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Commentary

THE publication of a personal account* of the history and development of radar by Sir Robert Watson-Watt has been long overdue and its appearance, late though it may be, has brought back such vivid memories and stirred such thoughts that it would be wrong to allow the occasion to pass without comment. It contains a number of lessons which should not be forgotten.

The years pass quickly and it is difficult to realize that the young men who flew the Hurricanes and Spitfires in the Battle of Britain are now approaching their forties. Those not old enough to share their vivid recollections can have little idea of the narrow margin by which Fighter Command, facing odds of more than three to one, succeeded in defeating the Luftwaffe. There was at least one occasion during the battle when not a single fighter remained in reserve upon the ground yet, thanks to radar and the early warning it gave of an approaching raid, we were always able to control and direct our few fighter squadrons to intercept and break up the enemy formations. Without radar we could not have won the Battle of Britain, Britain would have suffered invasion and, unprepared as we were, we should almost certainly have been defeated.

Such, then, is the debt which we owe to Sir Robert Watson-Watt, to A. F. Wilkins, E. G. Bowen, L. H. Bainbridge-Bell and a handful of others for their foresight, their brilliant ingenuity and the efforts which initiated an entirely new conception of air defence.

Watson-Watt was not the first to observe radio echoes, he was not the first to use pulse techniques, but he was unquestionably the first to propose and apply these principles in a co-ordinated scheme of early warning and fighter control. Granted that the initial conception was a brilliant one, granted its unique timeliness in that there was no other practical solution in sight to the problem of adequate defence against the impending threat of the mid-thirties, we must yet be ever thankful that we had a man with the vision, personality and drive of Watson-Watt, not only to direct the technical development of his scheme but also, by battling continuously against the petty obstructions and frustrations which seem so invariably to permeate all Government departments, to succeed in building around our coasts a chain of early-warning stations which were ready, just in time, to face the German onslaught of 1940.

It is never enough merely to invent something; it is the application that really matters and unless the inventor possesses those personal gifts which can overcome all the obstacles which beset the road towards achievement, his invention is likely to wither in the desert air. The greater the invention, the more numerous will be the difficulties in its development and application and not the least of these is likely to be the lack of appreciation and imagination which so often characterizes the administrative channels of Government departments.

By 1936 it had become evident to the 'man in the street' that war was a possibility; to those with inside knowledge it seemed virtually inevitable, but even with this certain threat the brilliant teamwork among the little band of radar pioneers would have been of little avail without the energy and irresistible tenacity with which Watson-Watt pursued his chosen task. Never a shy man, never a 'respecter of persons', he had early learned that a "worthy end may sometimes be better served by an unofficial and informal short-cut than through official channels". Somehow he found the way to break through the restraints of departmental administration; somehow he found the means to augment his little team with brilliant recruits from the universities; somehow, in those few short years, the whole great scheme was planned, the stations built, personnel trained and a revolution evolved in the operational tactics of Fighter Command.

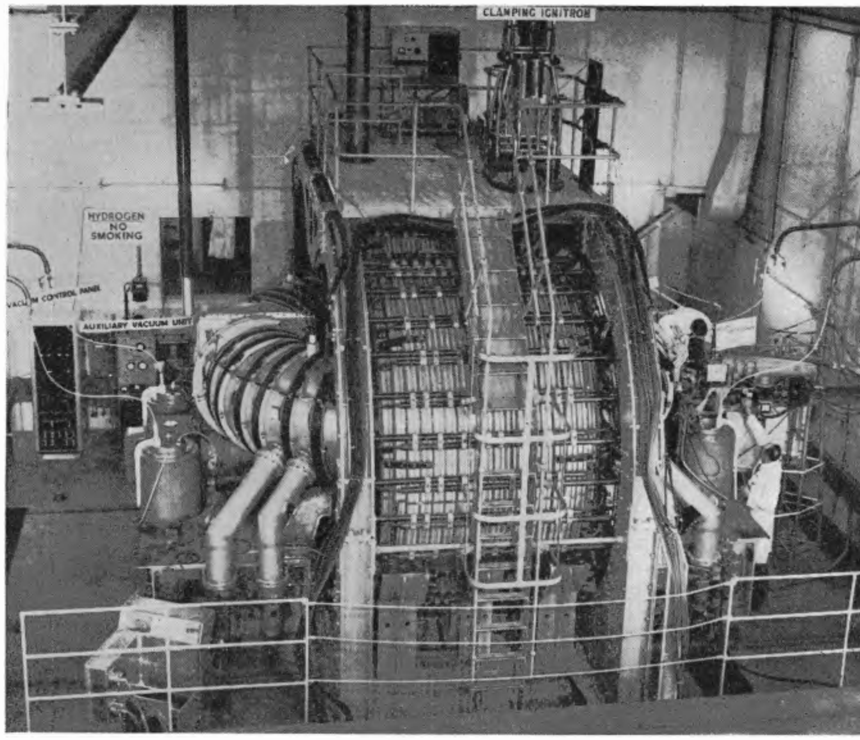
Speaking later of the young pilots upon whom the burden fell, Sir Winston Churchill was to make his famous declaration that "Never in the field of human conflict was so much owed by so many to so few". Had it not been, however, for the chain of radar stations around the coasts, their courage and their sacrifice would have been in vain; few as their numbers were, they only just sufficed; without the aid of radar they must have failed.

Veiled in secrecy as was radar throughout the war, it is certain that the general public has never fully appreciated the debt we owe to Sir Robert Watson-Watt. To him more than any other single individual is due the credit for the winning of the Battle of Britain and, when one recalls also the tremendous part which radar played in the Battle of the Atlantic and in the final Battle for Europe, the magnitude of our debt surpasses estimation.

Let it be hoped that such a need will never arise again. If it be true, however, that it was the proper application of science which won the war, let us all realize that it is only the proper application of scientific achievements which will enable us to remain on the winning side in the even greater economic battle of modern life.

* WATSON-WATT, R. Three Steps to Victory (Odhams, 1958). (*Reviewed on p. 158 of this issue*).

ZETA
the Zero Energy Thermonuclear Assembly



A General View of ZETA

RESEARCH at Harwell has led British scientists to the conclusion that control of thermonuclear reactions for electricity generation may well be a possibility for the future, though its practical application is still a long way off.

Results obtained suggest that 'thermonuclear neutrons' have been obtained, but further experiments will be necessary before this can be proved conclusively.

On 12 August, 1957, a large experimental apparatus known as ZETA (Zero Energy Thermonuclear Assembly) for studying the controlled release of energy from thermonuclear reactions was started up at the Atomic Energy Research Establishment, Harwell. On 30 August this apparatus was first operated under conditions that produced nuclear reactions; neutrons emitted in these reactions were observed when deuterium gas was heated electrically to temperatures in the region from 2 to 5 $\times$ 10^6°C. The hot gas was isolated from the walls for periods of 2 to 5msec. The heating process was repeated every 10sec. The high temperatures achieved, together with the relatively long duration for which the hot gas has been isolated from the tube walls, are the most important experimental results obtained so far.

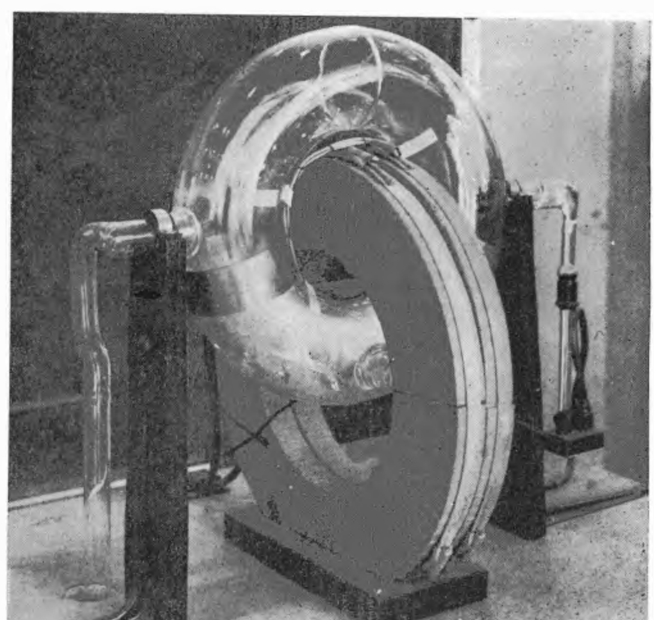
The source of the observed neutrons has not yet been definitely established. There are good reasons to think that they come from thermonuclear reactions, but they could also come from other reactions such as collision of deuterons with the walls of the vessel, or from bombardment of stationary ions by deuterons accelerated by internal electric fields produced in some forms of unstable discharge.

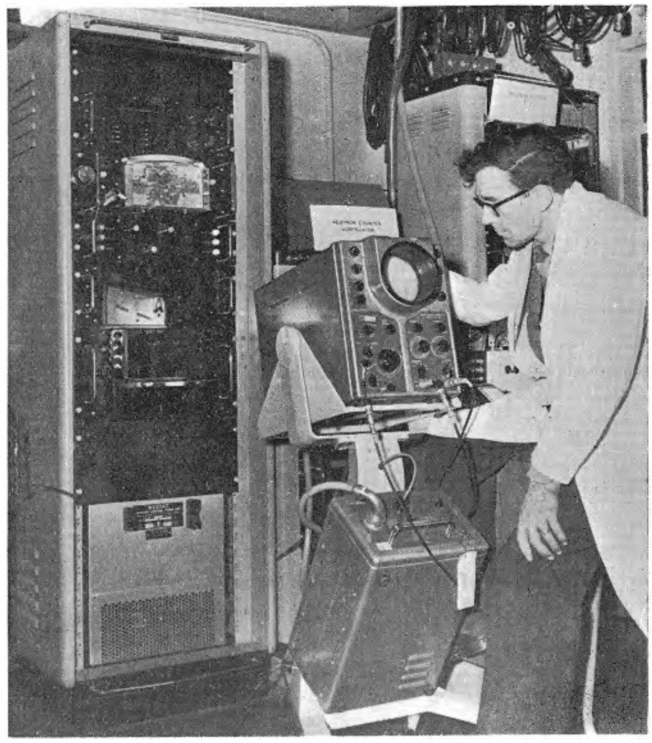
In the ZETA apparatus the number of neuterons produced by each pulse of energy as the current was doubled was roughly that which might be expected from a thermonuclear reaction at the measured temperatures. These temperatures have been definitely established.

Operation of Zeta

The principle adopted is to pass a large electric current through the deuterium gas. This circuit sets up an electric discharge in the gas which heats it and also produces an intense magnetic field around the column of hot gas. This magnetic field causes the discharge to become constricted and hence heated still more. Since it also causes the discharge to wriggle about, this field by itself is not enough to

A small discharge circuit using a glass toroid. The metal bands round the glass toroid are connected to an r.f. source and are used for pre-ionizing the gas





A section of the Neutron Counting Laboratory

keep the discharge away from the walls. The wriggling has been suppressed by applying an additional steady magnetic field parallel to the axis of the tube.

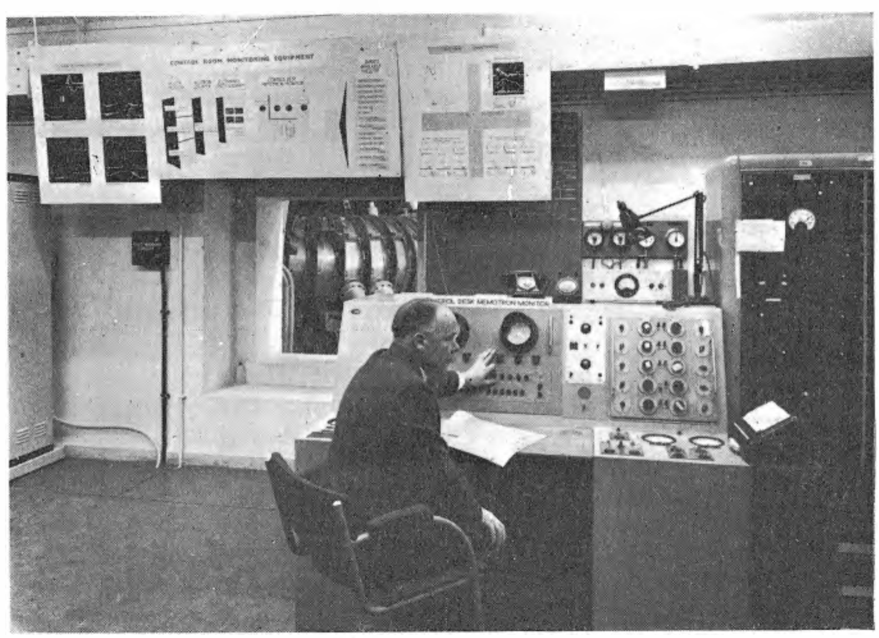
In Zeta the discharge chamber is a ring-shaped tube or torus of 1 metre bore and 3 metres mean diameter, containing deuterium gas at low pressure. The tube is linked by the iron core of a large pulse transformer. A current pulse of electricity is passed into the primary winding of the transformer from a bank of capacitors capable of storing 500 000 joules of energy. The bank of 1 600μF is charged to a potential of 25kV from a conventional metal rectifier equipment through a saturable reactor. Charging and discharging is carried out by a combination of ignitrons and mechanical switches, the design for the latter being based on high voltage airblast current breakers. The discharge pulse induces a very large unidirectional pulse of current in the gas, which forms a short-circuited secondary for the transformer. Peak currents up to 200 000A have been passed through the ionized gas for periods up to 5msec. The current pulse is repeated every ten seconds. Emission of neutrons throughout the current pulse is observed regularly in routine operation of ZETA with deuterium; there are up to 3 million neutrons emitted per pulse.

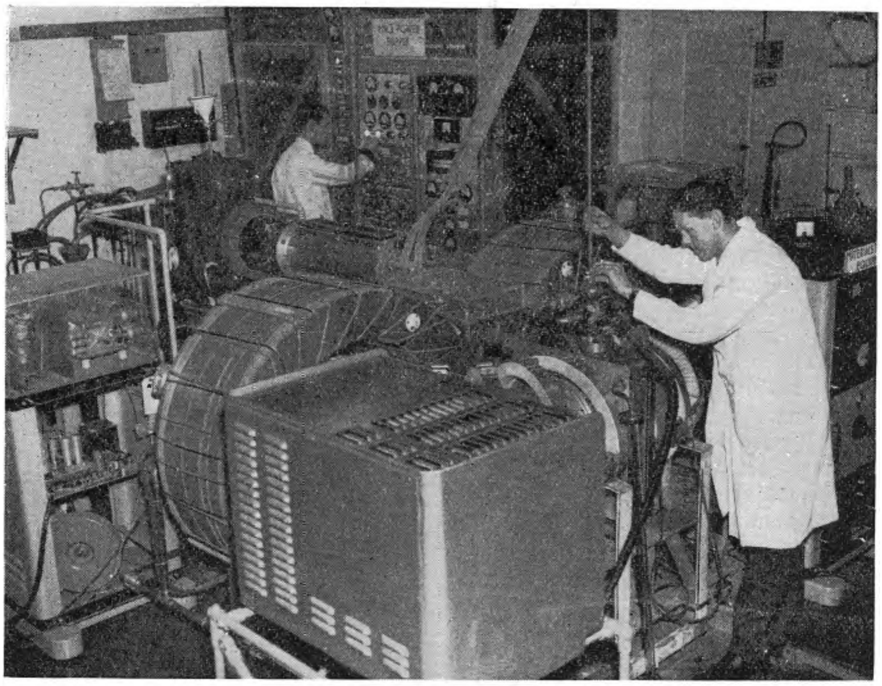
The temperature of gas discharges may be determined from measurements on the light emitted by the gas atoms but measurements of this kind in these experiments present problems because, at the temperature of the discharge, the hot deuterium atoms are completely stripped of their electrons and therefore do not emit a line spectrum. One method of solving this problem is to mix with the deuterium a small quantity of some heavier gas, such as oxygen or nitrogen, the atoms of which are not stripped of all their electrons under these conditions, and to study the spectral lines emitted by this impurity; the random motion of the high-energy impurity atoms which make many collisions with the deuterium atoms and so reach the same energy causes the spectral lines to broaden, owing to the Doppler effect, and the amount of broadening is a measure of the ion energy. Many measurements by this method have indicated temperatures in the region of 2 to 5 $\times$ 10^6°C. While temperatures in this range are required to explain the observed rate of neutron production on the basis of a thermonuclear process, electric fields in the gas arising from instabilities, can also accelerate deuterium ions and lead to nuclear reactions. Such a process was described by Academician Kurchatov of the U.S.S.R. in his lecture at Harwell in 1956. Therefore it is not altogether certain that the observed neutrons come from a thermonuclear reaction. Experiments are continuing to study the details of the neutron producing processes.

In order to obtain a net gain in energy from the reaction it would be necessary to heat deuterium gas to temperatures in the region of 100 $\times$ 10^6°C, and to maintain it at this temperature long enough for the nuclear energy released to exceed the energy needed to heat the fuel and lost by radiation. Lower temperatures would suffice for a deuterium-tritium mixture. The high temperatures achieved in ZETA, and the relatively long duration for which the hot gas has been isolated from the tube walls are the most important experimental results obtained so far. While a much longer time (perhaps several seconds) is required for a useful power output there appears to be no fundamental reason why these longer times, together with much higher temperatures, cannot be achieved. There are, however, many major problems still to be solved before its practical application can be seriously considered and the work must be expected to remain in the research stage for many years yet.

The work on ZETA has been done in the General Physics Division at Harwell which is under the direction of Mr. D. W. Fry. The group responsible for the work has been led by Dr. P. C. Thonemann and senior members concerned

The control desk of Zeta. Part of the toroid of Zeta can be seen through the window





The MK 3 ZETA, a prototype apparatus one third scale, the design of ZETA was completed with its aid. The torus is now used for materials testing work. Currents up to 800 000A are used here

with Zeta have been Mr. R. Carruthers and Mr. R. S. Pease, with Mr. J. T. D. Mitchell of the Engineering Division, and Dr. W. B. Thompson of the Theoretical Physics Division.

Full collaboration in the controlled thermonuclear reaction field of research was established with the U.S.A.E.C. in October, 1956.

Research work in the field of controlled thermonuclear reactions is also being carried out at The A.E.I. Research Laboratory (Director, Dr. T. E. Allibone) on behalf of the Atomic Energy Authority and with the advice of Sir George Thomson. Experimental work was started at Imperial College by Sir George Thomson in 1947. It was moved to the A.E.I. Laboratory in 1951. The senior members of staff engaged are Mr. D. R. Chick and Dr. A. A. Ware.

A relatively small apparatus known as SCEPTRE 3 was used and consists of an aluminium torus or ring made of 12in diameter tubing, the mean torus or ring diameter being 45in, and is threaded by an iron core weighing 4 tons. Very many experiments have been performed with this apparatus, discharging into it energies up to 40 000 joules. This energy has caused currents up to 200 000A to flow in the gas and temperatures of nearly 4×10^6°C have been measured. At temperatures of over $2{\cdot}5 \times 10^6$°C neutrons were emitted copiously over a time of several hundred microseconds. Much work has been done to establish the conditions under which the neutrons are emitted, the manner in which the number emitted varies with gas pressure, gas temperatures, and with different values of the stabilizing magnetic field.

The work has been supported for the most part by a contract from the Atomic Energy Authority, the balance being financed by the Associated Electrical Industries.

ZETA
(The Control Room Monitoring and Recording Instruments)

By E. P. Butt*, B.Sc., A.M.I.E.E.

ZETA is a large experimental device and it is, therefore, necessary to make a large number of measurements and recordings of them. In this article the necessary measurements are detailed and the control room monitoring and recording instruments are described.

ZETA, the Zero Energy Thermonuclear Assembly at Harwell, is a large experimental device designed to investigate the thermonuclear fusion of deuterium. A description of the machine, together with some preliminary results has recently been published[1,2].

ZETA is, essentially a large pulse transformer for producing high currents in a low pressure gas. The primary winding is connected to a large bank of charged capacitors every 12sec. An electrical discharge in the gas contained in the torus, usually deuterium, takes the place of the secondary winding. Currents of up to 2×10^5A flow in the discharge channel heating the gas to temperatures which may reach a maximum of over 5×10^6°K. At this temperature, in deuterium, neutrons are produced at the rate of about 2×10^6 per pulse. This discharge has been repeated at 12sec intervals† approximately 100 000 times.

* A.E.R.E., Harwell

† *10sec intervals were used in the earlier experiments*

The automatic cycle of operations and the control of gas pressure, high tension voltage etc. are centralized in the control room which also acts as the focus for experimental observations. The control room monitoring and recording instruments are used for recording information derived from a number of measuring techniques used simultaneously at every pulse. By this means the maximum information is derived from a particular experiment.

The Controls of Zeta

The cyclic operation of Zeta is an important feature of the machine. Several hundred discharges are usually required before steady conditions are reached if the machine has been open to the atmosphere for an overhaul. Even after a brief stop, one or two discharges are necessary before normal conditions are re-established. This "running in" time is believed to be connected with impurities on the walls of the torus. These impurities are gradually removed by the discharge. Consistent results are not obtained from occasional single discharges.

The cycle of operations required to produce each discharge is shown diagrammatically in Fig. 1 and the circuit of ZETA is given in Fig. 2. These operations are performed automatically by a continuously rotating timing drum. Relay circuits ensure that the sequence is started in the correct order, irrespective of the position of the drum at the time of starting. There is also an interlock system which prevents a cycle of operations being commenced if there is a fault in any part of the equipment.

The sequence of operations, as shown in the figure, is mainly self explanatory, but it will be noted that the axial magnetic field can either be continuous or pulsed. If the

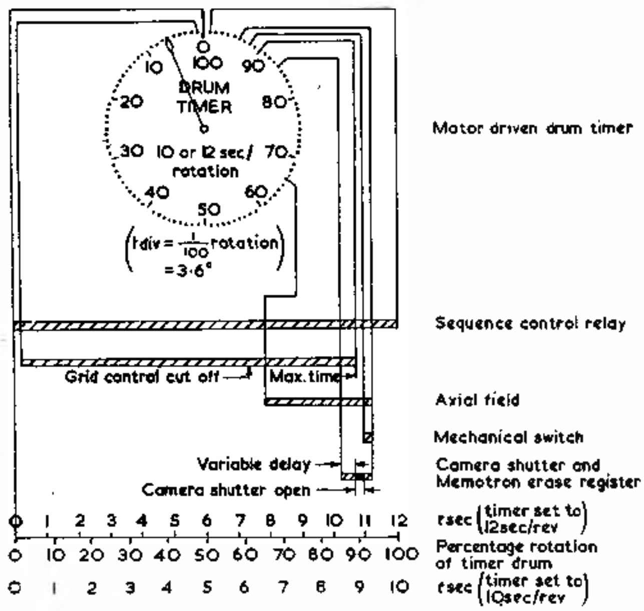


Fig. 1. Timing sequence for ZETA

current required in the coils exceeds their rating for continuous operation, then the field of the motor generator set is pulsed. It will also be noted that a pulse occurs, shortly before the discharge, which opens the shutter of the recording camera of the oscillograph. This pulse also operates counters which record the number of the discharge. Counters are located in the control desk and in each of the five laboratories which surround the machine. The discharge number and date is also recorded automatically by the oscillograph so that each frame can be identified and correlated with other records.

(a) Gas Controls

An ionization gauge measures the pressure of the gas in the body of the torus. This pressure may be controlled in two ways. Firstly by varying the pumping speed of the system by controlling the fine side valves. This provides a coarse control of pressure. Secondly by altering the rate at which gas is let into the torus. In the case of hydrogen or deuterium, the input of gas is varied by altering the temperature of a thin walled nickel tubing through which the hydrogen diffuses at a rate increasing with temperature. The pressure of other gases is controlled by a motor operated needle valve.

(b) Capacitor Voltage

This is controlled by adjusting the input voltage to the e.h.t. rectifier set which is connected to the capacitor bank by means of a large transmitting triode. This valve is used as a constant current regulator and also isolates the capacitors from the supply before the actual discharge.

(c) Stabilizing Axial Field Supply

The current through the axial field windings is controlled by remotely adjusting the field supply to the motor generators.

In addition to these controls and the monitoring and recording instruments which will be discussed in a later section, an intercommunication system links the control room with the other laboratories in the plant area.

It will be seen from the circuit diagram Fig. 2, that the capacitor bank is connected to the primary by means of a mechanical switch. The time taken for this switch to operate is uncertain to within a few milliseconds, so that it is not possible to derive signals from the main timing circuits which are required to trigger oscillographs and

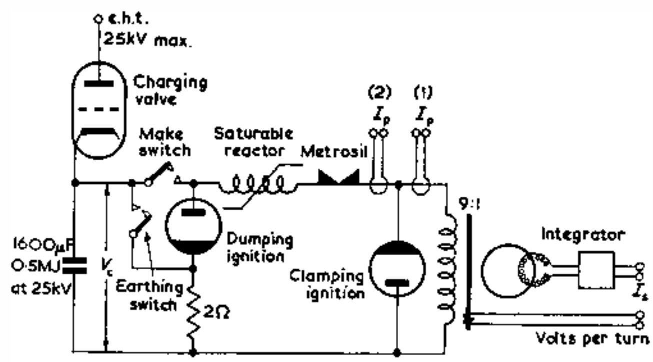


Fig. 2. ZETA main circuit diagram showing the principal electrical measurements

other electronic equipment before the event. Such signals are obtained from the saturable reactor in series with the switch. This provides a roughly square waveform of from 500μsec to 1 500μsec duration depending on the capacitor voltage. This signal is applied to a ten channel electronic timing unit* which provides the two trigger pulses required for the eight channel oscillograph and eight other trigger pulses for general use in the laboratories. The ten trigger pulses supplied by the electronic timer unit may be delayed up to 10msec from the front or rear edge of the master pulse from the saturable reactor.

The Recording and Monitoring Instruments

A schematic diagram of the recording and monitoring instruments is shown in Fig. 3. The main recording instrument is an eight channel oscillograph† with which signals from a particular experiment are recorded on 35mm film. This information is not immediately available as the film must be processed. It is therefore necessary to provide monitoring facilities on which the required information is instantly presented. A two channel storage type monitor enables any two channels, of the eight being recorded, to be displayed. By this means the experiment can be controlled and the satisfactory working of the machine ensured. This information is reinforced by that provided from a neutron detector and fast scaler unit, and a health monitor consisting of a rack containing three rate meters for fast neutrons, slow neutrons and gamma rays respectively.

The choice of an eight channel recording oscillograph was determined by practical considerations. Four square

* *Supplied by Metropolitan-Vickers Electrical Co. Ltd.*

† *Supplied by A. E. Cawkell (Electronic Engineers), Southall*

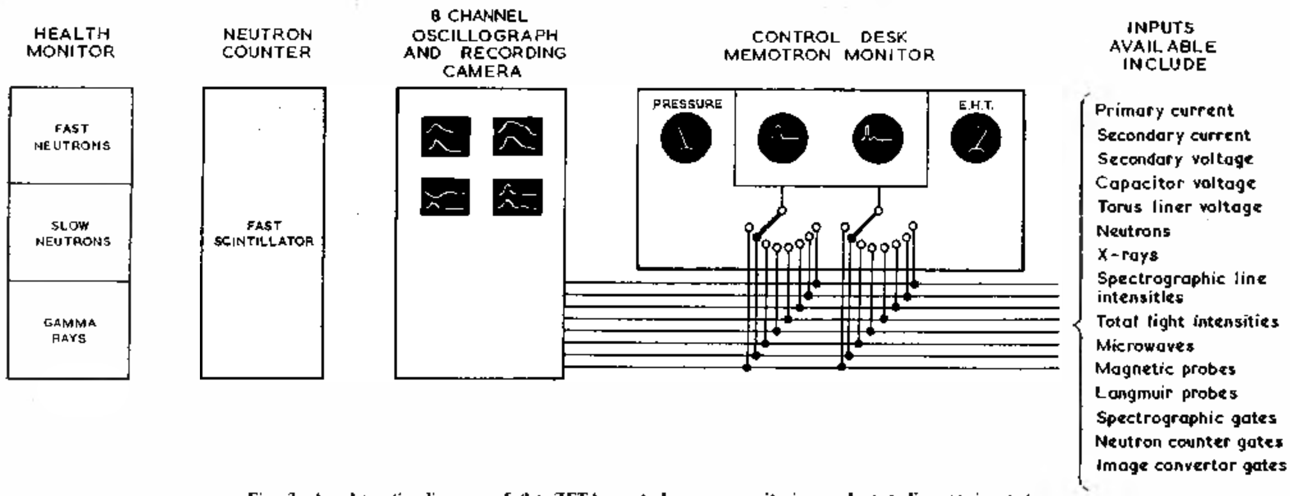


Fig. 3. A schematic diagram of the ZETA control room monitoring and recording equipment

5in cathode-ray tubes, each having two electrode assemblies, form a display which is 12in square. This can be readily photographed on 35mm film* and enlarged to nearly full size, for measurement purposes, by a commercial projector giving an accurate ×10 enlargement†. This arrangement represents the optimum compromise, for the present purposes, between the number of channels and the recorded detail. Photographic grain size imposes a limit on the recorded information, and to increase the resolution available it is proposed to use 70mm film when the occasion demands, if a suitable camera can be developed.

