circuit and the actual components are mounted behind the panel. The engraved circuit is illuminated by edge lighting of the Perspex front panel and additional in-



ternal lighting can be used to illuminate the components. The various circuit waveforms are also shown on the Perspex.

The power unit is built on a separate chassis which slides into the main box and is accessible by removal of the rear cover. A special compartment for the mains lead is provided at the bottom of the carrying case.

The counting circuit comprises:

(a) Signal amplifier,

(b) Pulse shaper and counting ratemeter,

(c) Loudspeaker amplifier.

Standard practice is followed whereby the negative going pulses from the Geiger counter are fed into the control grid of a cathode coupled signal amplifier. The amplifier pulses are fed into a pulse shaper and thence into the output amplifier which drives the loudspeaker.

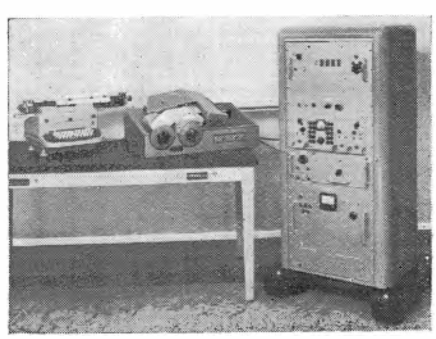
The Geiger counter used is of the halogen quenched type employing a gas mixture of neon, oxygen and bromine. One of the standard college experiments is to plot a counter plateau of this type of tube and the monitor lends itself admirably for such demonstrations.

The monitor is of robust construction contained in a portable cabinet 15in by 11in by 7in and the weight is approximately 27 lb.

TRACE READING EQUIPMENT

Southern Instruments Ltd, Frimley Road, Camberley, Surrey

ON page 614 of the December 1957 issue a description of the automatic trace reading equipment type K1020 was given. The photograph accompanying this description was, however, of the simpler type K1015 equipment. A photograph of the K1020 is reproduced below.



Notes from ——————

NORTH AMERICA

I.R.E. News

ELECTION OF PRESIDENT

Donald G. Fink, Director of Research of the Philco Corporation, has been elected President of the Institute of Radio Engineers for 1958. Mr. Fink succeeds John T. Henderson, Principal Research Officer of the National Research Council, Ottawa, Canada, as head of this international society of 62 000 radio engineers and scientists.

During 1933-1934, Mr. Fink was a research assistant in the M.I.T. departments of electrical engineering and geology. From 1934 to 1952 he was on the editorial staff of "Electronics," becoming Editor-in-Chief in November, 1946. Since 1952, he has been associated with the Philco Corp. as Director of Research for radio, television and appliances.

FELLOWSHIP AWARDS

Seventy-five leading radio engineers and scientists from the United States and other countries were named Fellows of the Institute of Radio Engineers by the Board of Directors at its November meeting held in New York City. The grade of Fellow is the highest membership grade offered by the I.R.E. and is bestowed only by invitation on those who have made outstanding contributions to radio engineering or allied fields.

The recipients of the Fellow award, which takes effect January 1, 1958, include one from Brazil, one from Canada, and three from England—J. A. Smale, Frank H. Wells and Arnold F. Wilkins.

Texas Instruments and IBM Agreement

Texas Instruments Incorporated announced recently that it had signed an agreement with International Business Machines Corporation under which both companies will work together in the area of transistors for data processing machines.

The agreement is expected to expedite progress in the area of interest of both companies through the exchange of technical information pertaining to transistors. Each company will protect the confidential nature of the information exchanged between them.

The agreement provides that IBM will purchase from TI a substantial portion of its expanding future commercial transistor needs. It places no restrictions on either company's present or future relations with other customers and suppliers.

Low Frequency Mechanical Oscillator

This miniature low frequency mechanical oscillator, manufactured by the Applied Science Corporation of Princeton is packaged in a hermetically sealed container measuring 2¼in square by 5in in length. The complete package weighs only 13oz. The output of the oscillator is a square pulse generated at 150c/s. The oscillator is essentially a sampling switch with a commutator-segment-type construction and two adjustable graphite brushes. This type of design allows the switch to handle currents up to 1A without adverse effects on switch life. The commutator is constructed with three make-before-break channels. The two brushes commutate the segments and pick up three pulses per revolution. The switch is designed so that the switch circuit is completed only when both brushes are on the same segment. This makes it possible to vary the width of the pulses from 10 to 80 per cent of total channel width by adjusting the distance between the brushes. This adjustment may be made by the operator while the oscillator is running. The model X-00006 is presently powered by a 9 000rev/min, 115V, single-phase, 400c/s drive motor; however, other motors are available.