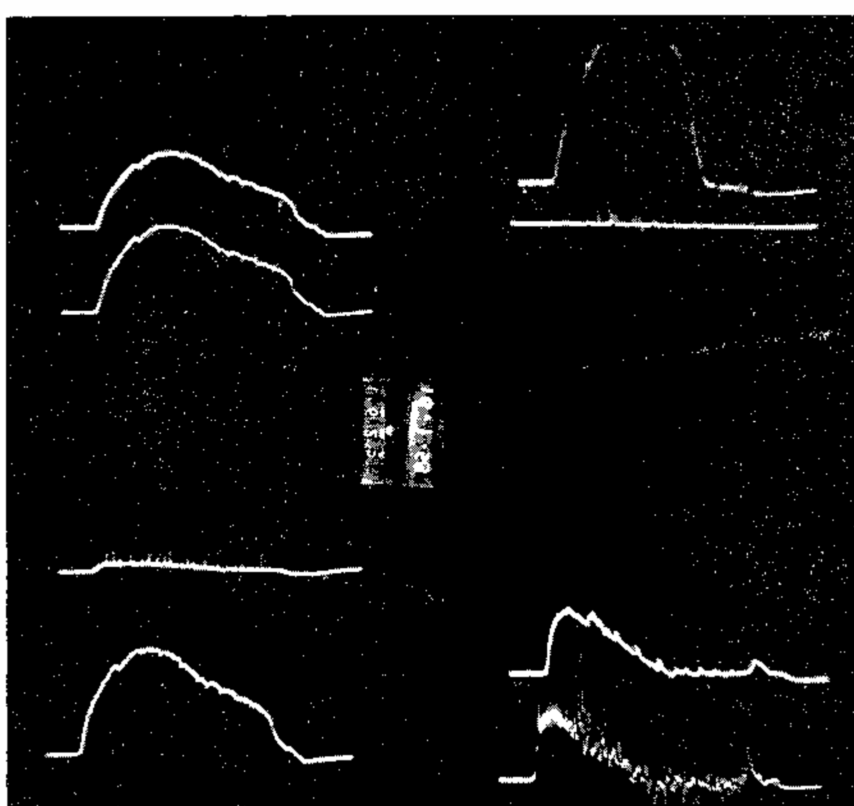
It is not uncommon for the present equipment to be in use for periods as long as twelve hours, continuously recording the eight traces every 12sec. It was therefore necessary to demand a high standard of stability and reliability from this instrument. All the main power supplies are stabilized including the high tension supply to the cathode-ray tubes. This latter point is important as no further adjustments are needed to the 'spot brilliance' controls once they have been initially set up at the beginning of the day.

Over forty inputs from the other laboratories together with a number of electrical measurements are available and the eight signals to be recorded are selected by means of a 56-way patching panel at the rear of the control desk.

The 'Memotron' monitor is a slave unit to the eight channel oscillograph, which enables any two of the signals being recorded to be observed on the control desk. It consists of two oscilloscopes with identical sets of X and Y amplifiers to the recording oscillograph. These are supplied with deflexion and beam brightening waveforms from the main oscillograph. The cathode-ray tubes employed, however, are storage tubes which have certain special advantages for the present purposes. The 'Memotron' is an American cathode-ray tube which stores indefinitely the trace on the screen. This is achieved by writing the transient waveform initially on to a storage mesh which is flooded with low velocity electrons. These electrons are allowed to pass through the mesh only at the points which have been traversed by the writing beam, and after suitable acceleration, excite the phosphor to give a picture of the waveform. The stored pattern can be instantly erased and there are facilities for automatic erasure before each pulse or local erasure by a push-button control. By this means any two signals may be observed at comparative leisure in the 12sec between each pulse. Alternatively, if the waveform is non-repetitive, a number of traces may be superimposed. Families of traces from the same channel may be stored to observe the effect of varying a particular parameter

Fig. 4. Typical eight channel records
Shot No. 0625, 16 January, 1958

Chan. 1. Is (1)	*Chan. 5. Ip (2)*
Chan. 2. Ip (1)	*Chan. 6. X-rays*
Chan. 3. Neutrons	*Chan. 7. V/In (filtered)*
Chan. 4. Is (2)	*Chan. 8. V/In*



* *Camera supplied by D. Shackman & Sons*
† *Supplied by Watson Monasty & Co. Ltd.*

or groups of traces collected from a number of different channels on consecutive discharges. These facilities have been of great value, but storage tubes also have their limitations. The amount of information which can be stored is limited by the mesh aperture size which is of the order of 112 to the inch. The writing speed is also limited. The storage mesh is in effect a two state device, and in consequence does not store information below a certain level of charge. Those parts of a waveform which involve the writing beam traversing the mesh at a speed in excess of that required to raise the potential to the minimum level are therefore lost. Its maximum value is of the order of 1cm/μsec, but varies greatly between tubes. The contrast ratio between the background illumination and the trace is of the order of 3 : 1, which is too low to photograph satisfactorily.

The stability requirements for the Memotron amplifiers and the e.h.t. supplies are similar to those required for the eight channel oscillograph.

A detailed description of both the eight channel oscillograph and the Memotron monitor will be found elsewhere in this journal[3].

The input signals indicated in Fig. 3 represent a wide range of measuring techniques. These can be roughly divided into three classes. Firstly, electrical circuit measurements, secondly the measurement of various forms of radiation from the discharge, and thirdly measurements from various form of probe. Each of these groups of measurements is briefly described in a subsequent section of this article in order to illustrate the use of the control room recording and monitoring equipment

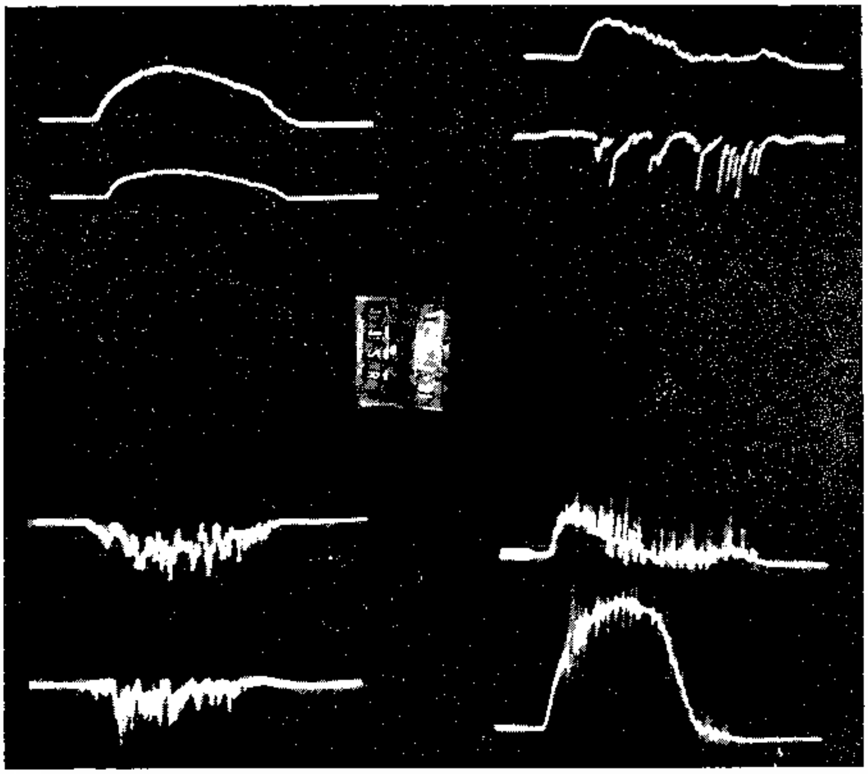


Fig. 5. Shot No. 1128, 18 December, 1957

Chan. 1. Is
Chan. 2. Ip (1)
Chan. 3. Nitrogen 4 3479A
Chan. 4. Total light
Chan. 5. V/tn (filtered)
Chan. 6. X-rays
Chan. 7. V/tn
Chan. 8. Ip (2)

Electrical Circuit Measurements

The main circuit of the machine is shown in Fig. 2. A mechanical switch connects the capacitor bank to the primary. A saturable reactor protects the switch by limiting the initial current. A bank of non-linear resistors protects the capacitors from a short-circuit across the primary or an open-circuit in the secondary. The secondary load is predominantly inductive so that it is necessary to short-circuit the primary, by means of a clamping ignitron, when the primary voltage reaches zero. This prevents the reversal of the voltage on the capacitors and prolongs the discharge. The capacitors are rated at a maximum of 5 per cent reverse voltage at full e.h.t. The principal electrical measurements obtained are shown in the circuit diagram and are as follows:

(1) The primary current I_p (1) is measured by a current transformer. This measurement includes the current which circulates after the clamping ignitron has fired.

(2) The current external to the clamping circuit I_p (2) is measured by another current transformer. This current falls rapidly to zero at zero volts.

(3) The secondary current I_s is measured by means of a Rogowski coil encircling the torus body. The voltage output from this coil is proportional to dI/dt and hence when integrated by a suitable valve circuit gives a measurement of the current enclosed by the loop. A similar coil has also been inserted into the torus to measure only the current in the gas discharge. Practical difficulties such as overheating and arcing prevent this being a permanent installation at the moment.

(4) The electric field applied to the gas discharge is obtained by measuring the voltage induced in a single turn round the core. This is known as the volts/turn signal. It is convenient on occasions to record this signal without the high frequency components. For this purpose a low-pass filter is inserted in the circuit.

(5) The voltage across the capacitors, V_c, is also measured and recorded.

These electrical circuit measurements can yield a great deal of information about the nature of the discharge. The difficulty lies mainly in interpreting the results, as it must be remembered that, while the current is changing, both the inductance and resistance of the discharge are also varying with time. Some typical traces can be seen in Figs. 4, 5, and 6.

Certain operational signals derived from the circuit are also recorded when necessary. An example of this type of signal is the trigger pulse for the clamping ignitron which is derived from the volts/turn waveform.

Radiation Measurements

A second main classification of input signal, available in the control room, is that obtained from apparatus which measures the intensity of the various forms of radiation emitted by the discharge.

Neutrons are detected by a plastic (proton recoil) scintillator and counted by a fast, gated, scaling unit in the control room. This acts as a very good guide to conditions within the discharge but does not provide a signal suitable for recording. Experiments are undertaken which involve special detectors and time and energy distribution measurements. For these experiments both scintillators and boron trifluoride detectors are used in conjunction with amplifiers which give a recordable signal.

X-rays are also detected by means of crystal scintillators and the output signals are recorded when required. Examples of both X-ray and neutron detector outputs are shown in Figs. 4 and 5.

There is very little visible light emitted by the discharge in deuterium after the first hundred or so microseconds, as the gas is then fully ionized. It has been found possible to estimate the temperature of the ions, however, by measuring the Doppler broadening of emission lines from highly ionized impurities. These may be present due to unavoidable contamination in the vacuum system but can be introduced deliberately when necessary. In order to know which lines to use, the intensity variations during the pulse are observed. Spectral lines which have a maximum intensity when the current is high are most suitable. When the gas concerned has not been introduced deliberately it is also desirable to know if the supply is maintained. For example, after continuously running for a long time all oxygen lines will disappear. The outputs from special photomultipliers used to measure the intensity of a particular spectral line are therefore recorded. The total light output is also recorded when required. Examples can be seen in Fig. 5.

The microwave noise emitted by the discharge is also being investigated as a method of measuring the electron temperature of the plasma[4].

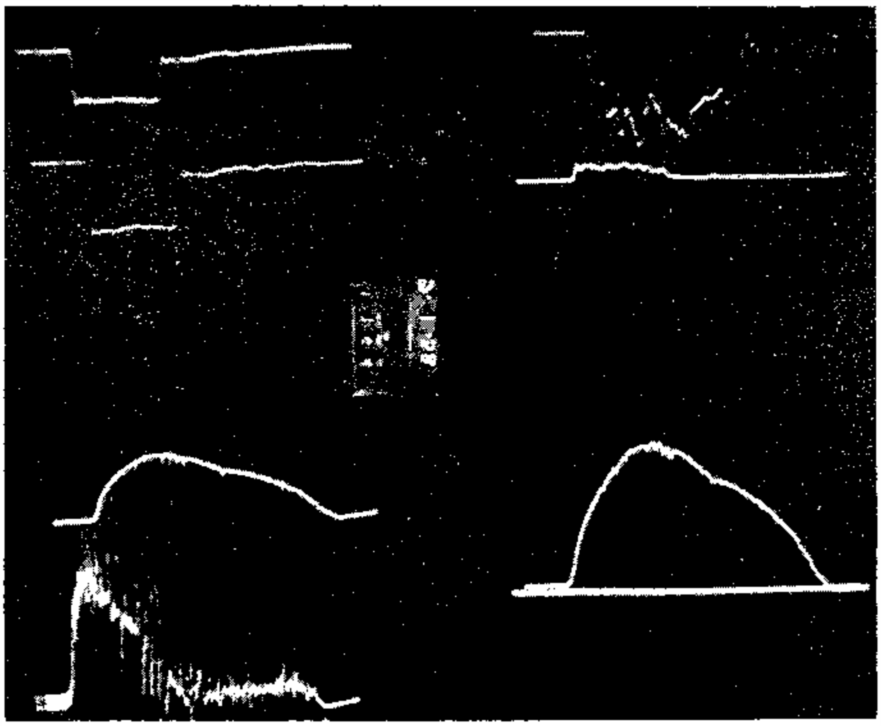


Fig. 6. Shot No. 0144, 9 January, 1958

Chan. 1. } *Langmuir Probes*
Chan. 2. }
Chan. 3. Ip (1)
Chan. 4. V/tn
Chan. 5. } *Langmuir Probes*
Chan. 6. }
Chan. 7. Is
Chan. 8. Nil

Probe Measurements

Probes inserted into the torus body provide another group of signals which may be used to measure certain properties of the discharge. These probes may be of three main types.

Magnetic probes are inserted into the discharge to measure the magnetic fields present in the plasma. The outputs from the search coils are integrated and up to eight measurements may be made simultaneously.

Langmuir probes may also be used to measure the ion current to the torus wall and at other points in the discharge. A four channel probe has been used for this work which may lead to another method of measuring electron temperature. An example of the signals obtained is shown in Fig. 6.

The measurement of the attenuation by the plasma of 4mm microwaves provides a means of estimating the density[4]. For this purpose transmitting and receiving horns are placed opposite each other and the signal received is recorded.

Operational Signals

Signals derived from other equipment are also recorded to establish a correlation between experiments. These signals include the gating pulse for the neutron counters and signals from the time resolved spectrograph gating circuit and the image convertor streak camera.

Conclusion

Unless great care is taken all these signals are subject to severe interference due to the large electric and magnetic fields associated with the machine. There is also a considerable risk of pick-up from the 3kW radio frequency pre-ionization equipment which is used to make the gas weakly conducting before the discharge. For these reasons it is desirable to use shielded balanced twin cables and balanced oscillograph amplifiers.

Acknowledgments

The techniques described in this article are the result of close collaboration with the members of the Controlled Thermonuclear Research Group, the Electronics Division and the Engineering Services Division at Harwell.

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NOTE

Numerous other references are given in a Bibliography on Thermonuclear Research by C. S. Sabel issued by the U.K.A.E.A.

ZETA

(The Main Recording and Monitoring Equipment)

By A. E. Cawkell* and R. Reeves*, A.M.I.E.E.

To operate and evaluate the results obtained with ZETA a large number of signals have to be recorded and monitored. The derivation of these signals has been described elsewhere[1]. In this article the main recording and monitoring equipment is described. This comprises an eight channel oscilloscope with photographic facilities and a two channel oscilloscope incorporating 'Memotron' transient storage tubes.

THE recording and monitoring equipment for ZETA consists of an eight channel measuring oscilloscope with photographic facilities, which is used for recording and measuring transient waveforms, and a two channel oscilloscope incorporating 'Memotron' transient storage tubes, which is used for visual monitoring.

Information from the ZETA torus is fed to the equipment through 600Ω lines, and the instrument was designed and developed for, and in very close consultation with, the Atomic Energy Research Establishment, Harwell. The derivation of the signals is described in an accompanying article by E. P. Butt[1].

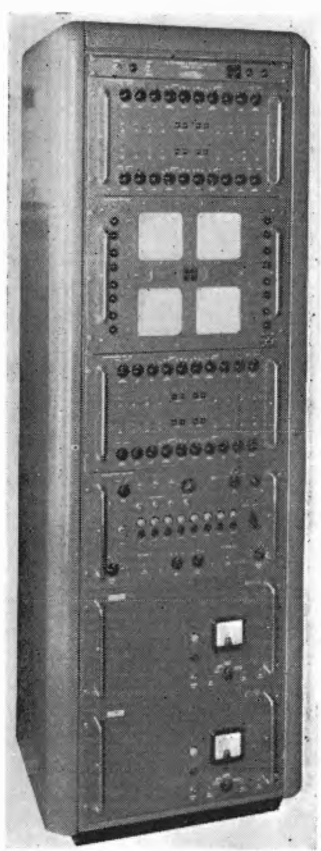


Fig. 2. Eight channel oscilloscope

General Arrangement (see Fig. 1 (p. 116))

The eight channel instrument (see Fig. 2) may be divided into a number of sections:

(1) Two units each consisting of four identical X and Y amplifier channels.

(2) The cathode-ray tube unit embodying four 20th Century D6BSQ double gun tubes, with the usual auxiliary circuits.

(3)(a) A unit consisting of a gated time marker generator, and mechanically part of the same unit.

(b) Two completely separate scan generators.

(4) Photographic control panel.

(5) Four Cawkell PE31A 250 mA three-outlet stabilized power supplies, inter-connected to provide a variety of operating voltages.

(6) A stabilized variable e.h.t. supply adjustable in the range 2 to 6kV at 3mA for the tube guns.

The instrument is built on panels and chassis occupying both the front and rear of a six foot rack.

The two channel oscilloscope (Fig. 3), may be subdivided as follows:

(a) A two channel X and Y amplifier identical to those fitted in the eight channel equipment.

(b) The tube section consisting of two identical 'Memotron' cathode-ray tubes together with the flood gun, writing gun, and erase networks and other networks and controls.

(c) A Cawkell PE31A stabilized h.t. supply.

(d) A continuously variable stabilized e.h.t. supply, of +3kV and −3kV 5mA maximum.

The equipment is mounted on one panel for fitting into a control console, but an alternative version is being made suitably sub-divided for fitting into a standard rack.

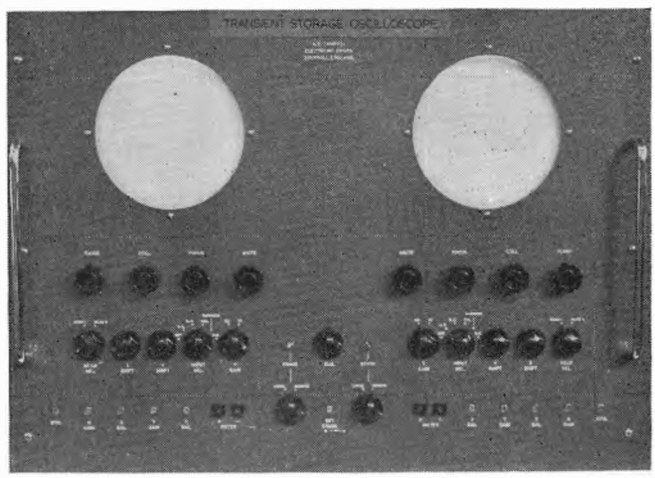


Fig. 3. Two channel transient storage oscilloscope

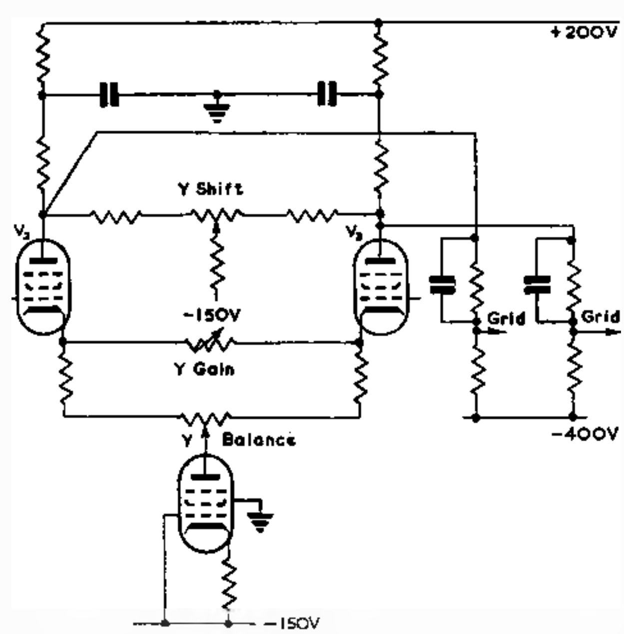


Fig. 4. Basic circuit of Y amplifiers

Some of the more interesting features will be described;

* Cawkell Research & Electronics Ltd.

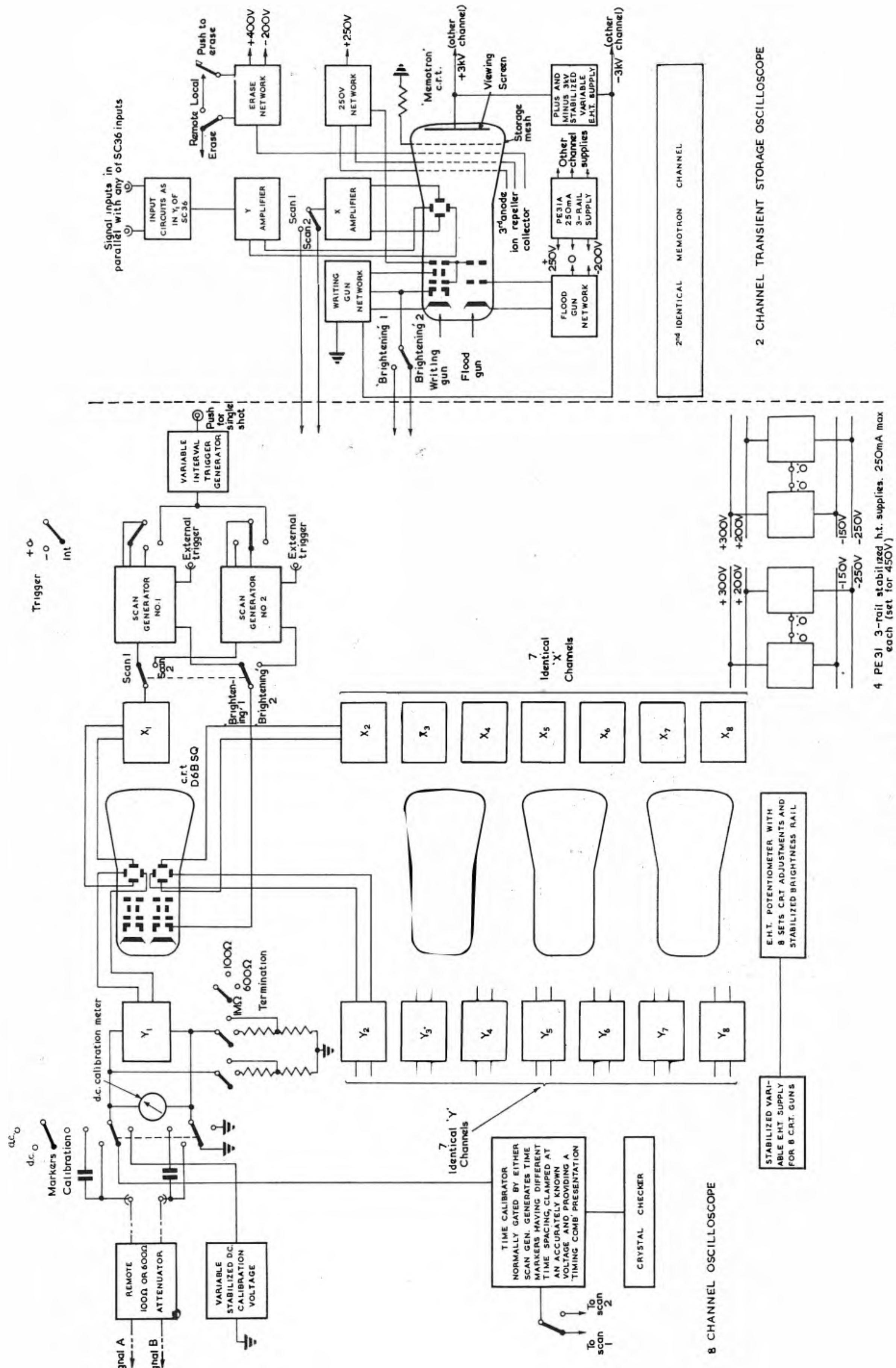


Fig. 1. General arrangement of the equipment

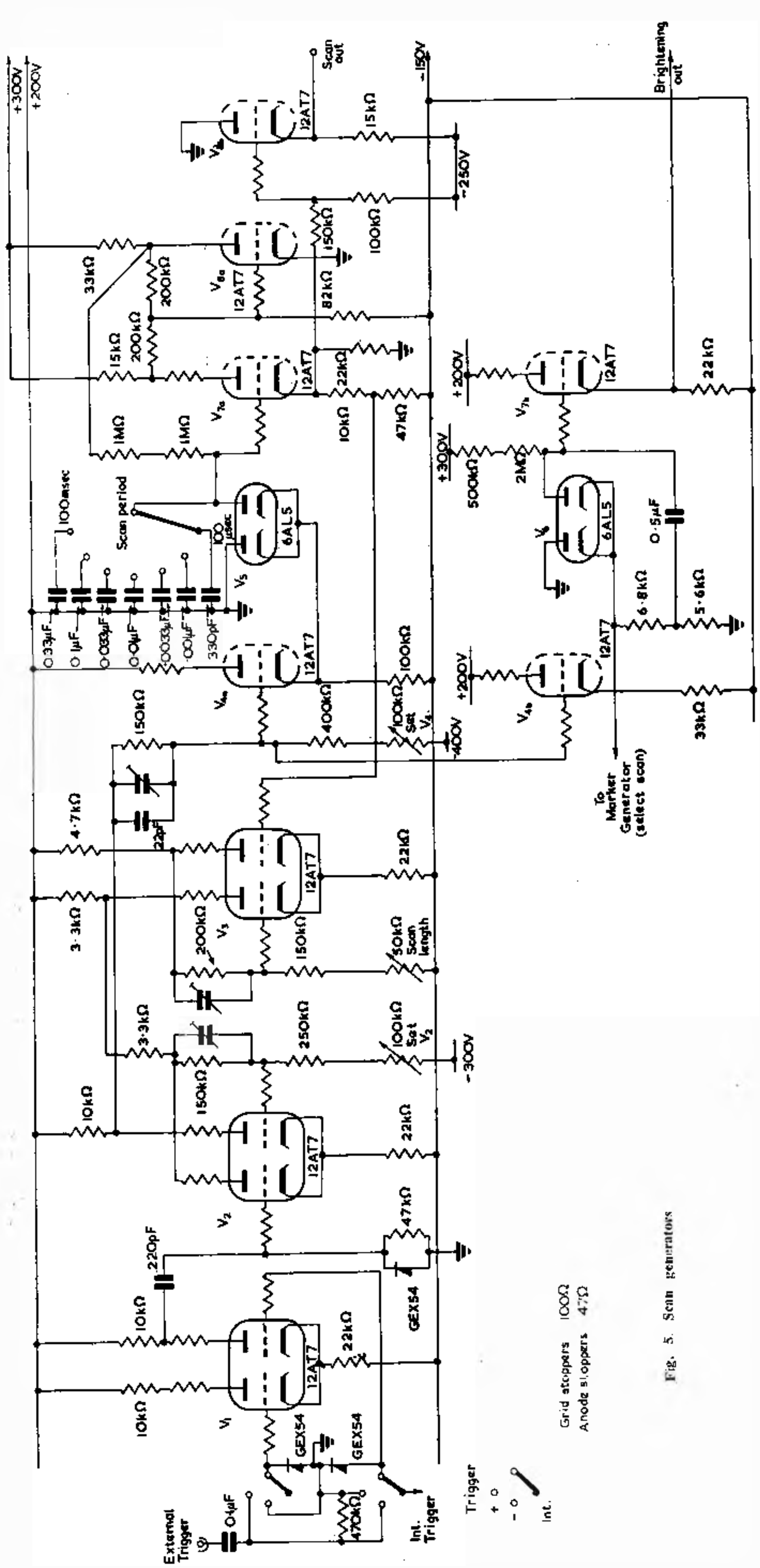


Fig. 5. Scan generators

in general, the description relates to the 'Mk.2' equipment, which contained some modifications.

Y Amplifiers

The basic circuit of the input stage is shown in Fig. 4.

A difficulty in the design of push-pull d.c. coupled oscilloscope amplifiers lies in the application of Y shift. In this case a potentiometer is connected between the anode of V_2 and V_3 as shown; when the slider is central, equal currents flow through the anode resistors to −150V. When the slider is moved, the difference in potential resulting from the unequal flow of current through the resistors is passed to the following stage.

This form of shift permits the amplifier to remain balanced throughout and does not affect the high frequency response.

The pentode 'tail' was introduced to increase the rejection of 'in phase' interfering signals from the intense magnetic fields created by ZETA.

Such an arrangement has been described in detail elsewhere[2,3] and a 'transmission factor' of 1 000 was achieved without selection of valves or components, which was adequate for the purpose.

Ganged switches permit the termination at the grids to be altered, depending on the type of external attenuator in use.

When the attenuators are in use the effective transmission factor is reduced, as 1 per cent tolerance resistors are used, and unequal grid voltages may be produced by interfering fields.

A push-pull signal input of 1V produces a half screen deflexion, which amounts to a gain of about 250.

High frequency compensation is introduced in the output stage cathode coupling, the rise-time being 0·3μsec (critically damped).

A simple form of d.c. calibration is provided by introducing a d.c. voltage, monitored by a sub-standard meter on to one input grid, the other being earthed.

The 'Y balance' control compensates for any differences between the $1/g_m$ of V_2 and V_3; when correctly set the 'Y gain' may be varied from minimum to maximum, without any spot movement occurring.

Scan Generators

The circuit contains some novel features and so is shown in detail (Fig. 5). The outstanding characteristics of the arrangement are

(1) Infallible triggering.

(2) D.C. coupled throughout, and therefore no undesirable effects as recurrence rate is altered.

(3) Very wide range; in conjunction with the X amplifier a maximum sweep speed of about 200V/μsec may be obtained. Using conventional 'speeding up' techniques much higher speeds should be possible. Apart from the 'brightening' arrangements, the circuit is capable of generating slow linear sweeps whose duration is only limited by the size of the main sweep capacitor. If necessary, this could be 'increased' by feedback methods[4].

(4) Constant 'brightening' voltage is applied to the c.r.t. grid without the use of an r.f. oscillator, even on the slowest sweep.

Assume for the moment that V_2 grid is the triggering terminal. Then any positive going 15V waveform, regardless of its shape or rise-time, will trigger the scan, as V_2 is a Schmidt trigger.

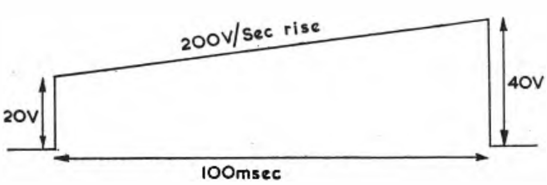


Fig. 6. Brightening waveform (100msec sweep)

The condition before receipt of a trigger is as follows:

V_{2b} on, V_{4a} grid negative, V_5 conducting holding scan capacitor voltage earthy, V_{3b} off.

Upon receipt of a positive trigger V_{2a} switches on and V_{2b} off; the rise in voltage at V_{2b} anode is communicated via V_{4a} to V_5 cathode which, becoming positive, is cut off.

The capacitor C, as selected by the 'scan period' switch, now commences to charge through the 2MΩ resistor R. This rising voltage is inverted in V_{7a} and V_{8a}, and fed back to the other end of the charging resistor R. The sweep voltage is therefore linearized by a 'd.c. coupled bootstrap' effect, the current through R being held constant.

The voltage on C is allowed to rise to about 10 per cent of the supply voltage.

The sweep terminating action is then as follows:

The sweep voltage is fed back to the comparator V_3 via V_{7a} cathode.

V_{3b} comes on, and V_{3a} off, the rising voltage from V_{3a} anode switching on V_{2b}. V_5 now comes on via V_{4a}, and the sweep voltage rise is terminated, C discharging through V_5.

The resulting fall in voltage at the cathode of V_5 switches V_{3b} off, and V_{3a} on. The circuit and potentials associated with V_2 are such that there is a form of backlash. When V_{2b} is on, its grid is well positive but when off only just sufficiently negative.

When V_{3a} comes on therefore, the negative drop applied to V_2 grid is insufficient to cut off V_{2b}.

The 'set V_2' control is adjusted so that the positive excursions of V_{3a} anode are sufficient to switch on V_{2b}, but negative excursions are insufficient to cut it off.

Brightening Circuit

V_{4b} cathode waveform, which is a rectangular pulse lasting for the duration of the sweep, is used to gate the time calibrator.

The brightening pulse is coupled via cathode-follower V_{7b} to each c.r.t. grid via an appropriate 5kV working capacitor, each grid circuit being completed by a 1MΩ resistor—the maximum permissible value.

If the brightening level is to be maintained constant to, say, 5 per cent for 100msec—the longest sweep duration, 5kV working capacitors of at least 1μF are required; these are inconveniently large and expensive.

Assume therefore that 0·1μF capacitors, which are a reasonable size, are to be used.

If 20V is a suitable value for the brightening pulse then the initial current drawn from C is $(20/10^6)$A. To maintain this current, V must be increased at the rate of I/C volts/second, which is (from $Q=CV$) 200V/sec. The brightening waveform required is therefore as in Fig. 6.

V_6 is normally conducting. V_{4b} is coupled to V_{7b}, the brightening output cathode-follower, via a suitable level adjusting potentiometer and a 0·5μF capacitor.

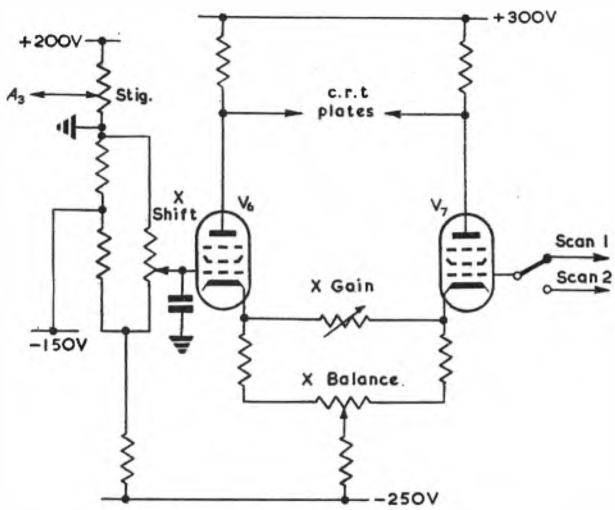


Fig. 7. 'X' amplifiers

Upon receipt of the positive rectangle from V_{2b} anode, V_6 is cut-off, and the 0·5μF capcitor charges at 200V/sec through R. The pulse at V_{7b} grid is therefore of the form shown in Fig. 6 and compensation for the short time-constant coupling to the tube grids is achieved. At the end of the scan period V_6 conducts and discharges the capacitor.

X Amplifiers (Fig. 7)

The single ended output from the scan generator is made push-pull in the cathode coupled stage V_6, V_7.

Markers are available for time calibration, and the 'X gain' control provides for expansion from ½ to 10 screen diameters.

The 'X balance' control is adjusted so that the spot may be positioned at the left-hand edge of the screen, and no movement is produced when the 'X gain' is varied.

To achieve optimum linearity, careful attention is paid to operating potentials. The velocity error is too small to measure on the screen.

Marker Generator (Fig. 8)

The square wave from the selected scan generator is inverted and applied to the cathode-follower V_{1b}. The diodes clip the waveform and clamp it between +10V and zero. At +10V (i.e. no applied square wave) V_2 is not operating. At zero (i.e. square wave applied), V_2 generates pulses 2μsec long, spaced by 10μsec, which in turn synchronize a second multivibrator generating markers spaced by 100μsec. A third multivibrator, synchronized from the second, generates markers spaced by 1msec.

The outputs from the three multivibrators are fed via three cathode-followers (V_{3b} only is shown as an example) to the adding and clamping circuits.

The output from V_2 is fed via an additional cathode-follower V_{3a} to the crystal checker.

To check that V_2 is generating markers time spaced accurately at 10μsec, it is made to 'free run' by depressing S_1 thereby altering the biasing conditions. VR_1 is then adjusted until the tuning indicator 'eye' closes, showing that signals are passing through the 100k/s 'crystal gate'. A similar arrangement has been described elsewhere[5].

Negative going markers are available from the cathode-followers (e.g. V_{3b}), at 10, 100 and 1 000μsec spacing, from where they are fed to the adding and mixing circuit which includes MR_1 to MR_6. When say, the '10μsec' switch is made, a pulse causes MR_1 to conduct, and current passes through R_3, MR_1, and R_1 to -150V. As a result a $\frac{1}{3}$V pulse appears across R_2. It will be observed that the diode resistance, being small compared with 50kΩ, has little effect upon the calibration.

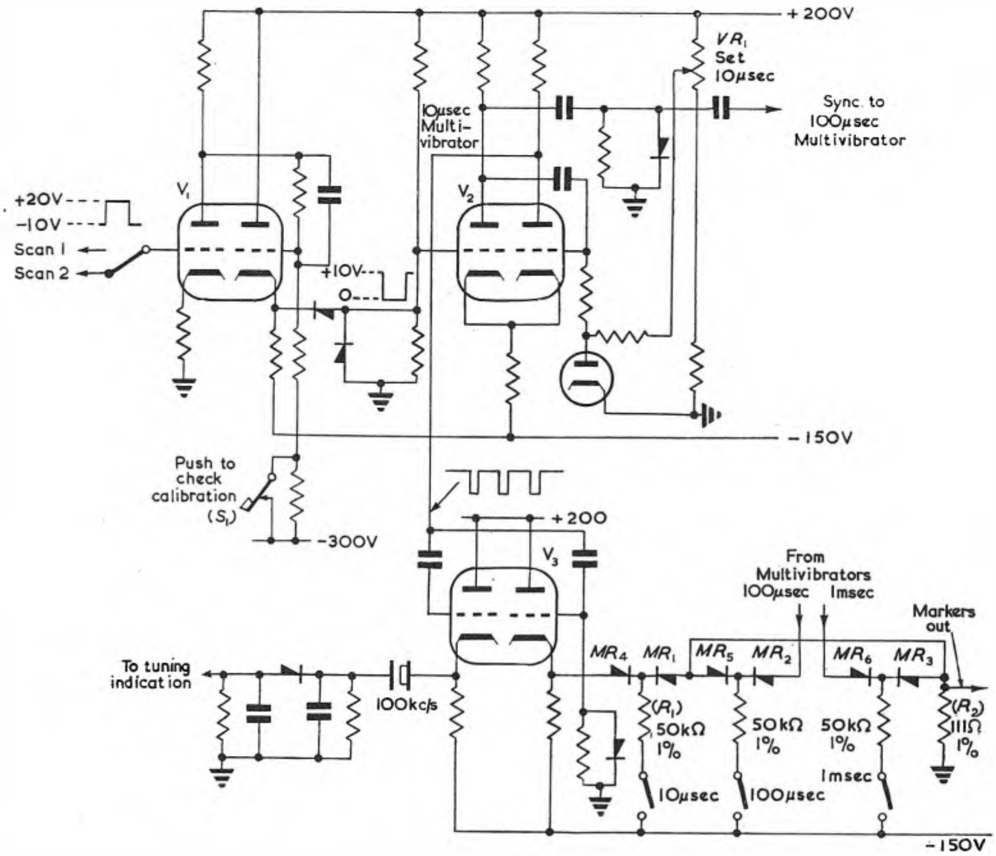


Fig. 8. Time calibration marker generator

If the '100μsec' and '1msec' switches are made, the pulses 'stack up' and as a result every tenth pulse will have an amplitude of $\frac{2}{3}$V and every hundredth 1V.

The pulses are fed into the Y amplifier as required for both voltage and time calibration purposes.

Stabilized Power Supplies (Fig. 9)

Some points of interest will be described. Four standard Cawkell PE31 model 'A' supplies are fitted (Fig. 9(a)). These result a wide range of output voltages is possible while the locking plug, whereby certain control resistors and a tapping on the transformer h.t. secondary are changed. As a result a wide range of output voltage is possible while the ripple level remains at less than 1mV, and the output impedance at less than 1Ω under all conditions.

V_1 is the usual 'series valve', which has a stabilized screen supply (not shown), V_2 being the controlling amplifier.

Three stabilized 'outputs' are available from one transformer-rectifier-smoothing system.

The 'positive' and 'negative' outputs are obtained in the normal manner.

The '0' rail is stabilized by V_4, and shunt stabilizer valve V_3. A total current of 250mA may be drawn, and up to 100mA more current (provided through V_3) can be taken between 'plus' and '0', than between 'minus' and '0', if desired. The '0' tap may be varied by 150V within, and relative to, the outer rail voltages.

In this case (see Fig. 1) two independent groups are provided by suitable inter-connexion, each giving plus 300, plus 200, minus 150, and minus 250V.

In the case of the X and Y amplifiers, by connecting the circuit between $+300$ and -250V, the output valves are capable of providing the necessary voltage swing to fully deflect the spot on the somewhat insensitive double gun tubes.

The greater part of the oscilloscope runs from the plus 200 and minus 150V supplies.

It will be observed that some current must always be taken from 'plus' and '0', otherwise V_3, V_4 circuit will not stabilize.

The e.h.t. supply, providing for 2 to 6kV variable stabilized voltage at 3mA is shown in Fig. 9(b).

V_3, V_4 comprise a self-excited oscillator operating at the resonant frequency of the h.t. coil L_5, about 25kc/s; L_1-L_5 are wound on a Ferroxcube double-U core. The output from L_5 is rectified and voltage doubled by V_5, V_6, with C_1, C_2. R_1, R_2 is a potential divider connected between e.h.t. negative and some fraction of $+300$V (obtained from one of the h.t. supplies).

Changes in the e.h.t. output voltage therefore appear at the grid of V_{1a}, being the junction of R_1, R_2. These changes are applied with the correct phase to V_3, V_4 via d.c. amplifier V_1, and cathode-follower V_2.

The e.h.t. output voltage is substantially R_1/R_2 times the voltage set by VR_1.

A change of this potentiometer causes a relatively slow change of e.h.t. voltage because of the large back coupling time-constants involved.

However C_3, C_4 comprise a capacitive divider which permits C_3 to be large enough for smoothing purposes without imposing a large delay in the feedback loop which could cause instability.

R_3, C_5, and R_4, C_6, prevent the oscillator from working at the natural frequencies of L_1-L_4 or any combination thereof. Such frequencies are much higher than 25kc/s.

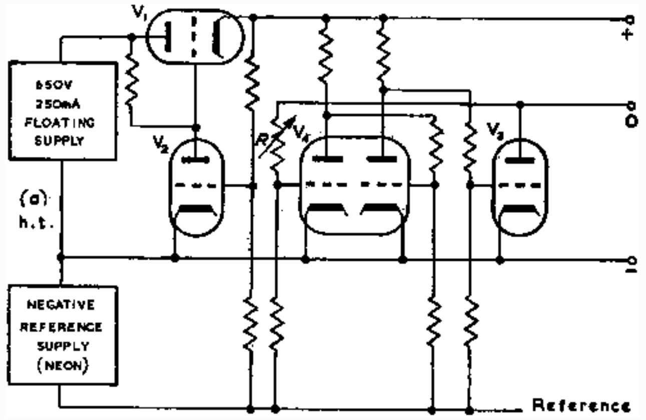
The heater of V_6 is supplied from a further transformer winding (not shown).

Two-Channel Transient Storage Oscilloscope

The heart of the instrument is the 'Memotron' transient storage cathode-ray tube, manufactured by the Hughes Aircraft Corporation.

The general arrangement of the tube, with its associated circuits, is shown in Fig. 1.

Two electron guns are contained within the glass envelope.



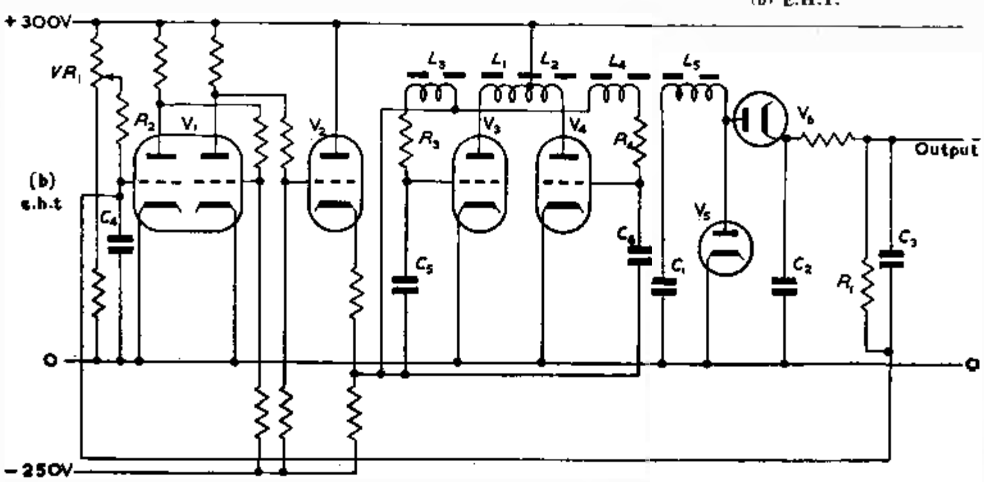


Fig. 9. Stabilized power supplies
(a) H.T.
(b) E.H.T.

The 'writing gun' has a normal electrode assembly, and the usual X and Y deflector plates to which a time-base and signal are applied. High velocity electrons are produced as a total accelerating potential of more than 3kV is used.

The 'flood gun' has a simple assembly, containing no grid or deflector plates, and a single anode only, to which an accelerating potential of only about 200V is applied; low velocity electrons are produced.

Between the two guns and the viewing screen (which is coated with a P1 green phosphor) are four additional electrodes—the 3rd anode, 'ion repeller', 'collector' and 'storage mesh', in that order.

The storage mesh is coated on one side with a dielectric material. It is continually sprayed with low velocity electrons from the flood gun, and under these conditions assumes the 'unwritten' potential, around 0 volts.

As soon as the brightening potential is applied to the writing gun, the beam comes on, and high velocity electrons impinge upon the storage mesh, as controlled by the deflector plates.

The bombarded areas of the mesh produce secondary emission, and assume the potential of the adjacent secondary electron collector—about 150V. These areas are transparent to flood gun electrons, which are accelerated to the viewing screen (which is at +3kV), producing a bright trace.

Flood gun electrons continue to pass through the 'holes' made by the writing beam, and the 'picture' remains upon the screen indefinitely at a constant brightness.

The collector is connected to the cathode of a cathode-follower, whose grid voltage may be altered.

The potentials of most of the other electrodes may also be altered, and the Memotron tube is adjusted so that there is the maximum contrast between the trace and the 'background' noise. The contrast is further improved by pulse modulating the storage mesh and collector voltages with a simple 'free-running' pulse generator (circuit not shown).

When present, the background appears as if by the projection of a fine wire mesh upon the screen.

The flood gun, by causing secondary emission, continues to hold the written areas of the storage mesh at +150V, until it is desired to 'erase'. This is achieved by momentarily lowering the collector voltage which is allowed to decay with a time-constant of 0·01sec.

'Erasing' is controlled by a push-button on the panel, or alternatively from a remote position.

The tube will behave like a normal cathode-ray tube if the collector is held at a low voltage. A further possibility is to arrange for a transient to remain for a pre-determined interval, being erased by an 'auto-erase' circuit.

At present the fastest transient that can be displayed is one lasting for perhaps 10μsec. However an integrating effect occurs if the waveform is repeated, by cumulative storage action. If the flood gun is biased off between sweeps the effect is enhanced, and regardless of the interval between transients, eventually a trace will appear, the number of repititions necessary being dependent on the speed of the transient.

The resolution is 50 lines/in.

Associated Circuits

A Cawkell type A stabilized h.t. supply, and a variable stabilized e.h.t. supply producing plus and minus 3kV stabilized at 5mA, are common to both 'Memotron' channels.

Their e.h.t. voltages are generated from a similar e.h.t. supply to that already described; the positive voltage is obtained by an additional rectifier system, which is not referred to for stabilization. It is held constant against mains fluctuations by the negative reference feedback, but variations arising from load changes are very small, as the positive rail only supplies the viewing screen.

Apart from the collector voltage already mentioned, the electrodes are supplied from suitable bleeder networks—heavy current being necessary in some cases. The ion repeller for example may pass up to 4mA.

The X and Y amplifiers are identical to those in the SC36 instrument. It is possible to connect the Y inputs to any of the eight channel inputs for monitoring purposes, while the scan and brightening voltages are derived from either of the two SC36 scan generators.

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Controlled Saturation in Transistors and its Application in Trigger Circuit Design

(Part 1)

By N. F. Moody*, A.M.I.E.E.

It is usually considered that the speed of a transistor switching circuit becomes grossly degraded if saturation is allowed to occur. That this need not be so, if the saturation is appropriately controlled, is demonstrated by the design of a trigger circuit. This saturated circuit is able to approach the performance of its non-saturated counterpart, which it may often replace with economy in both power consumption and cost.

Part 1 of the article is devoted to a study of carrier storage in both saturated and non-saturated transistors. This study leads to a concept known as 'controlled saturation,' which defines the maximum charge storage the transistor can exhibit whether saturated or not.

In Part 2 this concept is applied to develop a trigger circuit of excellent resolving time, whose output terminal is able to handle heavy currents and present a low impedance. The circuit finds an important application in digital computers employing passive routing.

TRIGGER circuits and switching circuits fall into two classes: those in which the transistors saturate, and those in which saturation is prevented. Circuits using saturated transistors are generally simpler than their non-saturated counterparts, but saturation can lead to large carrier storage which, in turn, destroys the speed of operation.

The first Part of this article is a theoretical study of saturation phenomena, directed towards minimizing the penalty of lost speed. A concept, known as 'controlled saturation,' is developed; and when this is applied, in conjunction with correct 'turn off' techniques, it is found that the speed loss due to saturation may be quite small.

The second Part of the article applies the theory to the design of a trigger circuit. This circuit is of a form peculiarly suited to digital computers which use passive routing. As such, it is required to meet heavy and variable output loads, and to present a low output impedance. The present design parallels the trigger circuit of a companion article, where saturation is not permitted[1]. The saturated version is able to meet its specification in all important respects, and has the advantage of reduced component costs and of lower power consumption.

Any conducting transistor stores carriers. For the alloy types, which are the most suitable for switching purposes, storage exists only in the base region. The process of turning the transistor off consists in removing this base charge. Thus a study of turn-off time requires that the charge storage be evaluated, and that charge removal processes be considered.

The theoretical part of the article commences with a comparison of the small signal storage of alloy junction transistors under linear†, and under saturated, conditions. This results in the concept of controlled saturation; a principle which provides a simple method of assessment and control of carrier storage. It is then demonstrated experimentally that the theory may be extended to large signal conditions for transistors of the class designed for switching operation.

The theoretical study concludes by showing that a major fraction of the stored carriers may be eliminated from a saturated transistor as rapidly as from its linear counterpart. A simple expression for this 'first order' turn-off time is then developed, which is applicable whether the transistor is saturated or not. A circuit technique which leads to minimum turn-off time is given.

* *Defence Research Board, Canada.*

† *The term 'linear' is hereafter used to mean 'non-saturated.'*

The Carrier Storage under Linear and Saturated Conditions

The charge storage in a transistor may be studied in any convenient circuit configuration, and the results may be expressed in terms of any pair of the three terminal currents. For the purpose of this article it is convenient to use a grounded collector stage, as shown in Fig. 1, and to develop the equations in terms of base and emitter currents.

If it be supposed that the base current is held constant at some arbitrary value, then the transistor may be made to saturate, or not, according to the value of the emitter resistance R. In the first place let R be made small, so that the limiting emitter current $I_e = I_b/1-\alpha_n \sim I_b\beta_n$, can flow without resultant saturation*. For this linear condition, the stored charge is simply proportional to the product of (rate of carrier flow across the base) and (average transit time). The rate of flow is defined by I_e; the transit time, T_n, by the inverse of the α cut-off frequency, ω_n, expressed in radians/sec. The storage factor Q for a non-saturated transistor, expressed in terms of base current, I_b, is thus

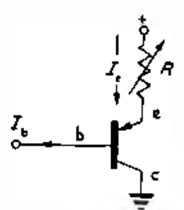

Fig. 1. The circuit used to study carrier storage in a pnp transistor

$$Q = K\,\beta_n\,T_n\,I_b \text{ coulombs} \qquad \text{(1a)}$$

where K is a constant of proportionality (taken as unity in this article).

If, now, R is increased while the base current is held constant, the transistor will eventually saturate, and will remain saturated for all emitter currents $I_{e(s)}$, such that

$$I_{e(s)} < I_e \quad \text{where } I_e \sim I_b\,\beta_n$$

The charge storage under saturated conditions, $Q_{(s)}$, may be determined from an expression developed in Appendix 1. This is

$$Q_{(s)} \sim K\,\beta_n T_n \left[I_b \left(\frac{1 + T_i/T_n}{1 + \beta_n/\beta_i} \right) + \frac{I_{e(s)}}{\beta_n + \beta_i} \left(1 - \frac{\beta_i T_i}{\beta_n T_n} \right) \right] \qquad \text{(7b)}$$

$$\alpha_n \to 1$$

The solution involves a knowledge of the normal parameters β_n and T_n and also of the inverse parameters β_i and T_i.

* *Throughout the main body of the article, I_e I_c and I_b are treated as positive quantities. The subscript n is used to define normal, or forward, parameters; and the subscript i will be used for the inverse parameters which arise when the functions of the collector and emitter are interchanged.*

Influence of Saturation on Charge Storage for Constant Base Current (Small Signal Conditions)

A first step in an assessment of the influence of saturation on turn-off time is a comparison of the stored charge under saturation with the value associated with linear operation. The storage ratio is given by dividing equation (7b) by equation (1a), using the same base current in each. It is permissible to equalize the linear and saturated base currents in this way, for it will be shown that then both circuit forms have very similar properties, and can perform similar duties. The ratio of charge storages thus obtained is

$$\frac{Q_{(s)}}{Q} = 1 - \frac{\beta_n(1-\eta) - \beta_i(T_i/T_n)(1-\eta)}{\beta_i + \beta_n} \quad \ldots\ldots\ldots \quad (7c)$$

where $\eta = I_{e(s)}/I_e$, the fraction of maximum emitter load current which is actually drawn.

When η is equal to unity, the circuit lies at the boundary condition between saturated and non-saturated states. Then the fraction vanishes, and $Q_{(s)} = Q$, as might be expected. Thus, at maximum emitter current, there is no additional storage.

The behaviour of the storage with emitter current in the saturated transistor may be obtained by differentiating equation (7c). This yields

$$\frac{d\left(\frac{Q_{(s)}}{Q}\right)}{d\eta} = \frac{\beta_n}{\beta_n + \beta_i}\left(1 - \frac{\beta_i T_i}{\beta_n T_n}\right) \quad \ldots\ldots\ldots\ldots \quad (8a)$$

and if

$$\beta_i T_i / \beta_n T_n < 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

the storage will diminish as the magnitude of $I_{e(s)}$ diminishes.

It will be noticed that the conditional equation (9) involves the ratio of two time-constants. These are the conventional grounded emitter time-constant $\beta_n T_n$, and the corresponding time-constant $\beta_i T_i$ which arises from inverted use. An insight into the relationship between these time-constants is given readily. A symmetrical transistor must, by definition, have equal time-constants, and therefore its storage is completely unchanged under saturated operation.

Most transistors are asymmetrical, geometrical changes being made in the emitter and collector areas in order to improve the forward (or normal) characteristics. Tests on ten alloy junction transistors (Table 1) show that as a result of these geometrical changes the normal time-constant is increased at the expense of the inverted time-constant. It seems plausible to expect that all alloy transistors will obey the same physical laws. Then, only symmetrical transistors will generate the boundary condition; hence all others must exhibit less storage under saturation.

TABLE 1

The Parameter $\beta_i T_i/\beta_n T_n$ for GE 2N123 Transistors
(Taken at I_c=5mA E_c= −2V)

SAMPLE NO.	NORMAL PARAMETERS		INVERSE PARAMETERS		$\beta_i T_i$ (μsec)	$\beta_n T_n$ (μsec)	$\frac{\beta_i T_i}{\beta_n T_n}$
	β_n	T_n (10^{-8})	β_i	T_i (10^{-8})			
53	88·5	31·2	19	76	1·44	2·86	0·5
341	59·5	32	12	120	1·44	1·9	0·76
269	70	24·8	14·6	82	1·20	1·74	0·69
439	62	30	8·4	136·5	1·15	1·86	0·62
107	115	28	11·5	120	1·38	3·2	0·43
203	91	27·4	10·8	76	0·82	2·5	0·35
175	89	28·6	11·5	125	1·44	2·55	0·57
359	66	39·9	9·75	120	1·17	2·61	0·45
182	61	36·4	9·1	108	0·98	2·21	0·44
253	98·5	24·1	11·5	76	0·87	2·38	0·37

The results of the preceding analysis may be stated formally as follows:—

The storage in a non-saturated transistor is given by the product of the base current and the grounded emitter time-constant. If, moreover, the grounded-emitter time-constant of the inverted transistor is less than the corresponding normal value, then the storage in saturation (for the same base current) will be less than the non-saturated value.

The corollary of this statement defines the concept of controlled saturation. It is:—

The value of the base current of a transistor sets the maximum stored charge which can exist. If the time-constants are related as shown above, the maximum storage will occur only in the non-saturated state.

Thus, control of base current yields controlled saturation and, all other things being equal, should yield a turn-off time which never exceeds, and should be usually less, than that of its linear counterpart. Since it has been suggested that both circuits can perform similar duties, the presence of saturation appears advantageous! Whether or not this conclusion is true will depend on the manner in which saturation influences carrier removal—a subject to be treated later.