Bell Telephone Developments

The following have been announced recently by the Bell Telephone Laboratories.

TRANSPARENT MAGNETIC OXIDES

A new class of magnetic oxides, structurally distinct from ferrites, has recently been discovered. These materials, known as rare-earth-iron garnets, are transparent, permitting the internal magnetic domain structure to be seen with a polarizing microscope.

The discovery of ferrimagnetism in these garnets was made first at the Laboratoire Electrostatique et De Physique du Metal of the Institute Fourier in Grenoble, France, and independently by S. Geller and M. A. Gilleo of Bell Telephone Laboratories.

Most magnetic materials, metals and ferrites alike, are opaque to visible light. Thus, their internal magnetic structure is not visible, and the way magnetic domains are oriented within them has been inferred from reflection of polarized light by their surfaces or by delineating domain boundaries at these surfaces with colloidal magnetic oxides.

Yttrium iron garnet, $Y_3Fe_2(FeO_4)_3$, is the most completely studied member of this new garnet family. It has a Curie temperature of 545°K and a spontaneous magnetization at zero temperature and infinite field of 4·96 Bohr magnetons per molecule, close to the theoretical value of 5·0. This magnetization results from super-exchange interactions through the O^{2-} ions between Fe^{3+} ions in crystallographically different positions in the cubic lattice. The number of these interactions per Fe^{3+} ion in yttrium-iron garnet is 3/5 that in a ferrite and, correspondingly, the observed Curie temperature of 545°K is 0·64 of that for magnetite (848°K).

Of great interest is the fact that yttrium-iron garnet contains magnetic ions with but a single valence. Both X-ray and neutron diffraction studies have shown clearly that, unlike the ferrites, the interactions between identical magnetic ions wholly occupying two different crystallographic sites are responsible for ferrimagnetism in the garnet structure.

In yttrium-iron garnet the width of the ferrimagnetic resonance absorption line at 9 300 and 2 4000Mc/s depends on crystallographic direction. Line widths of only 0·8 oersted were observed along the [100] direction in thin disks at 10 000 Mc/s, when the temperature was held at 540°K. These line widths increased with decreasing temperature and passed through peak values of several hundred oersteds between 65°K and 4·2°K.

A polarized light beam passing through the transparent magnetic domains in rare-earth-iron garnets has its plane of polarization rotated in one direction in one domain but in the opposite direction in an adjacent and oppositely magnetized domain. This Faraday rotation is wavelength-dependent and amounts to several degrees per mil of thickness; it makes the domains within the crystals clearly visible.

AUTOMATIC READING MACHINE

An experimental device the size of a portable typewriter that can read handwritten numerals or identify numerals as they are being written was announced recently.

The machine recognizes numbers as they are being written and indicates the numeral by lighting up the correct digit on a numbered panel. The writing is done with a metal stylus on a specially-prepared writing surface. Two dots, one above the other, are used as reference points. Seven sensitized lines extend radially from these two dots. Numerals are recognized by the machine, depending on which lines are crossed.

To clear the device for the next number, the writer touches the stylus to a special plate which causes the previous number to be 'erased.'

To recognize previously written numerals, such as those on a long distance telephone ticket, it is necessary to write with a pencil containing a conductive lead. The ticket is then inserted in the machine into a special slot under a plate that has seven sensitized lines. The machine then uses the same principle for recognizing numbers already written as it does for the ones written on the machine with the stylus. It determines which sensitized lines have been crossed and indicates the proper number by illuminating a numeral on the lucite panel. This information could be transferred to an accounting machine, computer or other data processing device.

The device is operated from flashpoint batteries and requires no outside power source. Its small size is made possible by the use of transistors.

Meetings this Month

THE BRITISH COMPUTER SOCIETY

Date: 17 February. Time: 6.15 p.m.
Held at: York Hall, Caxton Hall, London, S.W.1.
Lecture: Mental Arithmetic, an Historical Review with Demonstrations.
By: A. C. Aitken.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 5 February. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Radio Investigations during the International Geophysical Year.
By: W. J. G. Benyon.
Date: 26 February. (Time and place as above).
Lecture: A Long Range Navigational Aid.
By: C. Powell.

South Wales Section

Date: 26 February. Time: 6.30 p.m.
Held at: Glamorgan Technical College, Treforest.
Lecture: Industrial Television.
By: R. Swinden and J. E. H. Brace.

South Midlands Section

Date: 28 February. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: Some Advanced Applications of Information Theory.
By: P. H. Blundell.

West Midlands Section

Date: 12 February. Time: 7.15 p.m.
Held at: Wolverhampton Technical College, Wulfruna Street, Wolverhampton.
Lecture: Industrial Applications of Radio Isotopes.
By: R. F. Armitage.

North Western Section

Date: 6 February. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Manchester, 1.
Lecture: Electronic Musical Instruments.
By: Alan Douglas.

Merseyside Section

Date: 19 February. Time: 7 p.m.
Held at: Birkenhead Technical College.
Lecture: Closed-Circuit Television.
By: J. G. M. Downes.

Scottish Section

Date: 20 February. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, Elmbank Crescent, Glasgow.
Lecture: Frequency Response Instrumentation for the Analysis, Design and Production of Servomechanisms.
By: G. R. Swainston.
Date: 21 February. Time: 7 p.m.
Held at: Department of Natural Philosophy, The University, Edinburgh.
(Repeat of above lecture.)

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meeting

Date: 6 February.
The Forty-Ninth Kelvin Lecture: High-Polymers.
By: H. W. Melville.

Faraday Lecture

Date: 12 February. Time: 6 p.m.
Held at: Central Hall, Westminster, London, S.W.1.
Lecture: The Electrification of the British Railways.
Prepared by the late G. H. Fletcher.
Delivered by: R. Ledger.
(Admission by ticket available on receipt of a stamped addressed envelope.)

Measurement and Control Section

Date: 4 February.
Lectures: Temperature Transients in Gas-Cooled Thermal Nuclear Reactors.
By: J. H. Bowen and E. F. O. Masters.
Boron-Trifluoride Proportional Counters and The Design, Performance and use of Fission Counters.
By: W. Abson, P. G. Salmon and S. Pyrah.
The Application of Digital Computers to Nuclear-Reactor Design.
By: J. Howlett.
Date: 18 February.
Lecture: Magnetic Tape for Data Recording.
By: C. D. Mee.

Radio and Telecommunication Section

Date: 19 February.
Lecture: The Relation Between Picture Size, Viewing Distance and Picture Quality with special reference to Colour Television and to Spot-Wobble Techniques.
By: L. C. Jesty.
Date: 24 February.
Informal evening on Stereophonic Recording on Gramophone Discs.
Talk by: H. A. M. Clark.

East Midland Centre

Date: 4 February. Time: 6.30 p.m.
Held at: E.M.E.B. Service Centre, Derby.
Lecture: Automatic Circuit Reclosers.
By: G. F. Peirson, A. H. Pollard and N. Care.
Date: 13 February. Time: 7.30 p.m.
Held at: The George Hotel, Kettering.
Lecture: 275kV Transmission Lines.
By: J. D. Pierce.
Date: 20 February. Time: 7.15 p.m.
Held at: The Albert Hall, Nottingham.
Faraday Lecture: The Electrification of the British Railways.
Prepared by the late G. H. Fletcher.
Delivered by: R. Ledger.

Cambridge Radio and Telecommunication Group

Date: 4: February. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Informal evening on Problems of Sound and Television Broadcasting Coverage.
Talk by: G. Millington.
Date: 25 February. Time: 8 p.m.
Held at: The Physiological Laboratory, Cambridge.
Lecture: The Eye as a Television Camera.
By: F. W. Campbell.