Storage under Large Signal Conditions

The circuits of Part 2 involve saturation under large signal conditions. The storage equations then become non-linear, since the parameters are themselves dependent on current magnitudes. The equations must then be evaluated for the parameter values relating to a chosen current range, and taken on selected samples. Since the labour of making such measurements is great (see Appendices 1 and 2 for details) conclusions must rest on a limited selection of samples supported by practical circuit performance.

Three samples, among the ten listed, were selected for detailed large signal study. Of these, sample 53 displays rather average characteristics. In particular the ratio of the key time-constants $\beta_i T_i/\beta_n T_n$ is typical. Sample 341 was selected as having a time-constant ratio most likely to result in a storage increase with saturation; and sample 203 was chosen as the least likely to have a storage increase.

Storages were calculated and measured for each sample at a base current of 1·2mA; with a maximum emitter current (corresponding to linear operation), and also under saturated conditions with emitter currents of 21, 10, and 5mA. The saturated conditions correspond closely to the practical requirements of Part 2.

Details of the calculations are given in Appendix 1, and the method of measuring the storage is discussed in the section on the technique of carrier clearance. In this section discussion is limited to the results, which are given in Table 2.

It is seen from Table 2 that the measured storages give satisfactory support to the mathematical theory. Samples 203 and 53 show a considerable storage reduction under saturation, while the worst sample, 341, maintains a relatively constant storage. There is a possibility that a few samples may exhibit a slight storage increase in large signal saturation; but as engineering requirements are satisfied by a relatively approximate storage figure, this is not of practical significance. Circuit experience based on some 30 transistors indicates that the concept of controlled saturation remains valid for design purposes.

The inverse characteristics are relatively insensitive to operating conditions; and because of this, any transistor whose forward characteristics are maintained at high

TABLE 2

Large Signal Storages

(*Standardized to* I_b = 1·12mA, I_{br} = 20mA)

TRANSISTOR NO.	$\eta = \frac{I_b(s)}{I_b}$	$I_1(s)$ (mA)	$Q(s)/Q$ (CALCULATED)	$Q(s) \times 10^{-9}$ COULOMB (CALCULATED)*	$Q(s) \times 10^{-9}$ COULOMB (MEASURED)*
53	1†	50	1	5.0	4.13
	0.423	21.2	0.55	2.78	2.06
	0.188	9.42	0.48	2.39	2.30
	0.098	4.41	0.49	2.46	2.04
203	1†	51.5	1	4.12	3.1
	0.41	21.2	0.45	1.85	1.89
	0.183	9.42	0.39	1.63	1.75
	0.086	4.41	0.35	1.43	1.82
341	1†	33.6	1	1.85	1.92
	0.614	21.2	1.05	1.965	1.7
	0.28	9.42	1.12	2.07	2.02
	0.13	4.41	1.12	2.06	1.88

* Probable accuracy ±20 per cent.

† Boundary of saturation, $I_e(s) = I_e$; $Q(s) = Q$.

The measured value of $Q(s)$ at $\eta = 1$ is taken at $V_c = -2$V and is corrected to $V_c = 0$.

current should be acceptable for saturated use. Transistors designed for switching operation fall into this class.

The parameters used in the calculations were those which relate to zero collector voltage (as the theory requires), and the betas were direct current values. Such parameters are inconvenient to measure. Experience shows that little error is caused by substituting orthodox small signal parameters taken at low collector voltage (say 2V). The forward parameters are sufficient for engineering purposes, and transistor acceptability may be based on their values at the highest emitter current. It is common for the maker of switching transistors to supply this data, so that the engineer may assess his transistors for saturated use without the need to make further measurements.

Control of Saturation by Base Input Charge

The preceding work has established a maximum steady state storage relating to a maintained base current. If the base current should be interrupted at a time t, such that the input charge q is less than the saturated storage Q_s, that is

$$Q_s > q = \int_0^t I_b \, . \, dt$$

then, in fact, the storage will be limited to q.

The principle is illustrated in Fig. 2 by a circuit which will be used in Part 2. Here a capacitor defines the input charge, in terms of the applied voltage step. The collector of the transistor is able to deliver a considerable current and an enhanced charge to the external circuit. Under linear conditions it is shown in Appendix 1 of Part 2 that I_c is given by the expression:—

$$I_c \sim q/T_a \left[e^{-t/\beta_n T_n} - e^{-t/Cr} \right] \quad \ldots\ldots\ldots\ldots\ldots \quad (13a)$$

for $T_a \beta_n \gg rC$

and the charge Q_c delivered by the collector is

$$Q_c \sim q \, \beta_n \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (14)$$

These equations may be illustrated by application of a 6V step via 100pF to a transistor whose characteristics are:—

$r = r_b' = 100\Omega$; $\beta_n = 70$; $T_n = 2{\cdot}5 \times 10^{-8}$sec.

The input charge is 6×10^{-10} C, which gives rise to about $4{\cdot}2 \times 10^{-8}$ C at the collector. The collector current rises to 24mA in 10^{-8}sec, and this current decays with a time-constant of $1{\cdot}75 \times 10^{-6}$sec. Attention is drawn to the large collector current, to its rapid rise, and to its subsequent slow decay.

Removal of Stored Charge

Stored charge in a transistor may be removed by application of a reverse base current. The carrier elimination process which results may be idealized by dividing it into three separate phases. During the first phase the transistor behaves as a passive element, and the charge is eliminated as though stored in a capacitor. At the end of this phase a majority of the charge has been dissipated.

There follows a second phase, in which the elimination of carriers is controlled by the transistor parameters, rather than by the base current. This phase is associated with an exponential charge dissipation of time-constant T_a. It is substantially completed after a few time-constants. For many purposes the process of turn off may be considered as complete at the end of this second phase.

The elimination of carriers appears to conclude by a non-analytical process, phase 3, in which a small residual storage is dissipated. The third phase occupies more time than the preceding phases.

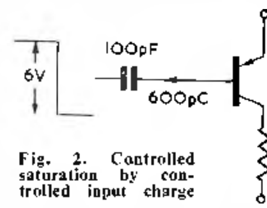

Fig. 2. Controlled saturation by controlled input charge

While all phases exist, whether the transistor is saturated or not, the third phase appears to be much more pronounced in saturated units. The discussion which follows will restrict itself mainly to the saturated state.

Phase 1—Reverse Base Current Source Limited

The three terminal currents of a transistor must always obey Kirchhoff's laws, so that any charge injected into the base must be associated with a departure of charge at the remaining pair of terminals. Now it is known that current flow across the emitter and collector junctions is sensibly limited to minority carrier flow. Since the stored charge consists of minority carriers, it follows that the charge injected into the base ejects an equal charge of minority carriers*. Thus the stored charge Q_s is simply removed at a rate

$$Q_s = \int_0^t I_{br} \, dt \quad \ldots\ldots\ldots\ldots \quad (15)$$

The first phase of turn off is characterized by an absence of any restriction on the value of reverse base current which may be injected; other than that imposed by impedances extrinsic to the transistor. The base spreading resistance, which is usually treated as extrinsic, is negligible at the usual currents involved, and all other impedances of significance are in the external circuit, and so can be controlled by the designer. The mechanism by which the transistor can accept unlimited base current is as follows.

At the instant of turn off, there is a finite carrier density at the base-emitter boundary of any transistor—and, for a saturated transistor, at the collector boundary also[3]. The transistor cannot restrict the base current until the density at these boundaries has been reduced to zero. This initial situation soon gives way to a state in which the maximum base current is limited by the carrier density gradients at both collector and emitter. Phase 1 ends when the sum of ejected currents due to these gradients is less than the available reverse base current.

* *In fact, there is another component of loss due to recombination of minority carriers within the base itself, so that the rate of charge elimination is further enhanced. For the massive reverse base currents considered in this article such recombination becomes a second order effect and it is neglected.*

In a practical situation the maximum available base current is limited by its source. If, as is quite usual, its magnitude is constant during phase 1, the time taken to leave the phase is

$$t_1 = Q_1/I_{br} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (15a)$$

where Q_1 is the portion of stored charge removed in the first phase.

At the end of phase 1 both the emitter and collector terminals will assume reverse potentials, since the transistor can no longer absorb the available base current.

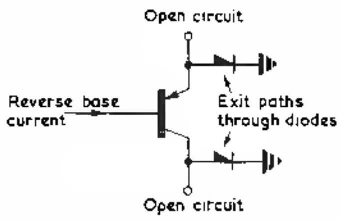

Fig. 3. Exit paths for reverse current aid rapid carrier clearing

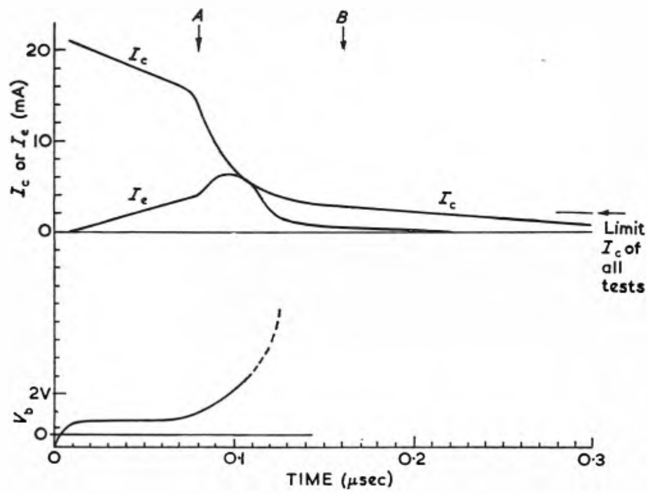

Fig. 4. Emitter, collector currents and base voltage during turn off
Sample 202 $I_e = 9{\cdot}4mA$ $I_b = 1{\cdot}2mA$ $I_{br} = 20mA$

In order that both collector and emitter may act efficiently as collectors during the first phase, it is desirable that a low impedance path be provided at each of these electrodes. Diodes, added as shown in Fig. 3, provide suitable exits.

In the measurement of carrier storage the circuit of Fig. 1 was essentially converted to Fig. 3 during turn off. (The full test circuit is shown in Appendix 2.) A reverse base current of 20mA was employed. An example of the waveforms is given in Fig. 4. The first phase continues to A, and is characterized by a low and almost constant voltage on the base. During this time the full 20mA is accepted by the base. The components of current leaving by the collector and the emitter were measured by the insertion of small resistors (24Ω) in series with the diodes. Their sum is always 20mA during this phase, but their relative instantaneous amplitudes are a function of η, of time, and of the relative external impedances in the emitter and collector exit paths.

Phase 2—Reverse Base Current Transistor Limited

The end of the first phase occurs when the base 'unclamps,' for then the base current is limited by the transistor itself. In the second phase, the elimination of carriers proceeds at the limiting rate for cut-off from the linear region: current is eliminated from the collector and decays with a time-constant associated with T_n and from the emitter with the longer time-constant T_i. The latter component is usually relatively unimportant.

Since phase 2 commences with a rate of elimination equal to I_{br}, and since this diminishes exponentially with a time-constant in which T_n is predominant, the fraction of charge eliminated in this second phase is in the order of

$$Q_2 \sim I_{br}\int_0^\infty e^{-t/T_n}\,dt = I_{br}T_n \quad \ldots\ldots\ldots\ldots \quad (16)$$

Normally the collector current terminates last: it will fall to 10 per cent of I_{br} in about 2·3 time-constants. The second phase may therefore be assumed to be completed in a time

$$t_2 = 2{\cdot}3\ T_n \quad \ldots\ldots\ldots \quad (17)$$

It is shown in Fig. 4 as the time interval from A to B.

A First Order Equation for Turn Off

At the end of phase 2 a majority of the stored charge has been eliminated, and for much circuit work the transistor may be regarded as being in a cut-off state. It is therefore appropriate to assess a turn-off time T_s in terms of these two phases alone.

Since

$$Q_s \sim Q_1 + Q_2$$

and from equation (16) $Q_2 \sim I_{rb}\,T_n$

$$Q_1 \sim Q_s - I_{br}\,T_n$$

By use of equations (15a) and (17), the turn-off time for phases 1 and 2 is given as

$$T_s \sim t_1 + t_2 \sim t_1 + 2{\cdot}3\ T_n$$

Hence

$$T_s = \frac{Q_s - I_{br}\,T_n}{I_{br}} + 2{\cdot}3\ T_n \quad \ldots\ldots\ldots \quad (18)$$

Equation (18) may be expressed entirely in terms of the transistor parameters $\beta_n T_n$ and I_b. According to the principle of controlled saturation, equation (1a) defines an upper bound for Q_s. When this upper bound is substituted in equation (18) a maximum value for T_s will result. Then

$$T_s < T_n\,[\beta_n\,(I_b/I_{br}) + 1{\cdot}3] \quad \ldots\ldots\ldots \quad (19)$$

Equation (19) has been applied to the three test conditions for each sample (Table 2). The calculated values of T_s always slightly exceed the observed turn-off time as they should. As an example, results on sample 203 will be quoted for the following operating condition:

$$I_{e(s)} = 9{\cdot}42\text{mA};\ I_b = 1{\cdot}2\text{mA};\ I_{br} = 20\text{mA}$$

The parameters were $T_n = 4{\cdot}7 \times 10^{-8}$, $\beta_n = 65$ and the calculated value of $T_s = 0{\cdot}25\mu$sec. The calculated time may be compared with Fig. 4 which relates to the same operating conditions.

In practical design the switching resolution may be made equal to T_s, provided the circuit can tolerate the small residual currents which continue in phase 3 (see next section). This can often be arranged. Then equation (19) becomes a key relationship since it determines the turn-off time in terms of the three major design criteria: the maximum permissible emitter current $I_b\beta_n$; the available reverse base current I_{br}; and the normal alpha cut-off. This concept will be employed in Part 2.

Phase 3—Residual Storage

Certain transistors may exhibit a third phase, particularly after saturation. At the end of the first and second phases there may yet remain a collector current, of a few per cent of the original saturated emitter current, which diminishes over a period which may be several times T_s. The analytical treatment of this third phase appears to present great difficulty, and it is not attempted here.

The sample used for illustration in Fig. 4 shows such a phase 3 collector current, in the region to the right of B. All three samples were observed, under each test condition, for the residual current at 0·3μsec after the initiation of turn off. The limit value of all 9 observations is marked in Fig. 4.

If, in practical design, a resolution of T_s is required, the circuit must be such that these small residual currents may be tolerated.

Appendix 1

The Calculation of Minority Carrier Storage

When a transistor operates in a linear manner, the collector current which flows can be expressed in terms of a minority carrier gradient which exists in the base region. Since the presence of such a gradient implies stored charge, it follows that any conducting transistor exhibits storage. The storage can be computed from a knowledge of transistor geometry, of diffusion constant, and of base width.

It is usual to employ a simplified geometry in order to ease the mathematical problems, hence this analysis will be based on a one-dimensional model[3]. Furthermore, when the lifetime of carriers in the base region is much longer than the average carrier transit time, the gradient is linear. It will be assumed further that minority carrier lifetimes are negligible in the emitter and collector regions, so that no storage can exist therein. The assumptions of concern to carrier lifetimes are well approximated by modern alloy type transistors: experience shows that the geometrical simplification yields results of acceptable accuracy.

For a pnp transistor the storage is in the form of holes. For non-saturated operation, conditions are as illustrated in Fig. A1, curve (1). When the required integration has been performed the storage is given as

$$Q_n = K\, I_{cr}\, T_n \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

In this expression Q_n represents charge storage associated with a collector current I_{cr}. The diffusion constant and base width have been absorbed into $K\,T_n$ where T_n is the inverse of the alpha cut-off frequency, expressed in radians/sec, and K is a constant.

If the transistor is inverted, so that the electrode previously employed as the collector is now used as the emitter, the carrier population arranges itself as shown by curve (2) of Fig. A1. There is now a corresponding equation to (1) which is

$$Q_i = K_1\, I_{er}\, T_i \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where T_i is given by use of the alpha cut-off for inverted connexion.

The constants K and K_1 of equations (1) and (2) have values near unity. They are treated as being equal in the analysis which follows, and are evaluated as unity in the concluding stages.

It has been shown by Ebers and Moll[4,5] that when the transistor is in a saturated state both forward and inverse conduction exists. Then, both the collector and the emitter may be considered to inject and collect minority carriers simultaneously. The resultant storage Q_s is shown by curve (3) and may be obtained by linear superposition of curves (1) and (2). Thus:

$$Q_s = Q_n + Q_i$$

and from equations (1) and (2)

$$Q_s = K\,(I_{cr}\,T_n + I_{er}\,T_i \quad \ldots\ldots\ldots\ldots \quad (3)$$

The storage can be evaluated from equation (3) if the terms I_{cr} and I_{er} can be expressed in terms of the terminal currents I_e, I_c and I_b. This is done in the equations which follow, where the sign conventions are those shown in Fig. A2.

$$\left.\begin{aligned} I_e &= I_{ef} - I_{er} \\ I_c &= I_{cf} - I_{cr} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots \quad (4a)$$

$$\left.\begin{aligned} I_{cr} &= \alpha_n\, I_{ef} \\ I_{er} &= \alpha_i\, I_{cf} \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots \quad (4b)$$

where α_n and α_i are the forward and inverse values of α. From equations (4a) and (4b)

$$I_{er} = \alpha_i \frac{I_c + \alpha_n I_e}{1 - \alpha_n \alpha_i} \quad \ldots\ldots\ldots\ldots\ldots \quad (5a)$$

$$\text{and } I_{cr} = \alpha_n \frac{I_e + \alpha_i I_c}{1 - \alpha_n \alpha_i} \quad \ldots\ldots\ldots\ldots\ldots \quad (5b)$$

By substitution of equations (5a), (5b) into equation (3)

$$Q_s = K \frac{T_n\alpha_n (I_e + \alpha_i I_c) + T_i\alpha_i (I_c + \alpha_n I_e)}{1 - \alpha_n\alpha_i} \quad \ldots\ldots \quad (6)$$

In the body of the article, storage is required in terms of emitter and base currents, rather than emitter and collector currents as in equation (6). The substitution

$$I_b = I_e + I_c$$

in equation (6) yields the required form:

$$Q_s = K \frac{I_e\,[T_n\,\alpha_n\,(1 - \alpha_i) - T_i\,\alpha_i\,(1 - \alpha_n)] + I_b\alpha_i\,(T_n\,\alpha_n + T_i)}{1 - \alpha_n\alpha_i} \quad \ldots\ldots\ldots\ldots \quad (7)$$

The expression, equation (7), may be given greater physical significance if it is expressed in terms of grounded emitter gains, β_n and β_i instead of the grounded base gains,

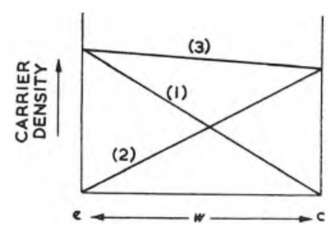

Fig. A1. Carrier gradient in the base of a transistor

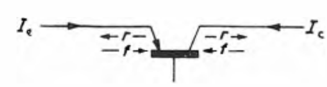

Fig. A2. The transistor terminal currents and the forward components (f) and reverse components (r)

α_n and α_i. This may be done as follows:

Divide both numerator and denominator by $(1 - \alpha_i)$ so that equation (7) becomes,

$$Q_s/K = \frac{I_e\,T_n\,\alpha_n\left(1 - \dfrac{T_i\alpha_i\,(1 - \alpha_n)}{T_n\alpha_n(1 - \alpha_i)}\right) + I_b\,\beta_i\,(T_n\,\alpha_n + T_i)}{1 + \beta_i\,(1 - \alpha_n)}$$

Then divide throughout by α_n,

$$= \frac{I_e\,T_n\left(1 - \dfrac{\beta_i T_i}{\beta_n T_n}\right) + I_b\,\beta_i\,(T_n + (T_i/\alpha_n))}{(1/\alpha_n) + (\beta_i/\beta_n)}$$

$$= \frac{I_e\,\beta_n\,T_n\left(1 - \dfrac{\beta_i T_i}{\beta_n T_n}\right)}{(\beta_n/\alpha_n) + \beta_i} + \frac{I_b\,\beta_n\,T_n\left(1 + \dfrac{T_i}{T_n\alpha_n}\right)}{(\beta_n/\beta_i\alpha_n) + 1} \quad \ldots\ldots\ldots\ldots \quad (7a)$$

where, as is usual, $\alpha_n \to 1$

$$Q_s \sim K\,\beta_n\,T_n\left[I_b\frac{1 + (T_i/T_n)}{1 + (\beta_n/\beta_i)} + I_e\frac{1 - (\beta_i T_i/\beta_n T_n)}{\beta_n + \beta_i}\right] \quad (7b)$$

Equations (7) may be used to study the charge storage as I_e is varied, with base current as a fixed parameter. Thus, from equation (7b),

$$1/K\ \frac{dQ_s}{dI_e} = \frac{\beta_n T_n}{\beta_n + \beta_i}\left(1 - \frac{\beta_i T_i}{\beta_n T_n}\right) \quad \ldots\ldots \quad (8)$$

and the storage will diminish as I_e is diminished if

$$\frac{\beta_i T_i}{\beta_n T_n} < 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

Notes on the Measurement of Carrier Storage

Determination of the Parameters and Calculation of Storage

According to the theoretical model used for Appendix 1, the four parameters α_n, α_i, T_n and T_i are required at zero collector voltage. The alphas should also be d.c. values. The V_{co} curves must be plotted as a function of current since the parameters are current dependent.

The derivation of the V_{co} curves is shown in Fig. A3, where β is plotted, rather than α. First the β values are plotted against V_c with I_e* as a parameter [(a) in the Figure]. The zero voltage intercepts then allow the curves (c) to be drawn for $V_c = 0$. Similarly the α cut-off curves are plotted, this time against $\sqrt{V_c}$ [curve (b)]. Theory shows that for small collector voltages such a plot will give a straight line. Extrapolation of the curves to zero collector voltage gives the required α cut-off curves (d).

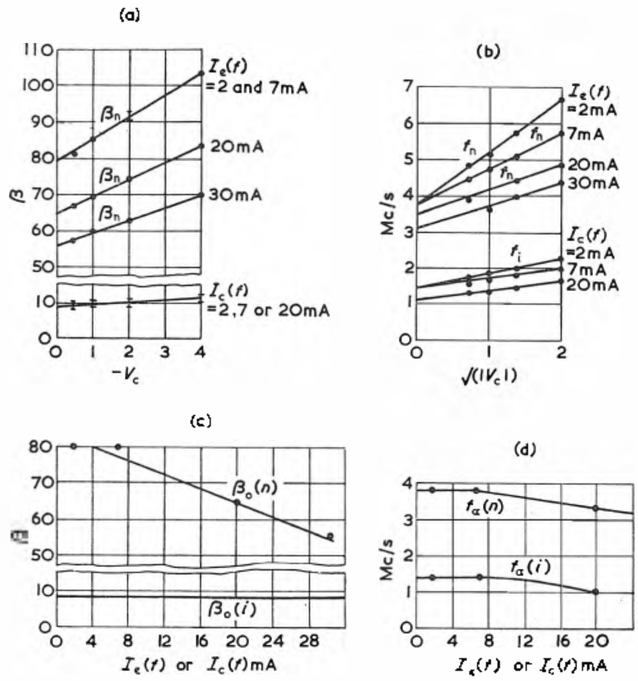

Fig. A3. Parameters of GE 2N123 sample 203 for saturated operation

(a) D.C. β as a function of I_e and V_c
(b) α cut off as a function of I_e and V_c
(c) D.C. β at $V_c = 0$ as a function of I_e (derived from (a))
(d) α cut off at $V_c = 0$ as a function of I_e (derived from (b))

The carrier storage in saturation is calculated by means of equation (6) Appendix 1, which requires a knowledge of I_e and I_c, together with the four parameters plotted. The particular value of each parameter must agree with the operating conditions, and the relationship is first established by successive approximation in equations (5a) and (5b).

Many measurements are involved in the calculation of charge storage, and the manner in which small errors combine in equation (6) makes it difficult to calculate the storage with precision. It is for this reason that the results of such calculations shown in Table 2 are quoted with an accuracy of ±20 per cent.

* $I_e = I_e(f)$ for a non-saturated transistor in normal connexion, and $I_e = I_c(f)$ when the transistor is inverted.

Circuit used to measure Storage

The circuit shown in Fig. A4 was use to study storage under the three defined test conditions described in Part I. Because of the small times involved in phase 1 of turn off, a rapid application of the reverse current was required. A Western Electric mercury wetted relay, mounted in a coaxial system, is known[6] to give a switching time of less than 10^{-9}sec. The circuit was developed to make use of such a switch. Currents were measured by means of a wide band oscillograph.

In order to maintain the resolution offered by the switch, oscillograph cables were correctly terminated at both ends. The diodes D_1 to D_4 are so placed in the circuit that transient effects in them do not disturb the measurements.

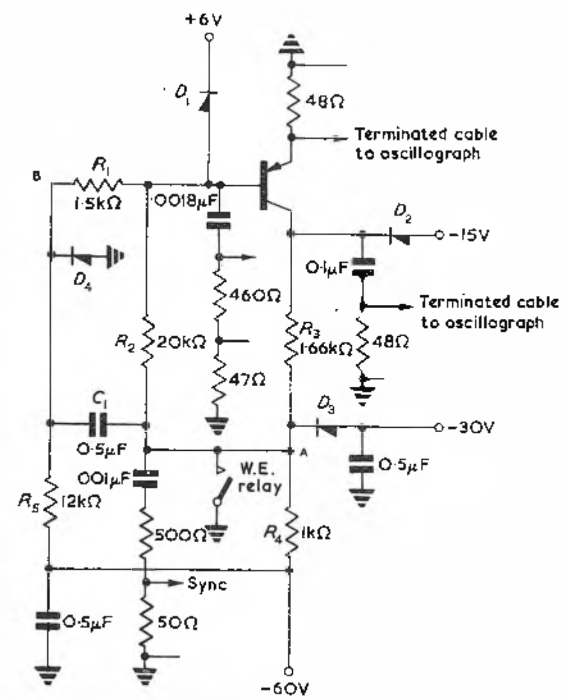

Fig. A4. Circuit for studying carrier clearing

Resistors R_1 R_2 and R_3 are adjusted to set the three test conditions. The values shown give the first condition, namely I_e = 18mA, I_b = 1·12mA, I_{br} = 20mA.

All voltages applied to the circuit were sufficiently large to allow currents to be treated (to a first approximation) as though generated by constant current sources.

The static conditions are set as follows (W.E. Relay open): —

A −60V supply holds diode D_3 clamped at all times irrespective of load. The point A is thus maintained at sensibly constant voltage of −30V, which determines the transistor collector current in terms of R_3, and the base current in terms of R_2. These two resistors, therefore, suffice to determine the first order static conditions.

A resistor R_1, also connected to the base, has its remote end maintained at ground potential by diode D_4, which is held clamped by current in R_5. The small current in R_1 must be included in the base current and it has been treated as a correction factor.

Closure of the relay grounds point A, and raises point B to +30V by means of the coupling capacitor C_1. Resistor R_1 now injects a reverse base current of 20mA into the transistor. At this time the original base current has been stopped since the voltage across R_2 is close to zero, and the current in R_3 has ceased for the same reason.

Current leaving the collector and emitter now flows into the terminated cables until the transistor cuts off. When cut off is complete diodes D_1 and D_2 conduct. These diodes are added merely to safeguard the transistor from destructive voltages.

It was required to study both the instantaneous emitter and collector current values, and their integrals over the whole turn-off operation. For this reason the circuit is arranged to give the currents directly. The current waveforms were photographed (Fig. 4, Part 1 is an example) and then graphical integration was employed to determine the net storage (Table 2, Part 1).

The size of the photographs and the nature of the waveforms made accurate integration difficult. For this reason an accuracy of ±20 per cent is believed realistic for the charge storage measurements.

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A 16mm Telerecorder

In recent years telerecording has advanced from a position of relative obscurity in the sphere of production technique to one of considerable importance. Nowadays, no television production centre is complete without equipment which will record complete programmes with consistently good quality results, and is simple in operation.

The latest product of High-Definition Films Ltd, Telerecorder type HD.800, is designed to fulfil these requirements. The equipment is self-contained, requiring only a composite video signal, a sound signal and 200-240V single-phase a.c. mains. Vision and sound are recorded simultaneously, the vision on 16mm double or single-perforated film, and sound on 16mm single-perforated magnetic film, the capacity of each being 2 500ft giving a total continuous running time of just over one hour. Provision is made for recording any C.C.I.R. or O.I.R. standard signal, and the equipment is suitable for one man operation.

All the electronic units are constructed on standard 19in rack mounting panels, which are arranged either in enclosed cubicles or on the control console. Access to the cubicles is provided by full length swing doors, front and rear, and the two large cubicles are fitted with extractor fans. The necessary test equipment is provided as an integral part of the design. This comprises a large-screen double-beam oscilloscope, a grey scale generator and a brightness measuring meter. A switching panel, in conjunction with the oscilloscope, enables the operator to examine the video signal at various stages in the unit, and also to check the locally-generated pulses.

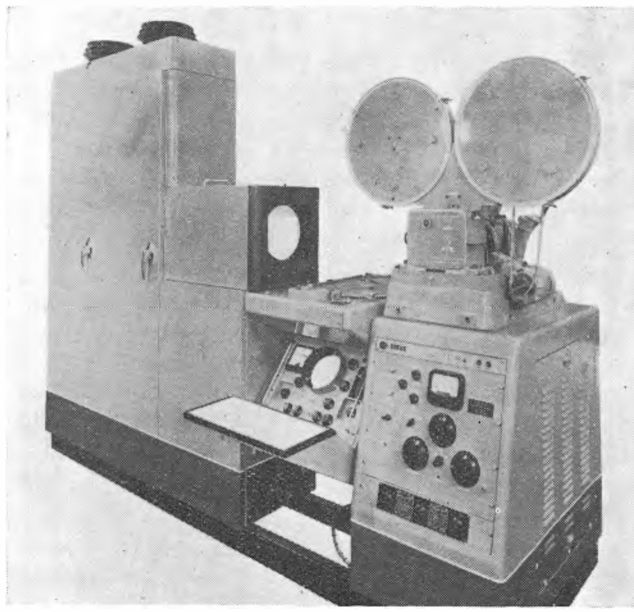
The displayed picture is focused on to the film which runs in a camera having a pull-down period of 15°. On 50c/s mains operation, this represents a time of 1·66msec, hence the film is stationary, and recording picture information for 38·34msec. Therefore, virtually a complete interlaced raster is recorded on each frame of film. It is necessary that the camera and the displayed picture should be in phase, and this is provided for by rotating the camera motor stator by the desired amount with a hand control. Visual indication of the correct condition is given on the displayed picture in the form of a thin horizontal bright line which can be moved until it is just out of the recorded area. A phase locking system can be supplied as an optional extra enabling recordings to be made of television pictures having a frame frequency different from that of the a.c. supply operating the recorder.

The camera is equipped with trip switches which ensure that the motor will not operate unless the film threading and loop tension is correct, the film take-up motor on, and the door closed. Film counters are provided operating in feet or metres as required.

The flexibility of the electronic circuits enables recordings to be made in three modes:

(1) Exposure of film to a positive displayed picture; development of the film yielding a negative, and printing of this negative to produce a positive image.
(2) Exposure of film to a negative displayed picture; development of the film yielding a positive image.
(3) Exposure of reversal film to a positive displayed picture; development of the film yielding a positive image.

A video high-frequency boost unit is provided, giving a variable boost from 0 to 15dB at any fixed frequency in the range 1 to 10Mc/s. This uses a transmission line as a circuit element, and the circuit arrangement is such that



The telerecording equipment

phase distortion is inherently zero. The line structure of the displayed picture is eliminated by use of spot wobble.

A flywheel synchronizing unit can be supplied for the regeneration of horizontal synchronizing pulses when using signals emanating from a distant source. Space is available in the standard equipment for this and for the phase-locking unit already mentioned.

The sound recorder is similar in specification with those used in broadcasting, all the facilities associated with such machines being provided. Either fully-coated or 0·1in striped stock may be used, the location of the track being in accordance with the relevant S.M.P.T.E. Standard. Exact synchronization between sound recorder and camera, including run-up and run-down, is achieved by driving the sound recorder from the camera using a Selsyn motor link.

The sound recorder can also be driven separately by its own synchronous motor, or by its own Selsyn motor controlled by another Selsyn motor connected to other equipment. Hence a local telecine machine to which a suitable Selsyn motor has been attached will run in synchronism with the recorder, and for re-transmission and monitoring purposes it is unnecessary to produce a married print, or run double-headed. Switching is provided to convert the sound recorder to run in this condition.

A Cine-theodolite Control System Used on Guided Missile Ranges

By R. J. Garvey*, B.Sc.

Cine-theodolites are used to determine the trajectory and velocity of experimental guided missiles; a number of them being dispersed on the range and operated by a central controller. This controller operates the cine-theodolite shutters and triggers flash lamps which expose the theodolite bearing and elevation readings on the cine film.

The lamps are flashed synchronously; the interval between flashes being 200msec ±100μsec. Shutter operations are synchronized to within ±1msec. These limits correspond to ±1ft in determining the missile trajectory and ±0·1 per cent of the velocity.

CINE-THEODOLITES are used to determine the trajectory and velocity of experimental guided missiles. A number of them are dispersed on the test range and controlled over land lines from a common station. The cine-theodolite, made by the Askanion Co. and shown by Fig. 1, consists essentially of

(a) A theodolite which can be trained in bearing and elevation by an operator tracking a missile through a sighting telescope or binoculars;

(b) A cine-camera, normally operated at the rate of 5 frames per second.

A precise record of the orientation of the theodolite axis is exposed on each frame of the cine film by flashing small gas discharge lamps near the bearing and elevation scales; the scale images being projected on to the film by a simple optical system (this is not shown). A shutter exposes an image of the missile on each frame with respect to the theodolite axis so that corrections can be made for tracking error when analysing the film. This tracking error is the angular displacement between the theodolite axis and the true line of sight to the missile. A typical film record is illustrated by Fig. 2; the cross hairs correspond to the theodolite axis. The scale shown can be read to within ±10sec of arc.

Simultaneous bearing and elevation readings of two cine-theodolites, placed at the extremities of a known base line, give the position of the missile in space at that instant and since successive readings are obtained at known time intervals, both the trajectory and velocity of the missile can be determined. The cine-theodolites have a limited range, it is therefore normal to employ a maximum of six to cover the whole trajectory. Readings are then taken from those cine-theodolites most suitably disposed to the part of the missile trajectory being plotted.

The function of the control system is to operate the cine-theodolite flash lamps and shutters. Ideally the lamps of each cine-theodolite should be triggered synchronously and at a regular rate. Asynchronization between the scale exposures introduces an error in determining the trajectory and hence the velocity, while an error in the repetition rate has a direct effect on the velocity evaluation. The shutter exposures, showing the position of the missile referred to the theodolite axis, should also be made at the same time as the lamps are flashed. This involves synchronizing the electro-mechanical operation of the shutters with the triggering of the gas discharge lamps. Tracking error can be changing rapidly from frame to frame as shown on Fig. 2; asynchronization between the operation of the shutter and scale lamps therefore gives an inaccurate recording of this error so that false corrections are made when analysing the film.

* *The Royal Aircraft Establishment.*

Cine-Theodolite Camera Mechanism

The cine-theodolite has a shutter mechanism as shown by Fig. 3. Shutter vanes are turned through 180° for each exposure by a gear sector engaged with pinions on each

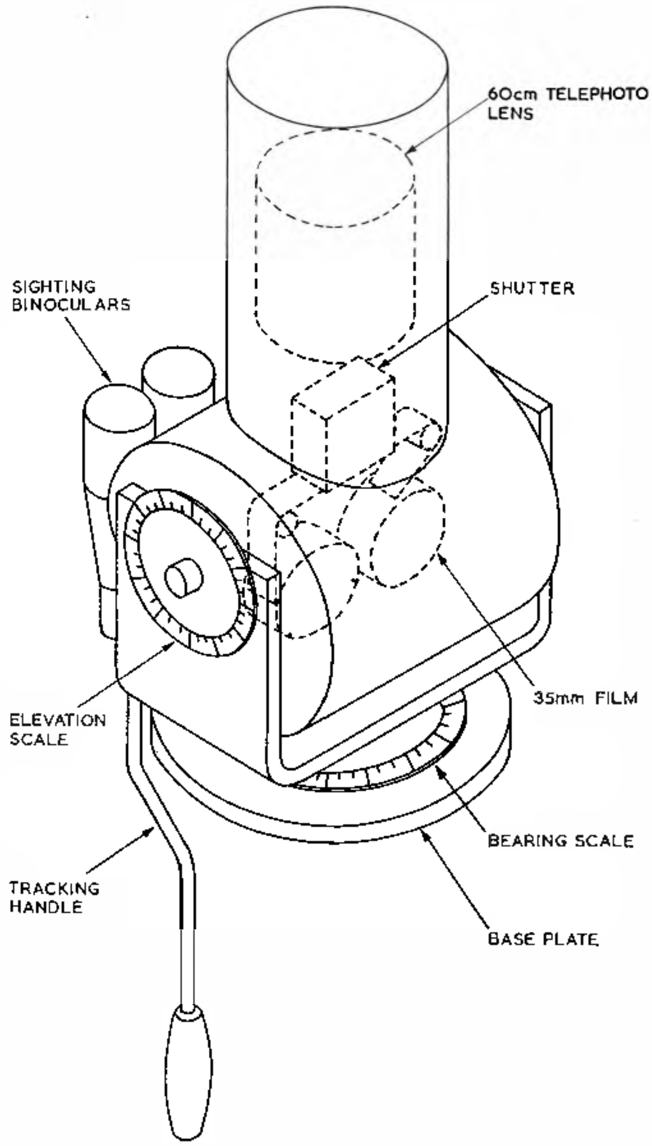


Fig. 1. General arrangement of cine-theodolite

vane. An electromagnet triggers the mechanism by releasing the gear sector so that it is pulled to the left or right by a spring. The wheel to which this spring is attached is

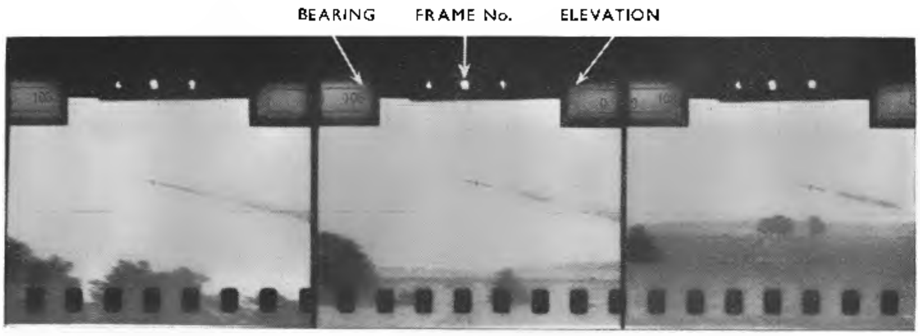


Fig. 2. Typical cine-theodolite film

rotated through 180° after each exposure, thus cocking the mechanism for the next operation. An electric motor automatically engages the wheel, cocks the shutter and winds the film on one frame each time the shutter is triggered. A rotary contact fitted to one of the vanes is made when the shutter is fully open; this provides a monitoring signal for synchronizing the operation of the shutter with that of the scale flash lamps.

The time taken for the shutter to operate depends on ambient temperature, local voltage and the mechanical adjustment of the mechanism. Effects of voltage and temperature on a particular mechanism are shown in Table 1; the time quoted is that for the shutter vanes to reach the fully open position after switching a voltage to the release magnet.

TABLE 1

TEMPERATURE (°C.)	OPERATING VOLTAGE (V)	OPERATING TIME (Msec)
−20		42·5
−10		40
0	12	38
20		36
40		34
	8	37
	12	35
	16	33
	20	31

The exposure time for each operation of the shutter is normally about 10msec.

Flash Unit

The flash unit operates two gas discharge lamps fitted near the bearing and elevation scales of the cine-theodolite. These give a high intensity flash of short duration so that a record of bearing and elevation is exposed on each frame of the cine-theodolite film. A short duration flash of about 100μsec is necessary to avoid blurred exposures of the scale readings at high tracking rates.

The circuit diagram of the flash unit is shown by Fig. 4. It consists of a power pack which supplies the necessary high voltage and a thyratron which acts as a switch in series with the flash lamps. Each input pulse fires the thyratron so that a 1μF capacitor is discharged through the lamps and the thyratron extinguished. A thermal delay switch protects the thyratron when the unit is first switched on by allowing the cathode to attain its working temperature before the h.t. is switched to the anode. The thyratron is hydrogen filled and does not require a bias voltage. The flash lamps are filled with xenon, and were specially made for the cine-theodolites.

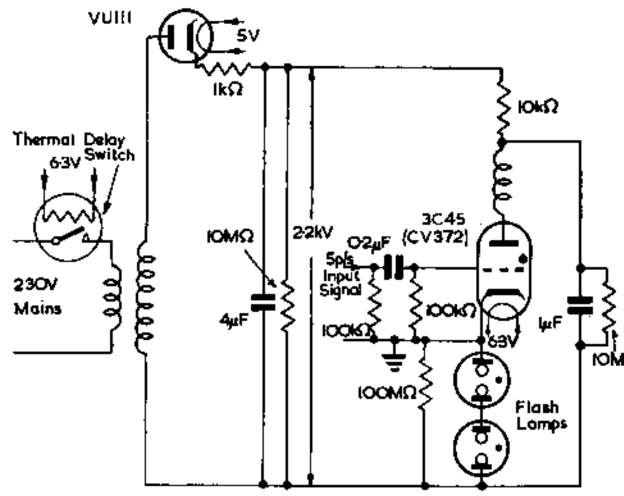
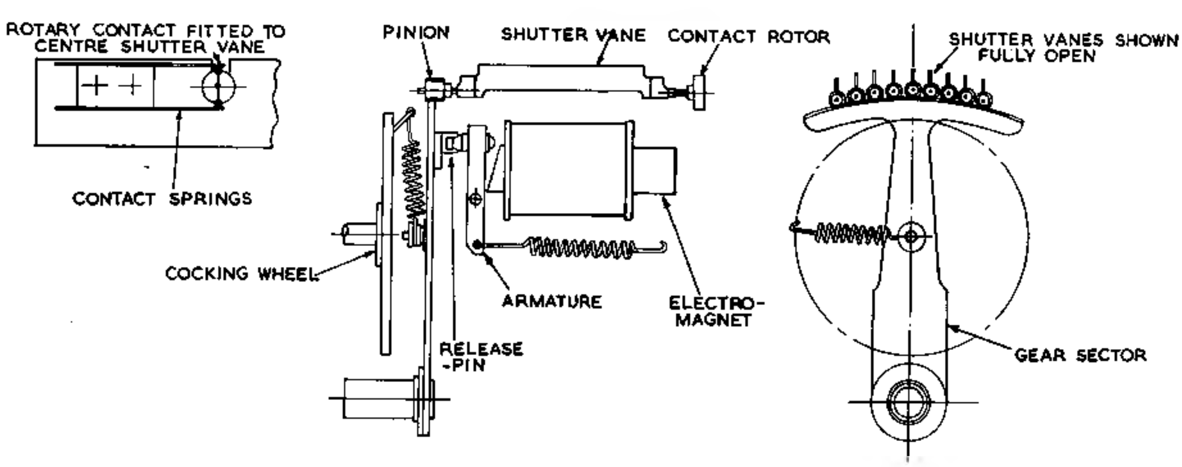


Fig. 4. Circuit of flash unit

General Description of Control System

A 5p/s operating signal is supplied to the control system by the range central timer[1]. This consists of a 10kc/s oscillator and frequency dividing circuits which supply timing and reference signals to the various measuring and recording instruments on the range; it functions as a separate

Fig. 3. Cine-theodolite shutter mechanism



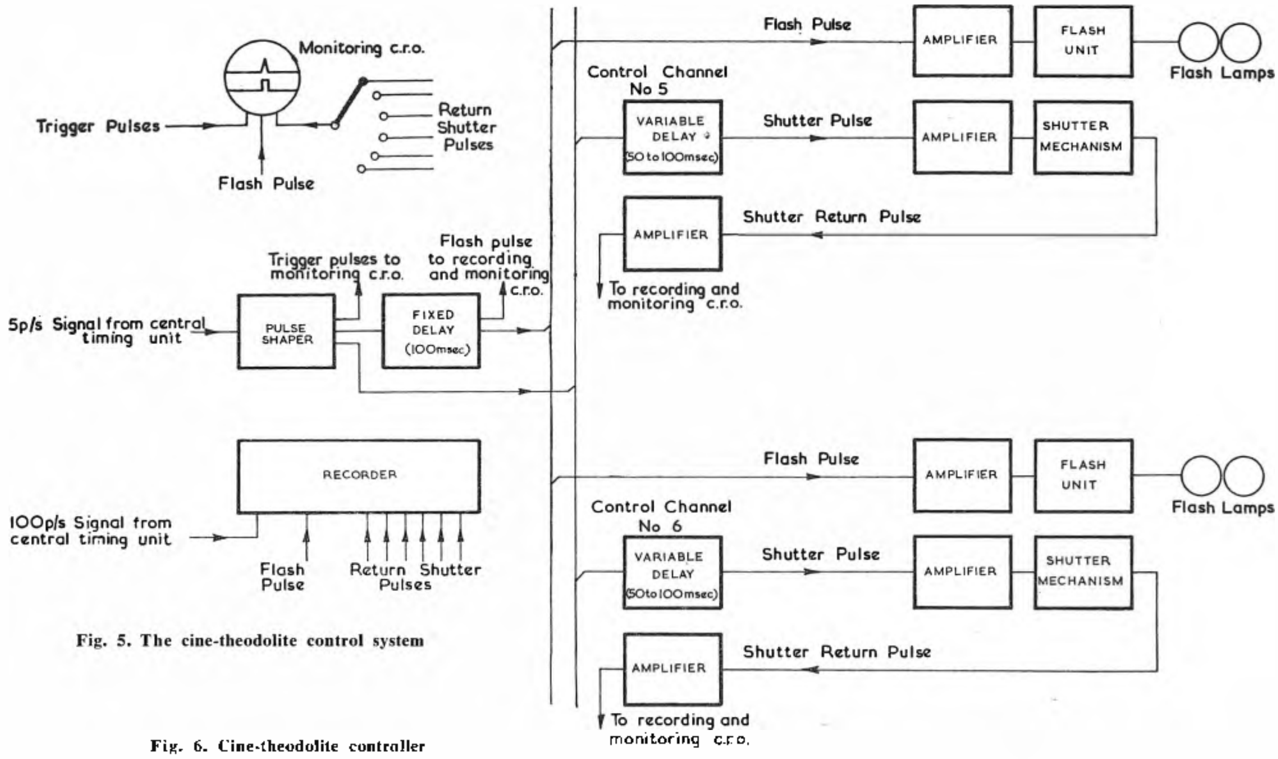
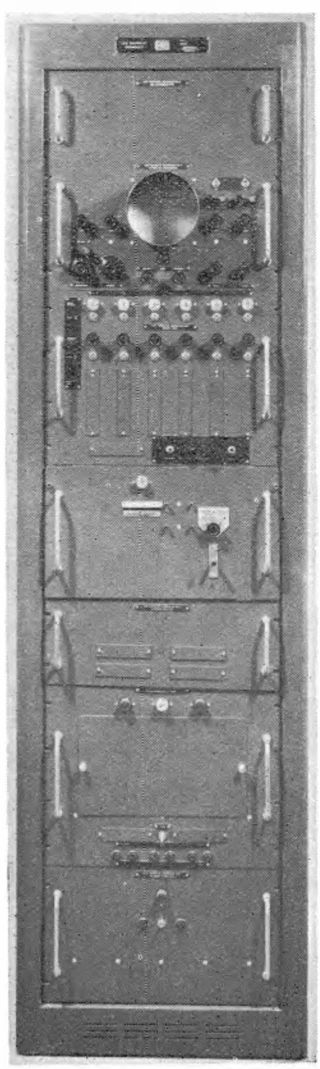


Fig. 5. The cine-theodolite control system

Fig. 6. Cine-theodolite controller



equipment. The control system is triggered from the range timer rather than with a separate oscillator so that cine-theodolite data can be related in time to that obtained by other range instruments.

Referring to the schematic diagram of Fig. 5: the 5p/s signal from the timer goes via separate variable delay circuits to each cine-theodolite shutter mechanism and via a common delay circuit to all the remote flash units. The variable delay circuits are adjusted so that the output pulse is sufficiently in advance of the common flash pulse for the shutter vanes to be passing through their mid-position as the scale lamps flash. The rotary contacts fitted to the shutters are closed when the vanes are in the mid-position and synchronization is effected by adjusting the variable delays until the shutter return pulses provided by these contacts are each in phase with the common flash pulse as seen on a double beam c.r.o. The time-base of the c.r.o. is triggered by pulses from the pulse shaper which are 100msec in advance of those displayed. Shutter return pulses are also recorded together with the flash pulse on a multi-channel recorder; this gives an overall check on the operation of the system. Pulses are transmitted to the remote stations via underground telephone cables where they are amplified and shaped as necessary by local amplifiers.

The pulse shaper, delay circuits, oscilloscope and recorder are assembled on a 6ft rack as shown by Fig. 6.

In operating the system each cine-theodolite is synchronized as necessary; the 5p/s signal from the range timer is then switched off in preparation for the actual run and the particular cine-theodolite and flash units that are to be used selected by closing appropriate switches on the central control panel. The run commences when the 5p/s signal is gated from the range timer at zero time; that is when the missile is launched. To conserve film each cine-theodolite is switched off locally when tracking has ceased, while the controller is switched off automatically at the end of the run.

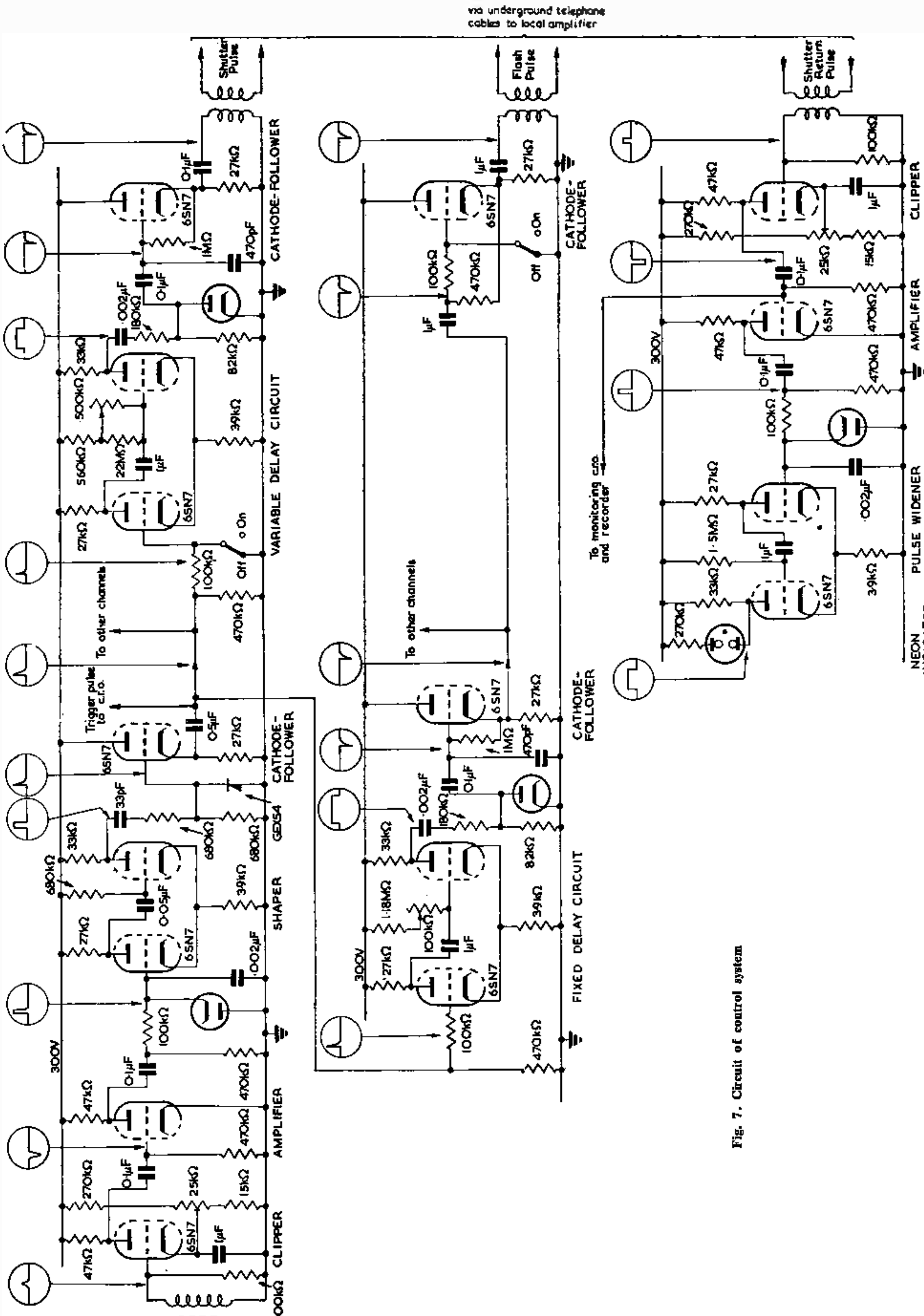


Fig. 7. Circuit of control system

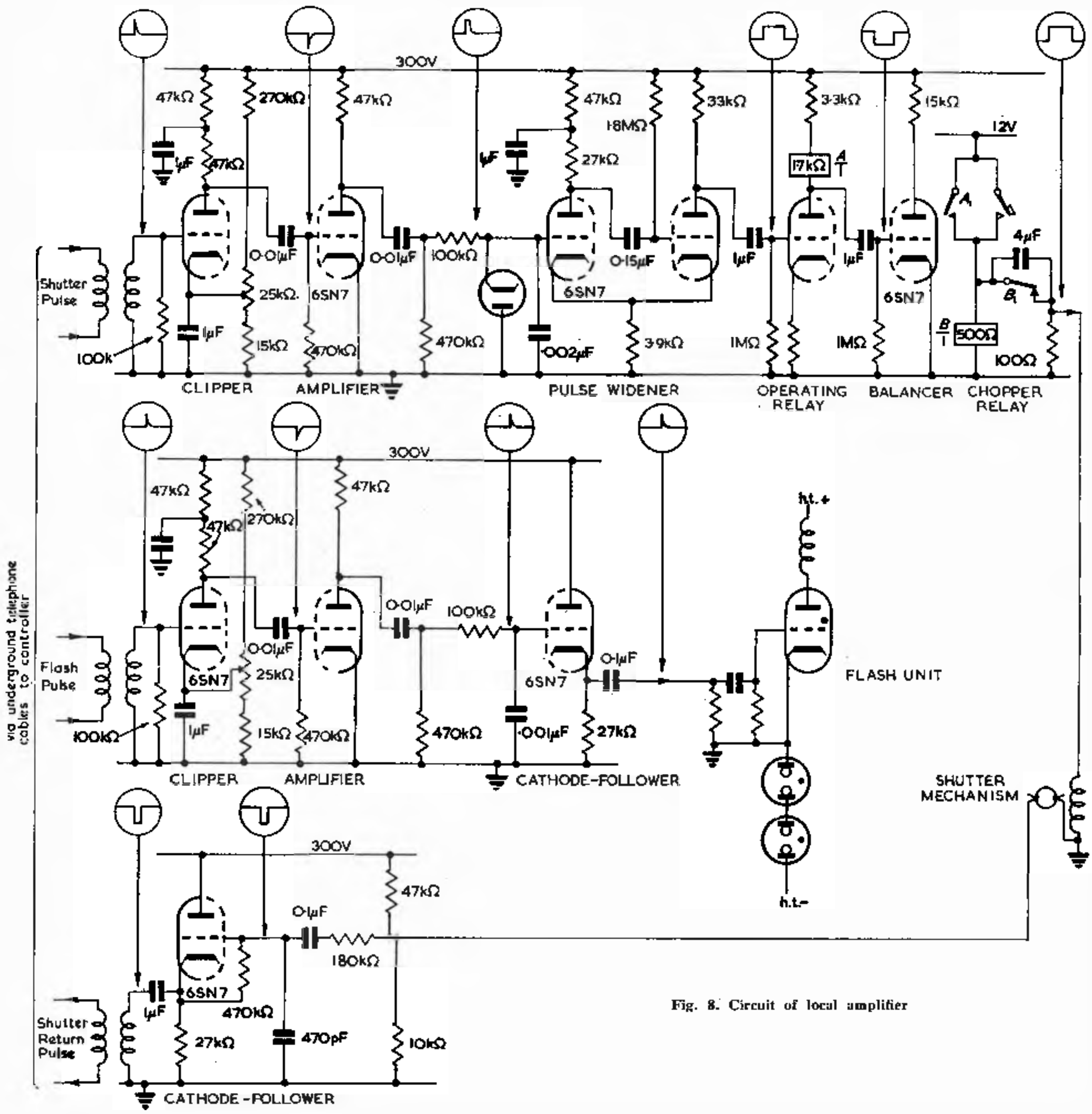


Fig. 8. Circuit of local amplifier

Circuit Details

Details of the control circuit are shown by Fig. 7. The 5p/s signal from the range timer goes to a pulse shaper; this consists of a triode amplifier biased beyond cut-off to clip any low amplitude noise, followed by another amplifier stage and a flip-flop circuit. The flip-flop acts as a squarer; the square wave output is differentiated and the negative pulses removed with a diode. A crystal diode is used here to avoid the introduction of 50c/s ripple that would result from the heater-cathode leakage of a thermionic valve. The diode connected to the input of the flip-flop removes any negative component from the input signal while the ·002μF capacitor short-circuits any high frequency noise picked up in the preceding stages of the shaper circuit.