Mersey and North Wales Centre

Date: 3 February. Time: 6.30 p.m.
Held at: The Town Hall, Chester.
Informal evening on Domestic High-Fidelity Reproduction.
Talk by: J. Moir.
Date: 17 February. Time: 6.30 p.m.
Held at: The Royal Institution, Colquitt Street, Liverpool.
Discussion of the Report on The Supply and Training of Teachers for Technical Colleges.
Introduced by: Willis Jackson.

North-Eastern Radio and Measurements Group

Date: 3 February. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Informal Lecture: Domestic High-Fidelity Reproduction.
By: J. Moir.
Date: 17 February. (Time and place as above).
Informal Lecture. The Measurement of High Voltages with Indicating or Recording Instruments.
By: G. W. Bowdler.

Tees-Side Sub-Centre

Date: 5 February. Time: 6.30 p.m.
Held at: Cleveland Scientific and Technical Institute, Corporation Road, Middlesbrough.
North-Eastern Centre Chairman's Address.
By: T. W. Wilcox.

North Midland Centre

Date: 27 February. Time: 6.30 p.m.
Held at: The Yorkshire Electricity Board, Ferensway, Hull.
Lecture: British Columbia-Vancouver Island 138kV Submarine Power Cable.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

South-East Scotland Sub-Centre

Date: 4 February. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: Electric Control of Stage and Television Lighting.
By: F. P. Bentham.
Date: 12 February. (Time and place as above).
Lecture: The Relation Between Picture Size, Viewing Distance and Picture Quality, with special reference to Colour Television and to Spot-Wobble Techniques.
By: L. C. Jesty.
Date: 18 February. (Time and place as above).
Lecture: British Columbia-Vancouver Island 138kV Submarine Power Cable.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

South-West Scotland Sub-Centre

Date: 5 February. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: Electric Control of Stage and Television Lighting.
By: F. P. Bentham.
Date: 11 February. Time: 7 p.m.
Held at: The Royal College of Science and Technology, George Street, Glasgow, C.1.
Lecture: The Relation between Picture Size, Viewing Distance and Picture Quality, with special reference to Colour Television and to Spot-Wobble Techniques.
By: L. C. Jesty.

South Midland Radio and Measurements Group

Date: 24 February. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Radio Research and the International Geophysical Year.
By: R. L. Smith-Rose.

South Midland Supply and Utilization Group

Date: 10 February. Time: 6 p.m.
Held at: The College of Technology, Birmingham.
Lecture: British Columbia-Vancouver Island 138kV Submarine Power Cable.
By: T. Ingledow, R. M. Fairfield, E. L. Davey, K. S. Brazier and J. N. Gibson.

North Staffordshire Sub-Centre

Date: 18 February. Time: 7.30 p.m.
Held at: The Victoria Hall, Hanley.
Faraday Lecture: The Electrification of the British Railways.
Prepared by the late G. H. Fletcher.
Delivered by: R. Ledger.

Rugby Sub-Centre

Date: 11 February. Time: 6.30 p.m.
Held at: Rugby Technical College.
Lectures: The Development of Variable-Speed High-Power Drives for Large Wind Tunnels, and A Variable-Frequency Power Installation for Large Wind-Tunnel Drives.
By: P. McKearney, L. S. Drake and E. G. Mallalieu.
Date: 25 February. (Time and place as above).
Lecture: The Importance of Research in Hearing and Seeing to the Future of Telecommunication Engineering.
By: E. C. Cherry.

Southern Centre

Date: 10 February. Time: 2.30 p.m. and 6.30 p.m.
Held at: The Guildhall, Southampton.
Faraday Lecture: The Electrification of the British Railways.
Prepared by the late G. H. Fletcher.
Delivered by: R. Ledger.
Date: 26 February. Time: 6.30 p.m.
Held at: The C.E.G.B. Offices, Portsmouth.
Lecture: Route Selection and Cable Laying for the Trans-Atlantic Cable System.
By: R. J. Halsey.

Hatfield District

Date: 7 February. Time: 7 p.m.
Held at: Hatfield Technical College.
Lecture: Colour Television.
By: P. C. Kidd.

THE TELEVISION SOCIETY

Date: 7 February. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: The Effects of Noise in Television Transmission.
By: T. Kilvington.
Date: 20 February. (Time and place as above).
Lecture: Test Equipment for Colour Television Receivers.
By: F. H. Cohen and D. C. Kidd.