The shaped pulses go via a cathode-follower to a common fixed delay circuit and to variable delay circuits in each of the six output channels of the controller. The delay circuits consist, in each case, of a flip-flop; the output square waves of these are differentiated and the positive pulse removed with a diode. Output pulses are thus derived from the trailing edge of the flip-flop square waves and are displaced from the input pulses by a time delay corresponding to the flip-flop action. Time delays are varied by altering the value of the resistance connected to the grid of the second valve of the flip-flop. A preset resistor permits the fixed delay to be adjusted to 100msec; the other delays can be varied from 50 to 100msec.

The output of each variable delay circuit triggers the cine-theodolite shutters; the pulses going to each remote station via underground cables and a local amplifier. The output of the fixed delay circuit goes to a cathode-follower in each control channel and then via the underground cables and local amplifier to the remote flash units. Flash pulses from each control channel are in phase, while the shutter pulses can be varied to be from 0 to 50msec in advance of the common flash pulses according to the delay of each shutter mechanism. The rotary contacts on each shutter transmit return pulses via the underground cable to an amplifier in each control channel. This consists of a triode amplifier biased beyond cut-off to clip any low amplitude noise followed by another amplifier stage and pulse widening flip-flop. The pulse widener operates a neon lamp which gives a useful indication by blacking out when the cine-theodolite shutter is operating. The shutter return pulse also goes to the recorder and via a selector switch to the monitoring oscilloscope.

The delay time of the flip-flop circuits is relatively independent of ambient temperature changes or of the steady h.t. voltage and any drift due to valve ageing can be taken up in the preset or variable resistance controls. Apart from slow changes, however, it is necessary that a given time delay be repeated for successive operations to within $\pm 100\mu$sec. It is, therefore, necessary to trigger the flip-flops with fast pulses and to ensure that there is no 50c/s or transient voltage superimposed on the exponential voltage rise at the grid of the second valve. The flip-flops are accordingly triggered via the pulse shaper described and supplied with a well smoothed and stabilized h.t.

Local Amplifier

This unit is installed local to the remote cine-theodolite and amplifies the pulses from the controller. The circuit is detailed by Fig. 8.

Shutter pulses go to a triode amplifier, biased to clip any low amplitude noise, a second amplifier and a pulse widening flip-flop. The flip-flop widens the pulse so that it is of sufficient duration to operate the shutter solenoid. A diode connected to the input of the flip-flop removes any negative component from the signal while the $\cdot 002\mu$F capacitor short-circuits any high frequency noise picked up in the cable or preceding amplifier stages. The output of the pulse widener goes to a triode with a Siemens' high speed relay in the anode circuit; this relay operates the shutter solenoid. Relay *B* is a Post Office type 3 000 relay which has an operating time of about 40msec; its function is to interrupt the inductive solenoid circuit before *A* releases and so protect the contacts of the high speed relay. The random variations in the operating time of a type 3 000 relay prohibit it being used to switch the shutter solenoid directly. The balancer is switched in anti-phase with the relay valve to equalize the load and thus prevent fluctuations in the h.t. voltage. A push switch shown in the relay circuit is used for manual operation of the shutter mechanism.

Flash pulses go via a clipper and amplifier to a cathode-follower which feeds the signal to the thyratron in the flash unit. The capacitor connected to the grid of the cathode-follower short-circuits any high frequency noise picked up in the preceding circuits and cable.

The rotary contact fitted to the shutter vanes earths the grid of a cathode follower so that pulses are transmitted via the telephone cable back to the controller.

Line Installation

Operating pulses are transmitted from the central controller to the remote cine-theodolites and flash units via underground telephone cables. These are 20lb, multi-core, lead sheathed cables; each pair of cores having the following characteristics:

Attenuation:	1 to 2·5dB/mile at frequencies of 1 to 10kc/s.
Characteristic Impedance:	450 to 170Ω at frequencies of 1 to 10kc/s.
Propogation Delay:	10μsec/mile.

The longest cable run on the range is about 4 miles and the corresponding delay of 40μsec has no appreciable effect on the synchronous operation of the cine-theodolite shutters and flash lamps.

The multi-core cables carry speech signals and other transmissions in addition to the cine-theodolite pulses. All signals are carried over twisted pairs of lines which are terminated in transformers so that signals and lines are balanced about earth potential. This eliminates 'cross-talk' between the various transmissions; the attenuation between any two signals is greater than 50dB/mile of cable.

Recorder

The common flash pulse, shutter return signals and a 100p/s time trace are recorded to give an overall check on the synchronization and operation of the system. Records are made by feeding the pulses to a row of fixed electrodes or pens which burn marks on a moving strip of carbon backed paper. The pens and a roller, which supports the paper and contacts the carbon backing, are rhodium plated to withstand mechanical wear and provide low contact resistance.

The paper is traversed by a motor and gear box driving a roller, spring loaded on to the paper. Recording can be made at speeds of 10 or 25in/sec.

Pulses to be recorded are amplified by the circuit of Fig. 9. Recorder pens are connected in the anode circuit of the second amplifier stage; a 100kΩ resistor in the anode circuit prevents the valve anode floating when the pen circuit is interrupted. The roller which supports the pens

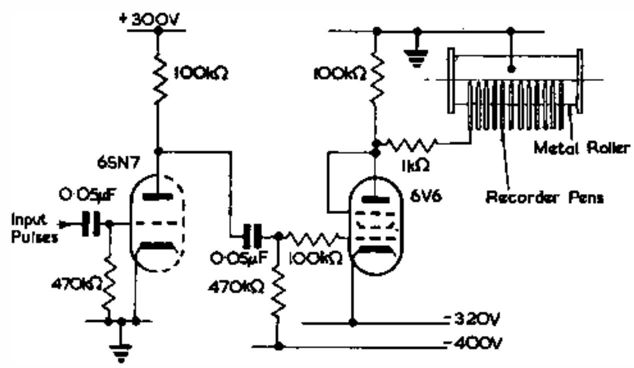


Fig. 9. Circuit recorder amplifier

and paper is not insulated from the recorder chassis so that it is necessary to earth the positive h.t. supplied to the second amplifier stage.

The 100p/s signal recorded is gated from the range timer at zero time; cine-theodolite operations can thus be related to a common time scale and hence to data obtained by other range instruments.

Accuracy

The accuracy to which the missile trajectory and velocity are determined depends on the magnitude of the timing errors involved in the operation of the cine-theodolite flash lamps and shutters. Control pulses are locked to the crystal oscillator of the range timer which has a frequency that is true and stable to within a few parts in 100 000. Apart from this precision, irregularities in the nominal 200msec interval between pulses will be introduced by jitter in the delay circuits; tests show this to be not greater than $\pm 100\mu$sec. Transmission of the flash pulses along various lengths of cable results in phase displacements between them; the maximum phase shift is 40μsec with the present installation. Accuracy in synchronizing the shutter operations depends upon the degree to which asynchronization can be observed on the monitoring oscilloscope and on the jitter or random time variation inherent in the design of the shutter mechanism. Tests show that the shutters can be synchronized to within ± 1msec. Shutter exposures give a correction for tracking error so that shutter asynchronization has only a second order effect in the evaluation.

As a summary of the timing errors it can be stated that:

(a) Asynchronization in operating the flash lamps is not greater than 40μsec.

(b) The interval between successive operations of the flash lamps is within $\pm$200μsec of the nominal value.

(c) Shutters are synchronized with the flash lamps to within $\pm$1msec.

The effect of timing errors in determining the trajectory and velocity depends on the disposition of the cine-theodolites and the skill of the operators in tracking the missile. It is estimated that with the present installation the figures quoted correspond to an error in determining the trajectory of not more than $\pm$1ft in space; this discrepancy is comparable with those due to errors in the surveying of the cine-theodolite and the orientation of the theodolite axis. Velocity evaluations depend upon the accuracy with which the trajectory fixes are made; errors introduced by irregular operation of the flash lamps, however, will be not greater than $\pm$0·1 per cent.

Acknowledgments

A prototype control system was designed and developed in the Guided Weapons Department of the Royal Aircraft Establishment and subsequently engineered by Cinema Television Ltd.

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A Simple Three-Channel C.R.O. Beam Switch

By W. F. Lovering*, M.Sc., A.M.I.E.E., and
M. P. Hearn*, B.Sc., A.C.G.I.

It is often desirable to display several different phenomena simultaneously on a single beam c.r.t. In this article details of a three-channel beam switch are given using a three-phase multivibrator circuit which utilizes only six valves.

SEVERAL different phenomena may be displayed simultaneously upon a single-beam cathode-ray oscilloscope by the addition of an electronic switch arranged to connect each of the phenomena in turn.

The basic circuit of such a switch is illustrated in Fig. 1. Several (n) gating valves share a common anode load, but are normally biased beyond cut-off. A positive pulse is applied to each valve in turn and causes it to conduct and amplify for the duration of the pulse. The switching waveforms each have a duration of $1/n$ period positive and $(n-1)n$ period negative and are displaced in time by $1/n$ period.

The switching frequency may be lower, or higher than the signal frequency. If the switching frequency is lower than the signal frequency, but exceeds 25c/s, each signal completes several cycles while displayed, and persistence of vision enables all traces to be seen simultaneously. The spot flies from one trace to another in a negligible fraction of the display time and each trace appears continuous and sharply focused against a dark background, (assuming that the tube controls have been adjusted to give a sharply defined spot). The phenomena displayed are displaced in time by several cycles of the signal and the method is suitable only for the examination of repeated waveforms or of transients which persist for many switching periods.

If the switching frequency is much greater than the signal frequency, each signal is sampled several times per cycle. A switching frequency corresponding to a harmonic of the signal frequency produces a dotted line trace; with non-harmonic or very high frequency switching the trace appears continuous. This method introduces no flicker and may be used for visual or for photographic examination of repeated waveforms, or for photographing single transients

High-frequency switching is more generally useful than low-frequency switching, but the limit of performance is determined by the inevitable stray capacitances in the circuit. Because of such capacitances the transition from one trace to another is not instantaneous and the screen is brightened by the spot in transit. If the transition occupies an appreciable fraction of the switching period the background illumination is noticeable.

* *Imperial College.*

A further effect of stray capacitance is due to capacitance between the switching grid and the output anode. The sharply rising switching voltage causes a current flow in the capacitance which results in an anode voltage waveform with a spike (overshoot) at the leading edge. The resultant signal trace on the screen is blurred. A similar effect results from feedback through stray capacitance between anode and signal grid of the gating valve. The

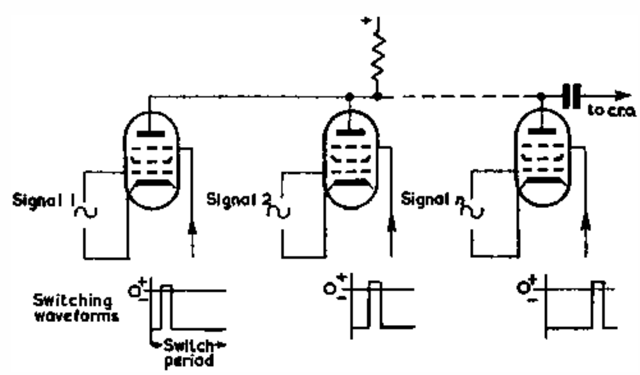

Fig. 1. Operation of a multi-channel switch

best available gating valves are pentodes with 'straight' $V_g - I_a$ characteristics using either control grid or suppressor grid, and with a high mutual conductance from either grid. Unfortunately such valves have no screen between anode and suppressor and so have an appreciable capacitance between these electrodes. With the pentodes available satisfactory switching with a sharp trace and little background illumination, is possible with a switching frequency of 13kc/s or a little higher. Double-triode gating valves have been used, but the stray capacitances are greater than with pentodes and the upper frequency limit of satisfactory operation is about 8kc/s.

The number of traces which may be displayed is limited in practice by the cost and complexity of the switch. For example a design based upon the use of ring-connected, Eccles-Jordan trigger circuits requires twelve valves and associated resistors and capacitors to switch three channels.

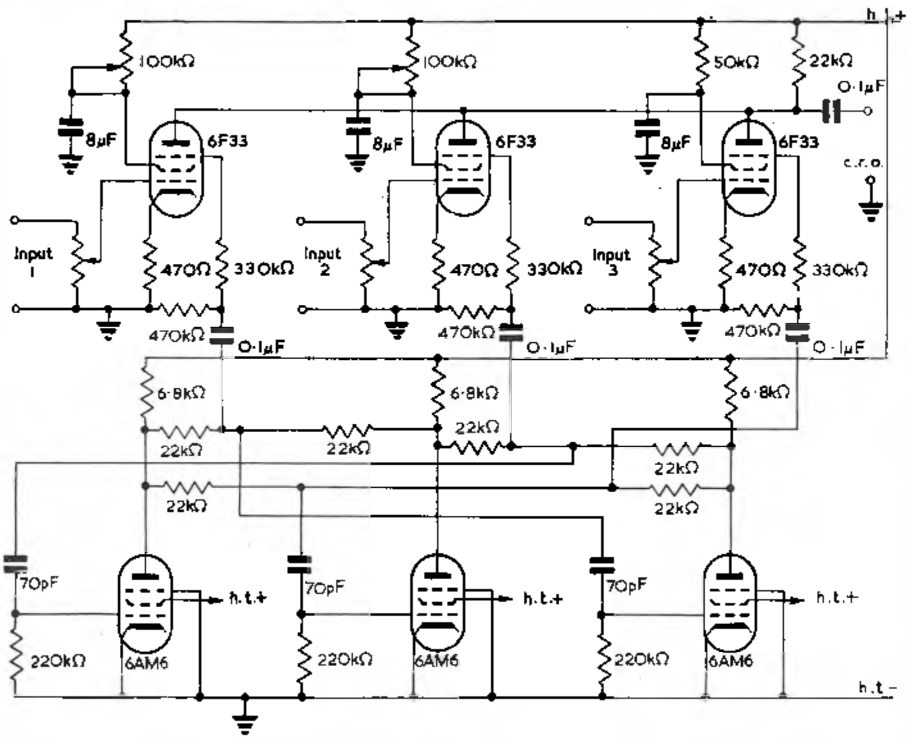

Fig. 2. Circuit of a three-channel switch

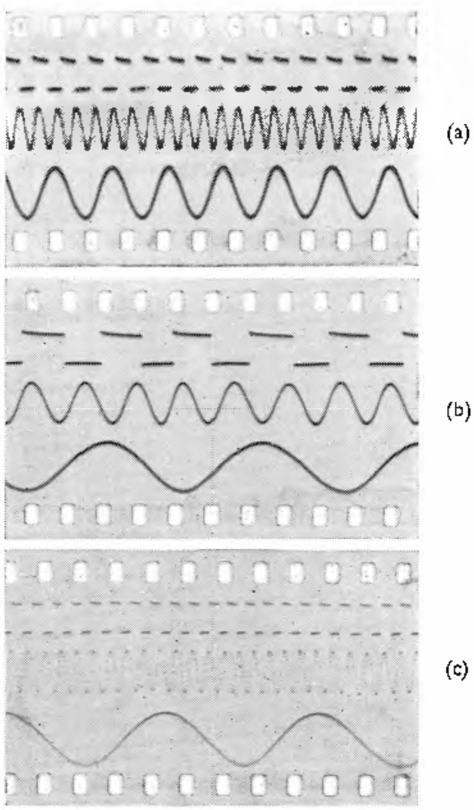

Fig. 3. Photographs of typical traces

(a) and (b) a 100c/s square wave, a 150c/s sine wave and a 50c/s wave (c) shows a 400c/s square wave, a 500c/s sine wave a a 50c/s sine wave

A three-channel switch using only six valves may be constructed if a three-phase multivibrator is used to generate the voltage waveforms required to operate three gating valves. The utility of a c.r.o. may thus be considerably increased at little cost.

Three-channel switching, using pentode valves, requires three rectangular switching waveforms, each having a positive duration of one-third of a cycle and each displaced in time from the others by one-third of a cycle. Such waveforms may conveniently be generated by a three-phase multivibrator.

Three-phase multivibrators have have been described elsewhere[1]; two modes of operation are possible. In one mode each valve conducts for two-thirds of a period and the anode voltage waveforms each have a positive duration of one-third of a cycle. In the other mode which is more readily generated, each valve conducts for a third of a period and the anode voltage waveforms have a positive duration of two-thirds of a cycle. With the latter mode of operation the required switching waveform is obtained from points mid-way in potential between two anodes.

The rise and fall times of the positive pulse are determined mainly by the anode load resistance and stray capacitances in the multivibrator circuit. Using miniature r.f. pentodes a transition time of less than 0·5μsec is possible so that a good waveform may be obtained at any frequency below about 50kc/s. The same multivibrator may be used for either high-frequency or low-frequency switching merely by a change of grid coupling capacitances.

The circuit of a complete three-channel switch using gating pentodes is shown in Fig. 2. With the circuit values shown the switching frequency is 13kc/s and it is possible to examine phenomena with any frequency up to 2kc/s. Photographs of typical traces taken with a moving film, are shown in Fig. 3. The photograph of Fig. 3(c) was taken with the maximum film speed; lack of detail in the 500c/s trace is due mainly to the inadequate film speed and lack of sensitivity of the film; the dotted nature of the trace contributes little to the lack of detail.

The traces may be separated upon the screen by arranging that the mean anode current of each gating valve is different from that of the others; this is achieved by control of screen voltage.

If high-frequency switching is used the output from the gating valves contains high harmonics of the switching frequency. The signal input to the gating valves contains signal frequencies only. The bandwidth of any amplifier following the switch must, therefore, be much greater than the bandwidth of any amplifier preceding it.

Low frequency switching introduces negligible high frequency components and the bandwidth required is little greater than that required for the signals.

An inadequate bandwidth following the switch results in an increase in the time of transition from one trace to another. The spot approaches its new position relatively slowly and the resulting trace is blurred.

The three-beam switch is a useful addition to any single-beam oscilloscope. With its aid the transient currents in each phase of a three-phase motor may be examined, or two phenomena such as voltage and current may be examined and the third trace used for a calibrating waveform.

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Transistor Circuits for use with Gas-Filled Multi-Cathode Counter Valves

By J. B. Warman*, A.M.I.E.E., and D. M. Bibb*

A technique is described which enables complex digital circuits using both transistors and Dekatrons to operate from a low-voltage power supply. A transistor d.c. convertor is used to generate the 475V h.t. supply. Output pulses from the Dekatron cathodes drive transistor circuits. A transistor blocking oscillator feeds stepping pulses to Dekatron guide cathodes and a similar circuit is used for resetting the Dekatron to its home cathode.

In this way a set of compatible circuit elements is obtained for designing logical circuits. A telephone exchange register constructed on these principles is described.

TRANSISTORS are being increasingly used for control applications due to their excellent switching properties, low power consumption, inherent reliability and general convenience in mounting. The switching speeds obtainable with transistors are high when compared with those of cold-cathode valves and relays, this fact often being used to overcome or justify the increased cost of such circuits. There are, however, some slow-speed counter applications in control circuits where it is not economic to provide transistor counters with much higher resolution rates than are required. For slow-speed digital counting (e.g. up to 3 000p/s), multi-cathode gas-filled counting valves such as the Dekatron[1] are ideal, being comparatively cheap and giving a convenient visual display. It may not be immediately obvious that Dekatrons can be blended with transistor circuits without the immediate economy being outweighed by the inconvenience of the additional range of power supplies necessary for the Dekatrons and associated driving circuits. It is the purpose of this article to show how this can be conveniently arranged.

Transistor oscillators have been used as d.c. convertors[2,3,4,5,6] by using one d.c. voltage to drive the oscillator and transforming and rectifying its output to provide another d.c. voltage. Such a convertor can be used to derive the 475V anode supply voltage required by the Dekatrons from a lower voltage supply used by the transistor circuits.

The supply voltage for the transistor circuits can be the same as the guide-bias voltage used by the Dekatrons. The output current from the Dekatron cathodes is about 0·5mA and so is adequate to drive transistor logical circuits. If one can also arrange for the logical circuits to step Dekatrons by means of transistors, then Dekatron counters can be incorporated into transistor digital circuits without additional power supplies. The use of a transistor circuit for driving a Dekatron has been proposed by Chaplin and Sharpe[7]. The circuits described in this article use transistor blocking oscillators for stepping the Dekatrons. A blocking oscillator is also used for resetting the Dekatron counters. In this way, a set of compatible circuit elements has been obtained which enables complex logical circuits to be designed using both transistors and Dekatrons operating from low-voltage supplies. A telephone exchange register constructed on these principles is described later. A pulse-train generator which was described previously[8] using Dekatrons and thermionic valves has also been redesigned to use the transistor circuits.

Transistor Dekatron Drive Circuit

Fig. 1 shows a simple blocking oscillator circuit which can be triggered by a tertiary winding on its transformer: a further winding transforms the signal so generated to a voltage suitable to step the Dekatron. The transistor X_1 is coupled regeneratively from the collector to the base by winding W_1 and W_2 as shown. A small positive bias derived from the potential divider R_1 and R_2 is fed via winding W_1 to bias off transistor X_1 and so prevent continuous oscillation. The collector of X_1 is fed from a -12V supply and clamped via rectifier MR_1 to a -20V supply. When the circuit is triggered by a pulse to winding W_4, then transistor X_1 is regeneratively driven hard into conduction and bottoms at earth potential until the core is saturated, whereupon the transistor is cut off regeneratively by the reverse

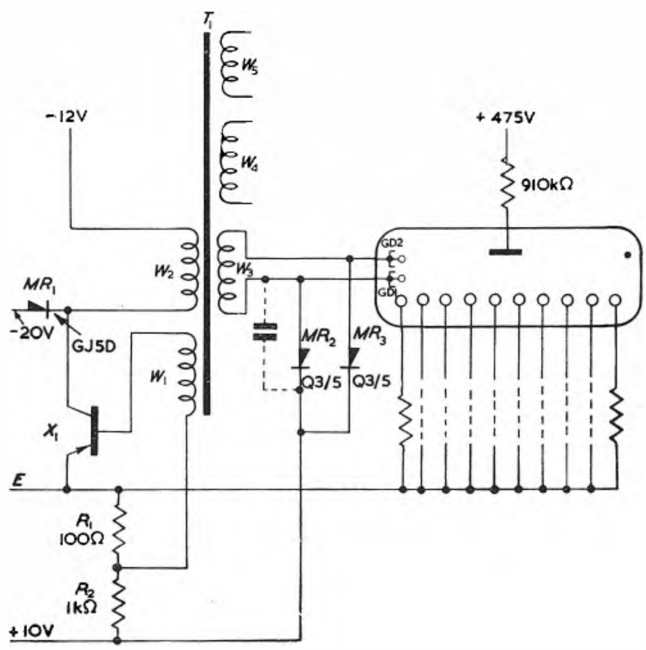

Fig. 1. Transistor Dekatron drive circuit

e.m.f. as the flux change collapses. The collector voltage tends to become very negative due to the back e.m.f. of the coil but is clamped by rectifier MR_1 to -20V. On collapse of the back e.m.f., the collector potential returns to normal awaiting the application of further stepping pulses. The collector potential therefore swings from -12V to earth and then to -20V before returning to -12V as shown in curve (b) of Fig. 2.

The Dekatron guides are normally returned to the guide bias potential of $+10$V via rectifiers MR_2 and MR_3. The signal from the collector of transistor X_1 is transformed into winding W_3 where, due to the action of rectifiers MR_2 and MR_3, the positive-going collector excursion is fed to guide 2 as a negative-going signal and the negative-going collector excursion is fed to guide 1 as a negative-going signal.

* *Siemen's Edison Swan Research Laboratory.*

It will be seen from curves (c) and (d) of Fig. 2 that the action of the blocking oscillator has generated and fed to the Dekatron guides two adjacent negative-going square pulses ideally suited to stepping the Dekatron. Using selenium rectifiers for MR_2 and MR_3, it has not proved to be necessary to delay the fall of the first pulse to overlap the second since the stray capacitance of the selenium rectifier MR_2 seems to suffice for this. When driving the high speed Dekatrons (type GS10D) at speeds of the order of 10kc/s it is usually necessary to use germanium point-contact rectifiers and in these circumstances it is desirable to insert a small capacitor across MR_2 to give this pulse lap-over. Details of a typical transformer suitable for this drive unit are given in the Appendix; the arrangement is in no way critical and the choice of bobbin and laminations quoted was primarily dictated by their immediate availability. Windings W_1 and W_2 are wound 1:1 and W_3 gives a step up from W_2 in order to feed 100V pulses to the Dekatron guides.

It has been found that the Dekatrons can be driven with reliability at higher stepping speeds than those quoted by the manufacturer for the recommended driving circuits.

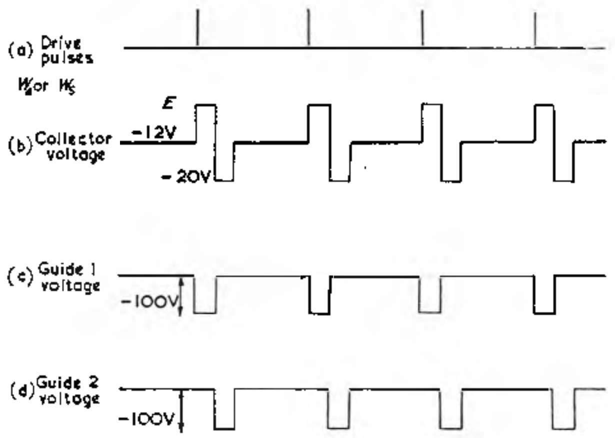

Fig. 2. Idealized pulse waveforms of transistor Dekatron drive circuit

This is probably due to the more ideal nature of the square pulses derived in this way and to the low value of cathode resistance used by the Dekatron when driving transistor logical elements. For instance it proved to be possible to run Dekatrons type GS10C up to 5kc/s with apparent reliability.

Generation of Anode Supply for Several Dekatrons

Fig. 3 shows a d.c. convertor suitable for supplying h.t. to 10 Dekatrons. It comprises a push-pull oscillator using two transistors coupled via a transformer with laminations of 'HCR' metal, the characteristics of which determine the frequency of the oscillation. When a constant voltage is applied across an inductor the current rises linearly and the time taken to saturate the core is $N\Phi_s/V$.

Where Φ_s is the flux change required to saturate the core.

V is the applied voltage.

N is the number of turns on the winding.

A secondary winding on the core will feed current only until the core is saturated, after which no significant further increase of flux can occur. Two transistors X_1 and X_2 are connected to the output transformer T_1 which provides positive feedback to their emitters. While X_1 is conducting, the voltage across half the transformer primary is V volts. This causes current to flow via the feedback windings to the emitter of X_1 to hold it fully conducting and to the emitter of X_2 cutting it off. When the core becomes saturated, X_1 is no longer held in conduction and the flux in the core drops to its remanent value.

In so doing it induces a voltage to cut off X_1 and bring into conduction X_2 which is held conducting until the core reaches saturation in the opposite sense, whereupon the conduction of the transistors is again reversed.

The core is taken into saturation twice per cycle and the period of the oscillation is $4N\Phi_s/V$ since the flux undergoes an excursion of four times the saturation figure in each cycle. The frequency of oscillation of a practical oscillator is thus $f = V/4\pi\Phi_s N.A.$

Where f is the frequency (c/s).

A is the core area (cm^2).

Φ_s is the flux density (lines/cm^2).

One advantage of using square-loop material (HCR) for the transformer is the rapid changeover between the alternate bottoming of the two transistors, made possible by the sharp saturation of this material which reduces the dissipation in the transistors. The HCR material also has a very low coercivity and a relatively high saturation flux

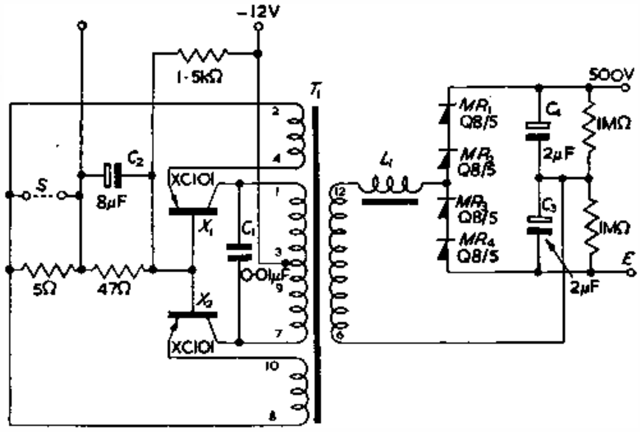

Fig. 3. Transistor d.c. convertor

Output (a) 500V at 3mA; (b) 500V at 5mA (with strap "S" inserted)

density, this combination allows an efficient transformer to be designed with small physical dimensions, as has been described by Royer[2] and others.

The small choke L_1 is necessary momentarily to reduce the load of the uncharged smoothing capacitor on the T_1 transformer, which might prevent the oscillator from starting. The capacitor C_1 absorbs transient high voltages which appear at the collectors due to the leakage inductance of the transformer and might damage the transistors. To reduce the number of turns on the secondary winding a voltage-doubler rectifier circuit has been used; the full-wave rectification draws a balanced load from X_1 and X_2 (a double winding feed to the rectifiers requires four times the secondary turns and a full-wave rectifier bridge requires twice the number of turns and more rectifiers).

The circuit was designed to serve up to 10 Dekatrons with +475V h.t. at 300μA apiece, giving a total of 3mA. It has shown itself capable of providing 5mA if the strap S is inserted as shown in Fig. 3 to reduce the common emitter load.

A Prototype Telephone Exchange Register

As an exercise in the use of these techniques a rudimentary telephone exchange register has been constructed with plug-in units.

The problem is to accept and store the dialled impulse trains from the subscriber's dial so that they can be subsequently used to control the setting up of the call through

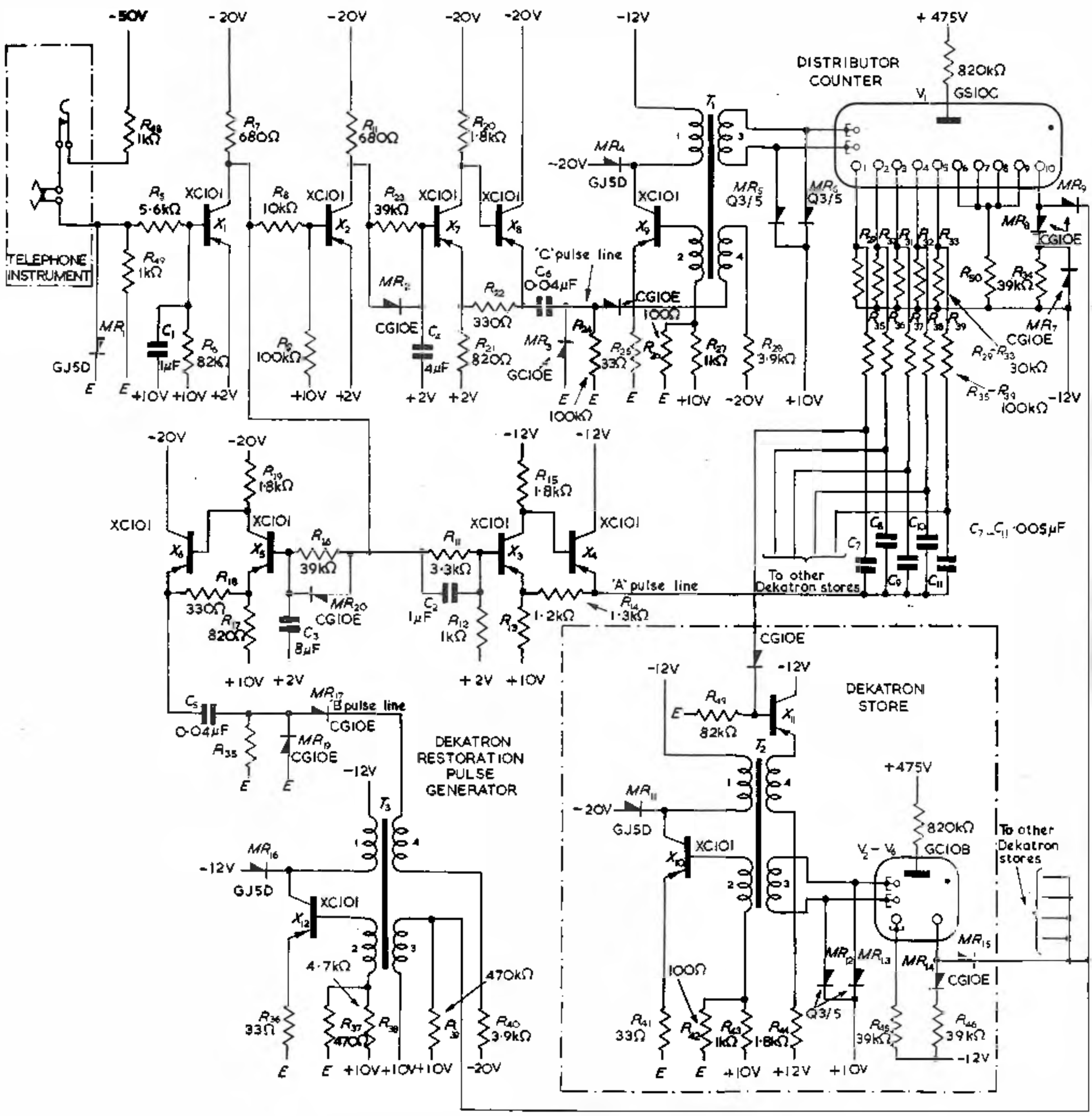

Fig. 4. Rudimentary prototype telephone exchange register

the telephone area. A subscriber's telephone instrument is connected into the exchange by a single pair of wires over which all the signals must be passed and the power fed to the telephone instrument. Since the two legs of the line must not be unbalanced during signalling, due to the possibility of cross-talk, the only signal available is the looping and the disconnexion of the line corresponding to lifting and replacing the receiver. The dial signals the digits by interrupting the loop a significant number of times at 33 per cent make and 66 per cent break ratio. The release condition is indicated when the receiver is returned to the hook, permanently disconnecting the pair. The dialling breaks are not confused with the release condition because they do not persist for long enough to initiate the release action. In order to separate the dialled trains, a further timing action is necessary to identify the persistence of the make condition beyond that normal between impulses of the same train.

Transistors X_1 and X_8 receive the dialled impulses and time the make and break periods in order to derive the 'end of train' and the 'end of call' discriminations. On lifting the receiver, a loop is presented to resistors R_{48} and R_{49} which represent the 50V battery feed normally supplied to the telephone instrument. This causes transistor X_1 to come into conduction and cut off transistors X_2, X_3, X_5 and X_8 which bring transistors X_7, X_6, into conduction. The changeover action here is positive and fast because these transistors are arranged in Schmitt trigger pairs. The positive-going excursion at the emitter of the X_8 transistor is fed to step the distributor Dekatron V_1 via its transistor drive circuit (which is of the type described previously), from the normal cathode 10 to the first cathode to prime the pulse-plus-bias gate[9] C_7R_{35} of the first Dekatron store V_2.

When the subscriber dials the first train of digits, transistor X_1 responds to the disconnexions of the loop and is in turn followed by the Schmitt trigger X_3, X_4 which feeds pulses to step the first Dekatron store V_2, since only the gate C_6R_{35} is opened by the distributor Dekatron V_1. The capacitor C_3, being charged positively by the MR_{20} rectifier during the make period of the dial impulse, is unable to discharge during the disconnexion via resistor R_{16} and so holds the Schmitt trigger X_5, X_6 from turning over during the break dial periods. This restrains the release signal. The Schmitt trigger X_7, X_8 turns over so that X_8 conducts during the first loop disconnexion (due to the capacitor C_4 charging rapidly via rectifier MR_2) and remains in this state

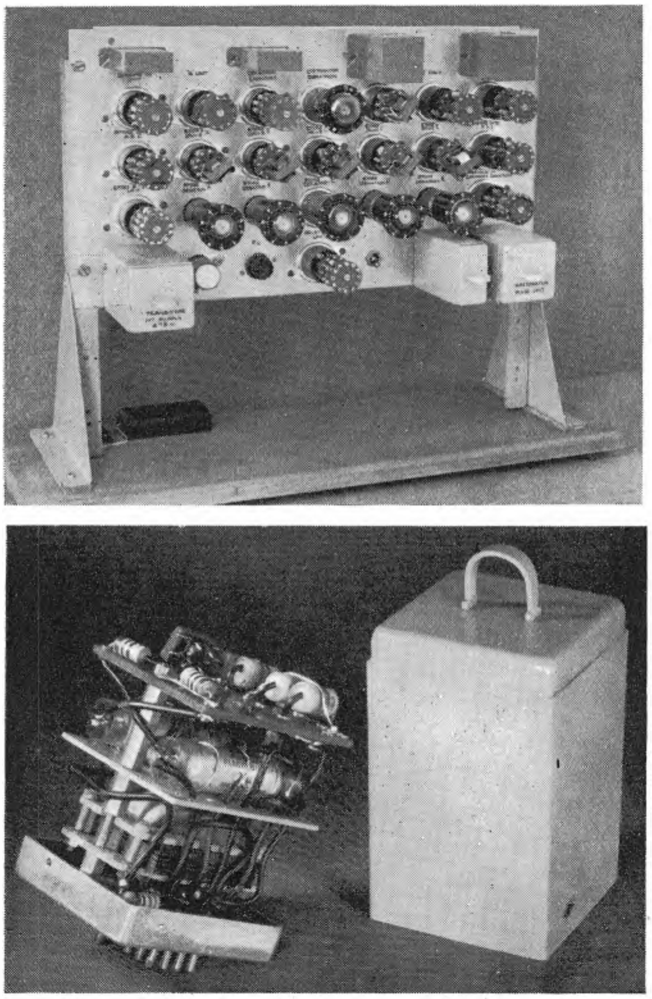

Fig. 5. (a) Prototype telephone exchange register (b) D.C. convertor

until the dial train ceases (due to the inability of the capacitor C_4 to discharge during the make period between dial impulses of the same train). Upon return of the subscriber's dial to normal, capacitor C_4 discharges and allows the Schmitt trigger X_7, X_8 to turn back thus delivering a pulse to step the distributor Dekatron V_1 to cathode 2.

The next train of impulses will be counted and stored on the next storage Dekatron whose pulse-plus bias gate is now opened by the new positioning of the distributor Dekatron V_1. This action proceeds until all the digits have been dialled and have been distributed by V_1 to their appropriate storage Dekatron. If the subscriber replaces the telephone handset during or after dialling, then the permanent disconnexion of the line allows C_3 to discharge and the transistor trigger X_5, X_6 turns over. The change of potential so generated is fed to the blocking oscillator circuit X_{12}, T_3 which generates a single -140V pulse which is fed to the home cathodes of the Dekatron and restores them to normal. The arrangement of rectifiers shown on the home cathode of the distributor Dekatron V_1 for feeding and isolating the reset pulse is necessary only when the restoring pulse is connected to a working outlet where a -140V pulse could cause damage or faulty circuit action. This isolation is not really necessary here but is added to demonstrate the technique. Rectifier MR_9 isolates the cathode from the pulse restoration common lead because it is back biased by the -20V from winding 3 of T_3. MR_8 isolates the output and MR_7 clamps the cathode load against any excessive negative excursion.

Test Results

The timing of the 'B' pause was measured to be 200msec and that of the 'C' pause 100msec.

The reception of impulses was satisfactory over a 2kΩ loop with 20kΩ line leak at speeds in excess of 40 i.p.s.

Conclusions

The technique described enables complex logical circuits to be designed using both transistors and Dekatrons operating from low voltage supplies. A telephone exchange register using the technique has operated reliably for a year. A pulse-train generator has also been designed using the same circuit elements.

Acknowledgments

Acknowledgment is made to the Director of the Siemens Edison Swan Research Laboratory for permission to publish this article. The authors are indebted to Dr. J. E. Flood for advice on preparing the article and to Mr. B. D. Simmons for advice on the d.c. convertor.

APPENDIX

WINDING DETAILS

Transistor Drive Circuit

Laminations Mumetal 4 mil thick
Service type 521 M and EA type 224

Windings	W_1	60 turns of	39 s.w.g.
	W_2	60 ,,	36 s.w.g.
	W_3	700 ,,	46 s.w.g.
	W_4	160 ,,	32 s.w.g.
	W_5	160 ,,	32 s.w.g.

D.C. Converter Output Transformer

Laminations H.C.R. Alloy 4 mil thick
Services type 522 M and EA type 225

Windings

Wdg.	Terml. Start	Direc. tion	Terml. Finish	Turns	Material	Size R
A_1	1	→	3	170	M Lewmex	31 s.w.g.
A_2	7	→	9	170	,,	,,
B_1	4	→	2	34	,,	35 s.w.g.
B_2	10	→	8	34	,,	,,
C	6	-→	12	3600	,,	41 s.w.g.

D.C. Convertor L_1 Choke

Core Mullard 1—FX1052 1—FX1053
Winding L_1 500 turns of 42 s.w.g. M Lewmex

Dekatron Restoration Oscillator Transformer

Core 2—FX1619

Winding	1	200 turns of	36 s.w.g.
	2	200 ,,	36 s.w.g.
	3	4000 ,,	42 s.w.g.
	4	150 ,,	36 s.w.g.

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THE BIRTH OF RADAR

By G. R. M. Garratt*, M.A., M.I.E.E.

Few great inventions have such a well defined and documented pedigree as that of radar. Two current events make the subject one of topical interest, the publication of an autobiography by Sir Robert Watson-Watt and the recent acquisition by the Science Museum, South Kensington, of the original historic apparatus.

As the year 1934 drew to its close, the increasing tempo of military preparations in Germany had begun to cause uneasiness in the minds of all intelligent observers; in the minds of those with inside knowledge of the developments on the Continent and of the utter inadequacy of our own defences there was a feeling of deep concern. One of those most anxious was A. P. Rowe, the Personal Scientific Assistant to H. E. Wimperis, Director of Scientific Research in the Air Ministry. Arising out of their mutual fears, Wimperis invited the Superintendent of the Radio Research Station, Robert Watson-Watt, to visit him to discuss whether there could be any form of damaging radiation—the so-called 'Death-Ray'—which could be used effectively against hostile aircraft.

The question was a serious one and Watson-Watt gave it serious attention. Formulating a specific problem, he considered the case of a stationary man at a range of 600 yards and, with his assistant, A. F. Wilkins, he calculated the amount of radio energy which would be required to raise the man's temperature by only 2°C in ten minutes. Using a single half-wave dipole, he showed that it would be necessary to radiate approximately five million kilowatts to give the required flux density. While the power required could be very greatly reduced by using a high-gain array, he showed that the array itself then became so large as to be quite impracticable.

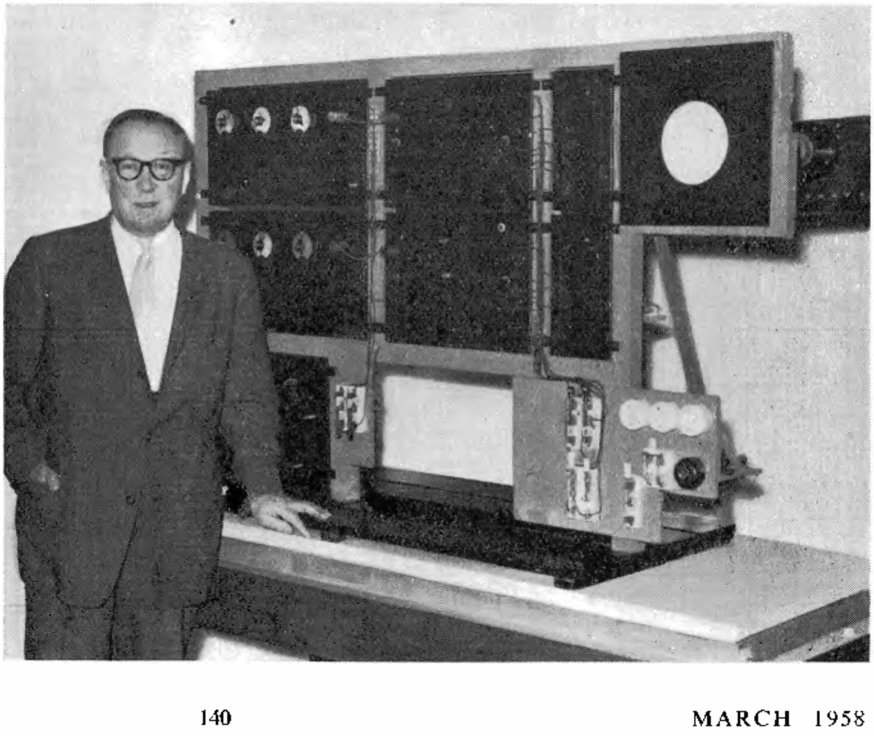
Watson-Watt summarized the results of his calculations in a Memorandum written towards the end of January 1935, but although he concluded that the pilot and structure of an all-metal aircraft must be quite immune to attack by any form of radio 'death-ray,' the final paragraph of his Memorandum contained a practical suggestion. In words which have become historic he wrote:

> "Meanwhile attention is being turned to the still difficult but less unpromising problem of radio-detection as opposed to radio-destruction, and numerical considerations on the method of detection by reflected radio waves will be submitted when required."

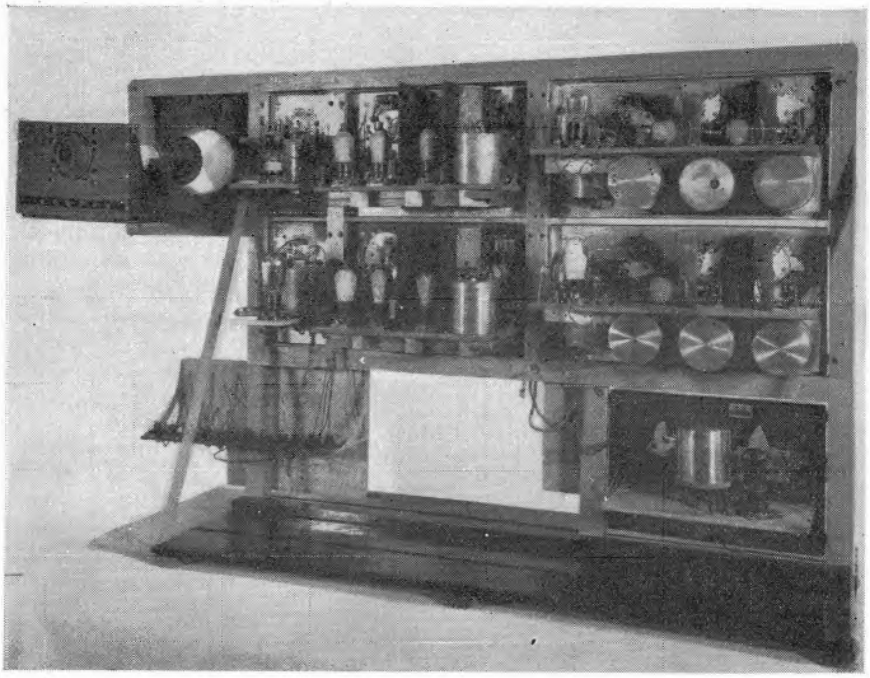
The ray, not of death but of hope, thus offered was seized upon by the Committee for the Scientific Study of Air Defence which had been set up a few months previously under the Chairmanship of Sir Henry Tizard, and Watson-Watt was promptly asked to amplify his proposal for a method of detection by reflected radio waves. Within a fortnight he had drafted a second memorandum which bore the title "Detection and Location of Aircraft by Radio Methods".

This second memorandum must surely be one of the most prophetic scientific documents ever produced. In barely two thousand words, it dismissed the possibility of using any form of primary radiation from an aircraft as the basis of a reliable detecting system, but it showed that secondary (i.e. reflected) radiation would seem to provide adequate signal strength for detection at reasonable ranges. It showed the great advantage to be gained from using pulse techniques and described how distance might be measured by timing the pulse echoes with a cathode-ray tube. While showing that position could be determined by distance-measurements from three stations, it proposed that the use of rotating beams would permit direction as well as distance to be shown on the cathode-ray tubes and, in addition, it gave promise of a means for height-finding. It stressed the need for vigorous development of techniques for using much shorter waves than those then available.

Sir Robert Watson-Watt with the receiving equipment used near Daventry on 26 February 1935 with which the reception of measurable echoes from an aircraft was first demonstrated.
The three upper racks contain the 2-stage h.f. amplifier, the 2-stage i.f. amplifier and the output (i.f.) stage of Receiver A. The three middle racks contain the same units of Receiver B, while the common oscillator for frequency changing is at the bottom left-hand corner. The phase-changing goniometer is now missing, having been destroyed some years ago.



* *The Science Museum, London.*



Rear view of the receiving equipment (with the screening covers removed).

wavelength of 49·8m from an array of vertical dipoles beamed in a southerly direction. By arranging for an aircraft to fly up and down the beam and by providing receiving equipment which could compare the phase of the ground waves received direct from the transmitter with those of the waves reflected from the aircraft, it was hoped to prove the existence of an echo of adequate strength by observing the beats as the ground and reflected waves passed in and out of phase with the motion of the aircraft.

In order to be sure of obtaining strong reflections, it was necessary that the experiment should take place near Daventry, but this introduced a difficulty in that the receivers would be subject to very powerful ground waves and it was necessary to provide means for very greatly attenuating the ground waves in order to avoid swamping the far weaker ones reflected from the aircraft.

As receiving equipment, it was decided to make use of some experimental apparatus which had previously been employed for the measurement of the angle of down-coming radiation from far-distant radio stations. Briefly, it comprised an aerial system made up of a pair of horizontal dipoles, a phase-shifting goniometer and a pair of matched super-heterodyne receivers arranged to feed the Y plates of a cathode-ray tube.

On 25 February 1935 a Morris van with a plain biscuit-coloured caravan-like body left the Radio Research Station at Slough en route for Daventry. Although their task was highly secret, it is doubtful whether the Scientific Officer in charge, Arnold Frederick Wilkins, or his driver A. J. Dyer could have realized the dramatic changes which their trip was to originate or the vast influence it was to have upon the nations of the world in the next few years.

About five miles from the Daventry transmitter, Wilkins and Dyer located a suitable site in a field near Weedon. The chosen spot was about one mile west of the Holyhead Road and situated on the line of maximum radiation from the transmitter.

Late in the afternoon, the aerial system was erected, two half-wave horizontal dipoles being set up on 15ft poles. The dipoles were spaced about 50ft apart and placed at right-angles to the direction of the transmitter so as to ensure maximum pick-up of radiation reflected from the aircraft which, on the following day, was to fly up and down the beam. Each of the dipoles was separately connected to the receivers through lengths of ordinary twisted lighting flex.

Attenuation of the ground waves was accomplished by the use of an adjustable phase-changing device in front of one of the two receivers as shown in Fig. 1. It will be realized that the ground waves produced e.m.f.'s in the two aerials of equal amplitude but with a phase difference depending on the wavelength and the spacing of the aerials. By inserting the phase-changer in the feeder from one of the aerials, it was possible to ensure that the phase of the ground waves entering one of the receivers was in opposition to the phase of the waves entering the other. In this way it was possible greatly to attenuate waves coming

but it declared that the urgency seemed to demand the making of a start on wavelengths which could then be handled. Finally, it referred to the need for being able to differentiate between friend and foe and suggested the means by which this could be accomplished.

This remarkable document—it has sometimes been termed the 'Birth Certificate' of radar—was forwarded in draft form to the Air Ministry by Watson-Watt on 12 February 1935. Containing as it did a hint that detection of aircraft at ranges of over one hundred miles was within the bounds of possibility, it is perhaps not surprising that it seemed to Wimperis and his colleagues almost 'too good to be true'.

To a scientist working in his own field, calculations—even on the back of the proverbial envelope—provide adequate assurance that some experiment will lead to a specific result. He does not need actually to perform the experiment to find the answer because he knows just what to expect already. Watson-Watt had no need to try any experiment in February 1935—he knew from his calculations very roughly what the answer would be—but it is perhaps not surprising that Wimperis and his colleagues, in particular Air Marshal Sir Hugh Dowding, should have expressed a wish for a practical demonstration. In response to this request, Watson-Watt replied in effect that a full-scale experiment would be costly and take a long time. Instead, he undertook to stage a quick demonstration, not of the full scheme he had proposed, and not even using a pulse technique, but aimed merely at showing that his arithmetic was approximately correct. "We will use 'lashed-up' equipment which was designed for quite another purpose. It will be just a rough experiment and if it doesn't work, it doesn't mean a thing, except that some little thing has gone wrong or that we have overlooked some quite insignificant point of detail".

For the purpose of this rather impromptu demonstration, it was decided to make use of one of the BBC's short-wave transmitters at Daventry which could radiate 10kW on a from the direction of the transmitter while maintaining almost full sensitivity to waves striking the aerial system from other directions.

Briefly, the phase changer comprised a pair of tetrodes in push-pull, the grids of which were both connected to the input from the aerial. The two field coils of a goniometer were connected in the respective anode circuits, one coil being detuned above resonance and the other below so as to produce a rotating field in the volume swept by the search coil. The phase of the e.m.f. in the search coil could thus be varied continuously from 0° to 360°.

The rest of the story has been vividly described for the author by its chief participant, A. F. Wilkins and in his own words:

"After erecting the aerials we commenced the task of connecting up the large batteries of accumulators for l.t., h.t., and the cathode-ray tube supply. In the excitement of preparing the apparatus at Slough the fact that the interior of the van was quite without illumination had been overlooked. No obvious means of improvising any electric light could be thought of and this mass of batteries was connected up, firstly, with the aid of matches struck and held by Dyer and, finally, by instinct. The difficulties of connecting up these batteries (some hundreds of volts of them) and the rest of the apparatus were such that the whole process was not complete until 0045 a.m. Just one-quarter of an hour was left in which to test the apparatus on a special transmission from GSA. By great good fortune all worked well after slight adjustments to the equipment. In particular, the neutralization of the ground wave was almost complete.

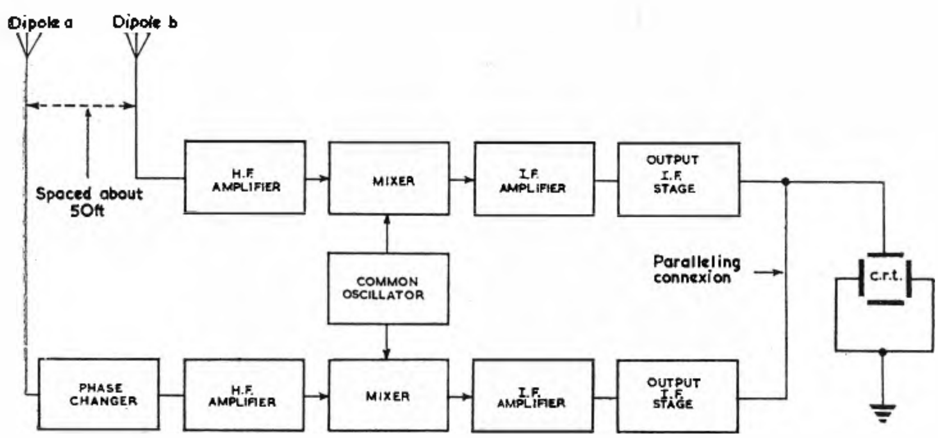


Fig. 1. General arrangement of the equipment

"It was decided to leave the aerials *in situ* and trust to luck that they would be there the next morning; the van with its valuable load could not be left, particularly as there was no lock on the doors. On attempting to drive away, however, Dyer found that the van would not move an inch. The wheels, which had sunk deeply into the soft ground on arrival in the field were now firmly gripped by earth which had frozen hard while the occupants were struggling within the van. Fortunately we found a spade in the equipment of the vehicle and after another half hour of hard work we were free. We ultimately let ourselves into our hotel at 3.0 a.m.

"Arrangements had previously been made for a Heyford bomber to fly up and down the beam at a height of 5 000ft between 9.45 a.m. and 10.30 on 26 February. On arrival at the site at about 8.0 a.m. We found the aerials still in place and, when we connected them up to the receivers everything proved to be in order.

"At the hour arranged by telephone the previous evening Watson-Watt and A. P. Rowe of the Air Ministry arrived at the site. Accompanying them was Watson-Watt's nephew, Pat, who had come for the sake of the ride but, as the proceedings in the van were classed as highly secret, he and Dyer were sent to the remote end of the field to explore.

"Our party in the van now started to peer anxiously at the cathode-ray tube on which only a short vertical line caused by the vestige of the ground ray could be seen.

"At about 9.45 a.m. the sound of a distant aircraft was heard and in due course the Heyford was identified flying towards Daventry but well off the scheduled course. After some minutes the aircraft appeared again flying south but still off the main beam, but now the expected beating of the signal was seen on the oscillograph and a maximum trace length of about two inches was observed. This performance was repeated as the Heyford flew up and down the beam and, by noting the time at which the deflexion disappeared and assuming that the aircraft was flying at the requested 100 m.p.h., we estimated that a detection range of about eight miles had been achieved.

"Unfortunately no record of the beats was made. We had provided ourselves with a siphon recorder and other necessary apparatus to make records but, when the apparatus was connected during the flight, no record appeared. It was discovered immediately after the test that a connexion in an obscure part of the apparatus had broken loose on the journey.

"Considering the crude nature of the apparatus used and the small amount of preparation the results of this experiment were regarded as quite creditable. Whatever the results we obtained, Rowe was move to exclaim that the experiment was the most successful one he had ever witnessed.

"It was clear to all of us who watched the tube on that occasion that we were at the beginning of great developments in the art of air defence and now that we had shown that aircraft could re-radiate radio waves in measurable amounts at useful ranges, we were anxious to demonstrate the efficacy of the other proposals set out in Watson-Watt's memoranda. Presumably the minds of Watson-Watt and Rowe were so full of such future hopes that it was not until they had passed through Towcester on the return journey that they remembered that poor Pat was still exploring the fields of Weedon!"

In those few and characteristically modest words, Wilkins has told the story of what was perhaps one of the most momentous experiments ever performed. Its success led directly to the spending of many millions on the development and production of radar equipment in the few years which were to follow. Radar saved us from defeat in the Battle of Britain, it paved the way for ultimate victory in 1945 and to those pioneers whose foresight, imagination and brilliant achievement laid the foundations we must for ever be deeply grateful.

A Twin-T Variable-Slope Filter

By G. B. Miller*, B.Sc., A.M.Brit.I.R.E.

By suitable modification a twin-T notch filter is converted into a low-pass filter in which the rate of attenuation up to a specified upper frequency is controlled by a single potentiometer. Design data are given and a detailed circuit is described.

IN electronic equipment designed to handle only audio frequencies some form of attenuation is required for frequencies above the pass-band of the equipment. The rate in dB per octave (or slope) at which these frequencies must attenuate is determined largely by the use to which the equipment is being put; for some applications a slope of at least 30dB is essential, while in others a slope of 6dB may be more than enough. To meet these varying requirements some means of readily altering the slope must therefore be provided in the equipment.

Low-pass *LC* filters in which the *Q* of the inductors is variable are readily adapted to provide variable-slope characteristics, but accurately wound inductances for use

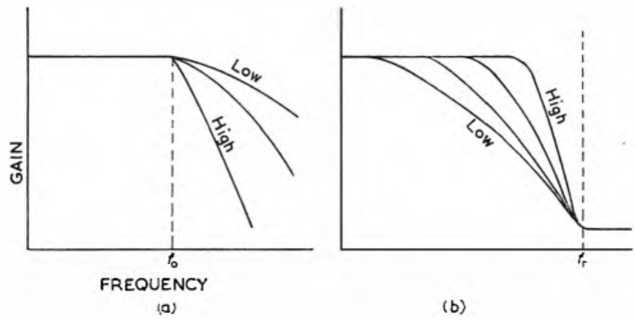


Fig. 1. Characteristics of variable slope filter

at audio frequencies are both expensive and bulky; in addition they are subject to interference from outside magnetic fields and require careful screening and placement. The same effect can be achieved by the use of *RC* networks and if these can be kept simple have much to commend them.

If a twin-T network is used in a position common to both the output and feedback paths of a feedback amplifier a notch filter characteristic is obtained. The slope of the characteristic both below and above the null frequency is determined by the amplifier gain (without feedback) and the feedback factor, and variation of either of these will change the slope. By suitable modification of the frequency/phase characteristic of the feedback loop it is possible to change the notch filter into a low-pass filter and permit the slope down to the null frequency to be controlled by a standard log-law potentiometer.

This article describes a circuit which will achieve this and gives practical details and response curves for three experimental filters.

The usual variable-slope filter has a characteristic similar to that shown in Fig. 1(a). The frequency f_o at which the response is 3dB down is kept fixed and the rate at which frequencies above f_o are attenuated is varied. However, there are many applications where this characteristic is undesirable. In the case of narrow-band transmitters for example, it is essential that all frequencies above a specified frequency should have high attenuation irrespective of the slope or cut-off frequency, otherwise excessive bandwidth may result. Modern high-fidelity pre-amplifiers, too, must have a high attenuation for frequencies outside the recorded audio band, and this attenuation must be independent of the position of the roll-off control. The type of characteristic required for these applications is similar to that shown in Fig. 1(b). In this form the filter gives a variable slope while maintaining a high attenuation for all frequencies above a specified frequency f_r.

Most of the variable filters giving this type of characteristic incorporate inductances with all the problems which go with them. The circuit to be described below enables the same results to be achieved using only *R* and

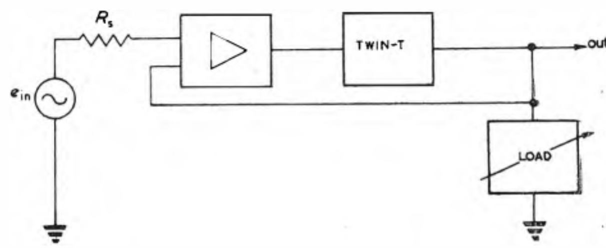


Fig. 2. Basic filter circuit

C. Details are given of three filters: the first two are suitable for modulators and give an attenuation of 35dB to all frequencies above 5·9kc/s and 10·6kc/s respectively, while the third is suitable for high-fidelity equipment since it has a pass-band of 14kc/s and a maximum roll-off rate of approximately 35dB per octave; in addition it has a minimum attenuation of 30dB for all frequencies above 21·2kc/s.

Description of Circuit

The basic circuit used is a modified form of the usual twin-T frequency-rejection negative feedback amplifier and is shown in Fig. 2. The use of a twin-T network for this purpose is well known and is fully discussed elsewhere[1,2]. If the network is driven from a low-impedance source and operates under unloaded conditions, the response is symmetrical about the null frequency of the twin-T. Thiessen[2] has shown that the characteristic can be made asymmetrical by the introduction of an additional phase lag in the feedback loop, while Buckley[3] has shown that the same result is achieved by loading the output of the twin-T. As the latter method is flexible and allows the characteristic to be controlled easily and simply, it is used in this design.

Fig. 3 shows the full circuit used. The valve V_{1a} is used to isolate the filter from the input circuit and to make it independent of the impedance of the signal source. The output from V_{1a} is cathode-coupled to V_{1b}, the grid of which is used to inject the negative feedback. The output from the anode of V_{1b} is fed through the twin-T network

* *The British Thomson-Houston Co. Ltd.*

and the voltage appearing across the slope control R_o is used to feed the voltage amplifier V_{2a} and also provides the feedback to the grid of V_{1b} via the resistor R_g. The output from the filter is taken from the anode of V_{2a}. The 220kΩ anode resistor of V_{1a} allows the filter to be switched out of circuit by taking the output from the anode of V_{1a} but when this is done a decoupling capacitor of 50μF must be switched in from the cathode of V_{1a} to ground. It must be specially noted that in all cases the response curves given are on the basis of the output feeding a load of 1MΩ shunted by 100pF.

Design Data

A detailed mathematical analysis can be made, but is not very helpful. The final response of the filter is determined not only by the twin-T network but also by the values of $C_k C_o R_o R_g$ and the C_{ag} and C_{ak} of V_{1b}. For practical design purposes the following general rules—some of them empirical—are enough to determine the main values.

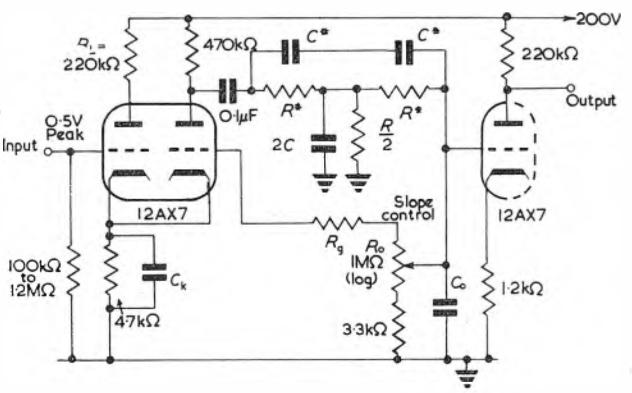


Fig. 3. The filter circuit

COMPONENT	FREQUENCY			TOLERANCE
	5·9kc/s	10·6kc/s	21·2kc/s	
R^*	270kΩ	150kΩ	150kΩ	5% ½W
C^*	100pF	100pF	50pF	5%
C_k	0·04μF	0·02μF	0·004μF	20% 350V
C_o	500pF	200pf	50pf	10%
R_g	1·2MΩ	1·2MΩ	470k	10% ½W

* Matched.

All resistors 20% ½W. All capacitors 20% 350V unless otherwise specified.

These can later be adjusted to give the final desired response.

f_r = the frequency of maximum attenuation. This is usually specified and the null frequency of the twin-T is made equal to f_r.

= $(1/2\pi/CR)$. In choosing the actual values of C and R, C should be made sufficiently large (at least 50pF) to swamp stray circuit capacitance. R should if possible be at least ¼ the recommended load for maximum gain for valve V_{1b}.

f_o = maximum frequency for which the filter response is less than 3dB down. This frequency is affected by $R_o R_g C_k C_o$ and indirectly (through R_g) by the capacitances C_{ag} and C_{gk} of valve V_{1b}. The ratio f_o/f_r is increased by an increase of R_o, a decrease of C_k and a decrease of R_g — and vice versa. Typical variations caused in the response by changes in C_o and C_k are shown in Fig. 5.

= $(0{\cdot}45 + 0{\cdot}04x)f_r$ where x is the maximum peak in dB permitted in the response. $(0 < x < 3)$. This formula is slightly pessimistic

$C_k R_k' = CR$ to $3CR$. See Fig. 5.

$$R_k' = \frac{r_a + R_L}{\mu + 1 + \dfrac{r_a + R_L}{R_k}}$$

where r_a is the anode resistance of V_{1a}

μ is the amplification factor of V_{1a}

R_L is the anode load of V_{1a}

R_k is the cathode resistor of V_{1a}.

C_o = the value used for C_o is determined by the maximum attenuation required above f_r and the value of f_o/f_r. A large value of C_o will give high attenuation above f_r but will reduce f_o and also cause a large peak in the response at a frequency just below f_o. This peak can be reduced by an increase in C_k, but this will cause a further reduction in f_o. As a rough rule C_o should be three times the value of C, but it must be remembered that the Miller effect capacitance at the grid of V_{2a} will be high and must be taken into account when determining the actual capacitance value.

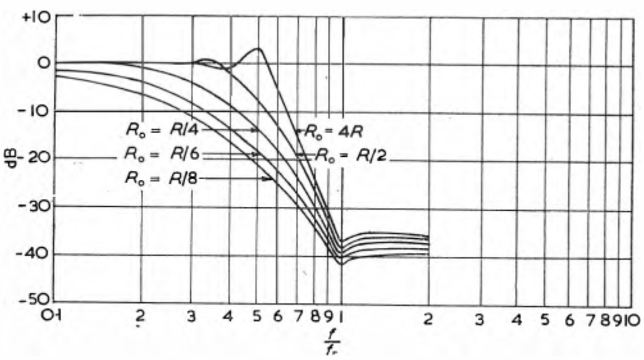


Fig. 4. Variation of cut-off slope with R_o for 5·9kc/s and 10·6kc/s filters

f = frequency of signal
f_r = frequency of twin-T
$R_g + R_o$ = 2MΩ for all curves
C_o = 6C for 5·9kc/s
C_o = 3C for 10·6kc/s

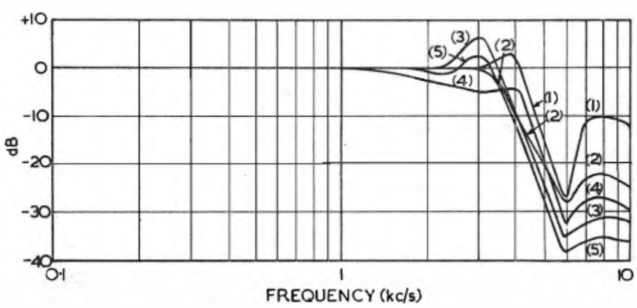


Fig. 5. Variation of cut-off slope and attenuation with C_o and C_k for 5·9kc/s filter

$R_o + R_g$ = 2MΩ for all curves
R_o = 4R for all curves
$C_k R_k'$ = 1·5CR (1) C_o < (C/10)
(2) C_o = C
(3) C_o = 6C
$C_k R_k'$ = 3CR (4) C_o = C
(5) C_o = 6C
C_o includes Miller effect capacitance

$R_g + R_o$ = maximum permitted grid to cathode resistance of V_{1b}. For a wide range of slope control R_o should be at least $4R$ and for high attenuation of frequencies above f_r, R_g should be as large as possible. As R_g will reduce f_o if it is too large a good compromise is to make $R_g = R_o$ for a ratio $f_o/f_r = 0{\cdot}5$ and $R_g = \frac{1}{2}R_o$ for $f_o/f_r = 0{\cdot}65$.

Performance of Actual Filters

Three filters have been designed using this basic circuit with f_r of 5·9, 10·6 and 21·2kc/s. Fig. 4 gives the response

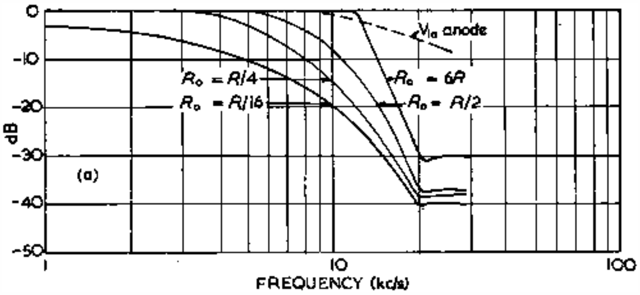
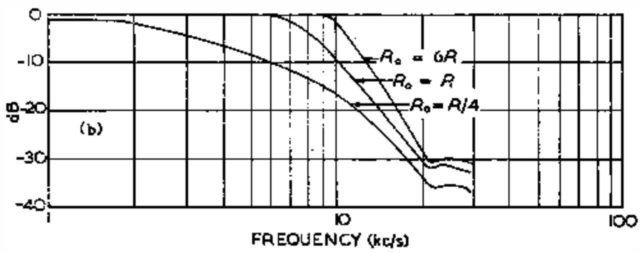


Fig. 6(a) Response of 21·2kc/s filter C_o = 50pF

$R_o + R_g = 1{\cdot}5M\Omega$
$C_k = 0{\cdot}002\mu F$
$= 50\mu F$ for V_{1a} anode curve



(b) Alternative 21·2kc/s filter

C_o = Miller effect only
$R_g = 1{\cdot}2M\Omega$
$R_o + R_g = 2{\cdot}2M\Omega$
$C_k = 0{\cdot}01\mu F$

of the first two filters for various positions of the slope control. A peak of 3dB has been allowed in the response of the filter and $f_o = 0{\cdot}57f_r$. The maximum slope is 40dB/octave, while the minimum slope is initially about 4dB/octave; the minimum attenuation for all frequencies above f_r is 34dB. If the peak in the response is not wanted the curve for $R_o = \frac{1}{4}R$ can be used and this will still give a maximum slope of 30dB/octave. The response can also be modified by reducing the values of C_o and C_k, and Fig. 5 gives enough information for the result to be predicted with reasonable accuracy.

Fig. 6(a) gives the response for the filter with $f_r = 21{\cdot}2$kc/s. In this case there is no peak in the response of the filter due to the choice of R_gC_k and C_o. R_g has been reduced to bring f_o up to 14kc/s, and the reduction in R_g requires a reduction in C_o if a peak is to be avoided. This in turn requires a reduction in C_k to avoid a response similar to that illustrated in curve 4 of Fig. 5. For comparison, the response of a filter with R_g at the maximum value of 1·2MΩ, $C_k = 0{\cdot}01\mu F$ and C_o = Miller effect capacitance only, is given in Fig. 6(b). It will be seen that the value of f_o is well below that obtained for the values used in Fig. 6(a).

The low-frequency response of these filters is very good and is equal to the level at 1kc/s for all frequencies down to well below 20c/s. In addition, the response at low frequencies is not effected by the slope control until f_o has been moved down to below 1kc/s. The filter with f_r = 21·2kc/s and whose response is given in Fig. 6(a) is therefore highly suitable as a slope control in high-fidelity equipment.

In all the filters the overall gain from the grid of V_{1a} to the anode of V_{2a} is 30dB and inputs of up to 0·5V peak can be handled without overloading. The gain from the grid of V_{1a} to the anode of V_{1a} is 35dB when the cathode is decoupled with a 50μF capacitor and the response of V_{1a} under these conditions is shown in Fig. 6(a). It must be repeated that in all cases the responses shown in Figs. 4, 5, 6(a) and 6(b) were taken with the output being fed into a load of 1MΩ shunted by 100pF in order to simulate a following high-gain triode.

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A New Shipping Control System at Southampton

On 17 January the Rt. Hon. Harold Watkinson M.P., Minister of Transport and Civil Aviation, inaugurated the Southampton Harbour Board's new Port Operation and Information Service which is believed to be the first of its kind in the world.

The equipment consists primarily of a Decca Type 32 Harbour Radar system and v.h.f. radio equipment supplied by Marconi's Wireless Telegraph Co. Ltd.

In practical terms the Information Service, by virtue of the data obtained by the v.h.f. radio and by radar, presents the Duty Operations Officer with an immediate and continuous awareness of the movements and positions of all vessels navigating within the Port Area and its approaches. With this overall picture as a background, relevant information is passed by radio-telephone to the Masters or Pilots of individual vessels.

The v.h.f. equipment has been chosen to conform to the recommendations of the International Maritime V.H.F. Conference which was held at The Hague in January 1957; in this respect the Port of Southampton has given a lead to the world by the adoption of internationally-agreed standards.

The ideal concept of the Hague Conference was to standardize harbour and shipboard v.h.f. radio installations to such a degree that any ship of any nationality would be able to enter or leave any port in full radio-telephone communication with the Harbour Authorities. Southampton is an important first step in this direction, for the fullest co-operation has been forthcoming from other Authorities. The G.P.O. has announced that Public Correspondence v.h.f. radio-telephone facilities are being planned from existing stations at the Land's End, Isle of Wight, North Foreland and Humber areas; the Corporation of Trinity House has equipped the pilot cutters in the Isle of Wight area and their motor launch "Jessica", stationed at Hythe, with the appropriate equipment; shipowners, oil companies and others immediately concerned have given the project their closest support.

The specification for the Harbour Surveillance Radar at Southampton was prepared by the Ministry of Transport and Civil Aviation Research Group after observations had been made for a period of two months with a trial radar installation. The Decca Type 32 which was chosen for the installation is a 3cm high definition equipment, using an aerial of 25ft span. For reliability and to ensure continuous operation during routine maintenance there are two entirely separate radar channels with remote control changeover switching facilities. There are also 3 separate displays which can be independently operated if necessary. A feature of the design concept of the station at Southampton is the facility of selecting on any display unit any one of 5 pre-determined display areas which together make up the planned coverage of the station. These areas are Southampton Water itself, the western approaches from the Solent (including Ryde and Spithead), the eastern Solent (including Cowes Roads), a specially large scale view of the petroleum terminals (including Shell-Mex and B.P. Limited's bunkering station and the Esso marine terminal just north of Calshot) and finally a general warning, all round picture extending approximately 18 miles from the radar station and including coverage of the Nab.

Measurement of Small Phase Shifts with a Phase Sensitive Voltmeter

By D. J. Collins*, B.Sc.(Eng.), A.M.I.E.E., and
J. E. Smith*, B.Sc.

The design of precision electronic circuits sometimes requires the measurement of very small phase errors to a reasonable accuracy. The article describes a method of measuring phase errors of the order of one degree using conventional instrumentation.

WITH the increase of systems using closed loop feedback control, there has been a greater demand for instrumentation capable of determining the transfer functions of elements employed in the control loop.

The usual method of transfer function analysis is to energize the unit under test at a number of fixed frequencies, covering the frequency range of interest, using a constant amplitude input signal. The output amplitude and phase-angle are noted in each instance and the results are then recorded on polar co-ordinate paper, in the form of a Nyquist diagram.

The Phase Sensitive Voltmeter

This voltmeter is ideally suited to the above method of measurement. It receives two input voltages: one, the energizing voltage fed to the unit under test, usually known as the reference voltage, and the other—the output from the unit—is known as the signal voltage.

The voltmeter is fitted with two 6in scale length thermocouple wattmeters. The function of these is to indicate in terms of true voltage the resolved components of the signal voltage which are in phase with and at 90° to the reference voltage input. The in-phase component is indicated on the reference meter and the 90° component on the quadrature meter. This simultaneous resolved component presentation permits direct plotting of a Nyquist diagram on ordinary rectangular graph paper.

Advantages and Limitations

The voltmeter has many advantages over other methods of measurement, e.g. high rejection of harmonics and noise, high input impedance, large voltage range, and good accuracy. Although the accuracy, ±2 per cent of full scale in each quadrant, is completely adequate for most purposes, it is not so for uses involving the measurement of small phase errors in units having a nominal 0° or 180° phase shift. This is because, under these conditions, the quadrature meter indication is very small compared with that of the reference meter. If, in fact, the quadrature indication is as low as 2 per cent of full scale, it is unlikely that the angle measurement will have any great accuracy, as the indication is comparable with the instrument error.

Obviously, there are two solutions to overcoming inaccuracies under these conditions:

(1) To improve the general accuracy of the voltmeter.

(2) To increase the quadrature indication until it is large compared with the inaccuracy.

The first solution can be ruled out on the grounds that the accuracy of indication is almost entirely dependent on the accuracy of the thermocouple wattmeter, and there appears to be no hope of any immediate improvement in this accuracy.

The second solution cannot be achieved by increasing the overall voltmeter sensitivity, since the reference meter would then be overloaded. A better method is to 'back off' the in-phase component of the signal, thus reducing the reference meter indication and allowing the voltmeter sensitivity to be increased, with consequent increase in quadrature indication.

Method of Measurement

This technique will be described as applied to the testing of the two general cases met in practice:

(a) To test a unit having a nominal phase shift of 180°.

(b) To test a unit having a nominal 0° phase shift.

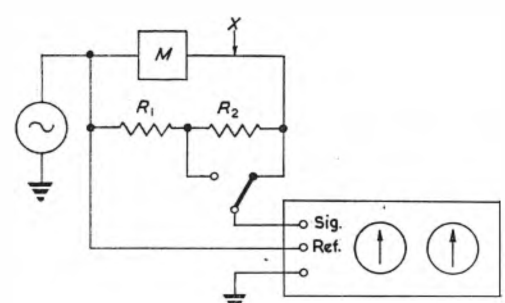

Fig. 1. Basic circuit configuration for unity gain

First, consider the nominal 180° phase shift case and, for simplicity, it is proposed to deal with an amplifier of unity gain. The conditions for the test are shown in Fig. 1.

The operator sets the input to a suitable level for the test, and adjusts the phase sensitive voltmeter on this signal. The voltmeter is then switched to the 'use' position, and is used to monitor the amplifier output. The signal amplitude measured is sensibly all 'reverse phase' component. This reverse phase reading is noted and the voltmeter is then switched to the junction of the two adding resistors R_1 and R_2. These resistors should be equal in value and, if the amplifier gain is unity, the signal appearing at their junction will be practically all quadrature. The sensitivity of the voltmeter can now be increased to give an accurate measurement of this component. It should be remembered that the reading should be doubled as the quadrature component at the amplifier output is halved by attenuator R_1: R_2.

Considering the case of the unity gain amplifier with a nominal 0° phase shift, it is obvious that the same

* *Solartron Research & Development Ltd.*

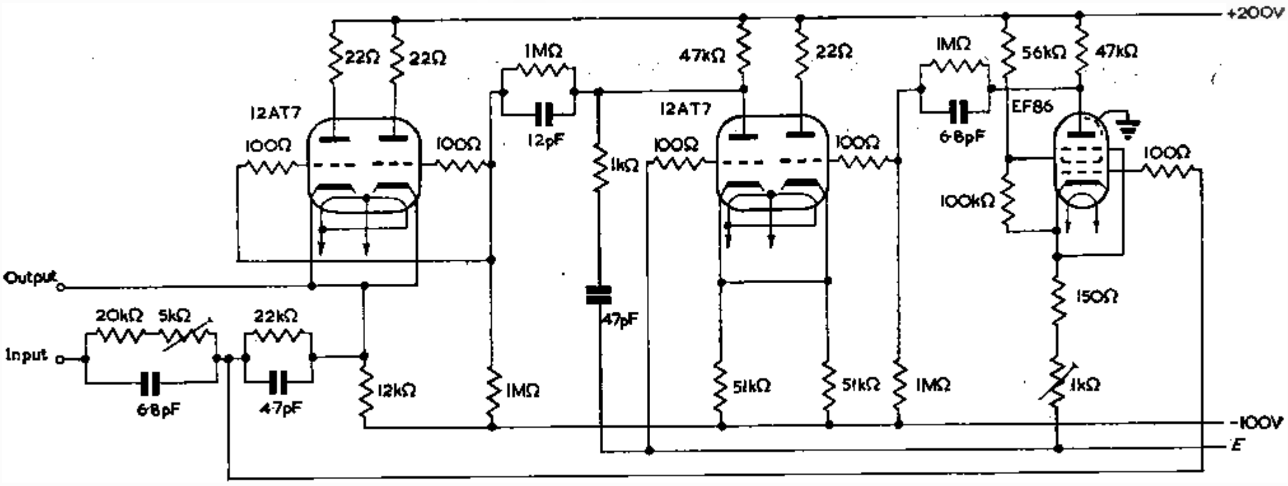

Fig. 2. Circuit diagram of image amplifier

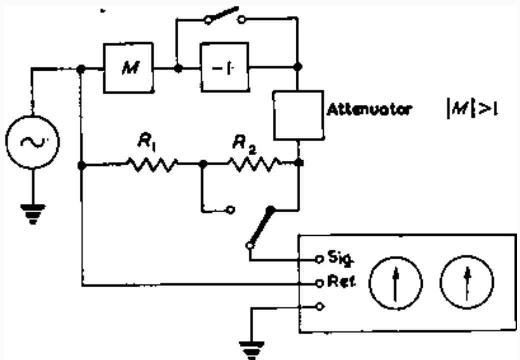

Fig. 3. Circuit configuration for devices having gain greater than unity

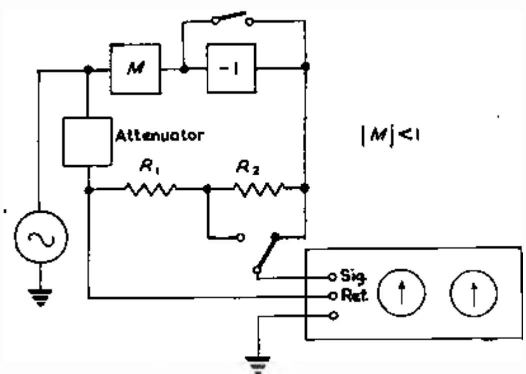

Fig. 4. Circuit configuration for devices having gain less than unity

technique can be applied as above, if the amplifier output is accurately phase reversed. A phase reversing element can be inserted at x, taking the form of a transformer or, more conveniently, an image amplifier of the type shown in Fig. 2. This type of amplifier has unity transmission from d.c. up to about 20kc/s, and the small phase errors occurring in it at the higher frequencies can be quickly compensated using the above 'backing off' technique.

To enable amplifiers of gain greater than unity to be tested, it is only necessary to insert an accurate attenuator at the output of the amplifier under test to give the correct balance of in-phase signals developed across R_1 and R_2. For amplifiers of less than unity gain, it is obviously necessary to attenuate the reference signal being fed to R_1. The general set up for these conditions is shown in Figs. 3 and 4.

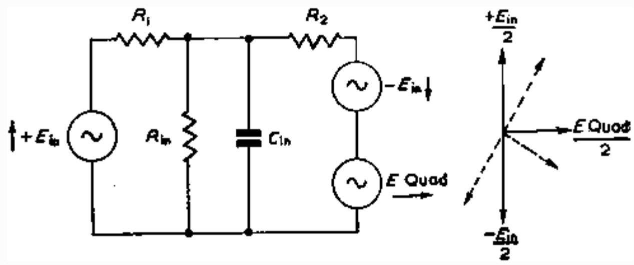

Fig. 5. Equivalent circuit of 'adding' system

PRACTICAL CONSIDERATIONS

It can be seen that the attenuators used in the above tests should have negligible phase error at the frequency concerned. The values of R_1 and R_2 should also be sufficiently high to prevent loading on the sources to which they are connected.

The phase sensitive voltmeter itself constitutes a load at the junction of R_1 and R_2, and can give rise to errors in the quadrature indication. Fig. 5 gives the equivalent circuit of R_1, R_2 and the input impedance of the voltmeter C_{in} and R_{in} in parallel. If R_{in} is much greater than R_1 and R_2, then amplitude errors due to pure resistive loading can be neglected.

The phase shift in the signal caused by C_{in} produces only a small amplitude error in the quadrature reading, providing that the cancellation of in-phase signal is good. Thus, for an accurate quadrature reading, it will be necessary to adjust the attenuator to give a zero reference meter reading.

By careful use of the above technique, it has been possible to measure a phase shift of 10′ to within an accuracy of $\pm 2'$ on a 1·0V r.m.s. signal level at 10kc/s.

Acknowledgments

The authors wish to record their thanks to the Directors of Solartron Research & Development Ltd for permission to publish this article.

A New Bistable Element Suitable for Use in Digital Computers

(Part 2)

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

Speed of Switching Off

The analysis for the switch-off case is given in Appendix B. Whereas in the case of switch-on one waveform is sufficient to compare theory and practice, in the case of switch-off this is not so. Because of this seven currents are plotted in Fig. 10 and each current is plotted first as calculated and then as measured. The circuit constants used were the same as those in the turn-on case with in addition the following:

i_5	=	5mA
X_3 cut-off frequency	=	8Mc/s ($T = 2 \times 10^{-8}$sec)
X_3 base-collector capacitance	=	10pF ($k_3 = 0{\cdot}1$)
Turn off input capacitance	=	100pF.

Because the turn-off time is broken into several stages, the times for some of these stages is quite short and most oscilloscopes introduce error in measurement. The rise-time of the oscilloscope used is also shown in Fig. 10 and in some cases the result to be expected when using this oscilloscope is shown in dotted lines on the theoretical curves.

Comparison of the theoretical and experimental results reveals agreement that is, on the whole, good. Various discrepancies occur which show the limit of the theoretical treatment. For instance i_{e2} is falling while i_{d8} is still reducing towards zero†, whereas if D_8 is a perfect switch i_{e2} cannot fall until D_8 is open. The same cause (of diode impedance) results in the curve for i_{d6} being flatter in its mid portion than it should be: some of the current which should flow to D_6 via the base-collector capacitances is diverted into the base of X_1 and is revealed by the hump on the waveform of i_{e1}.

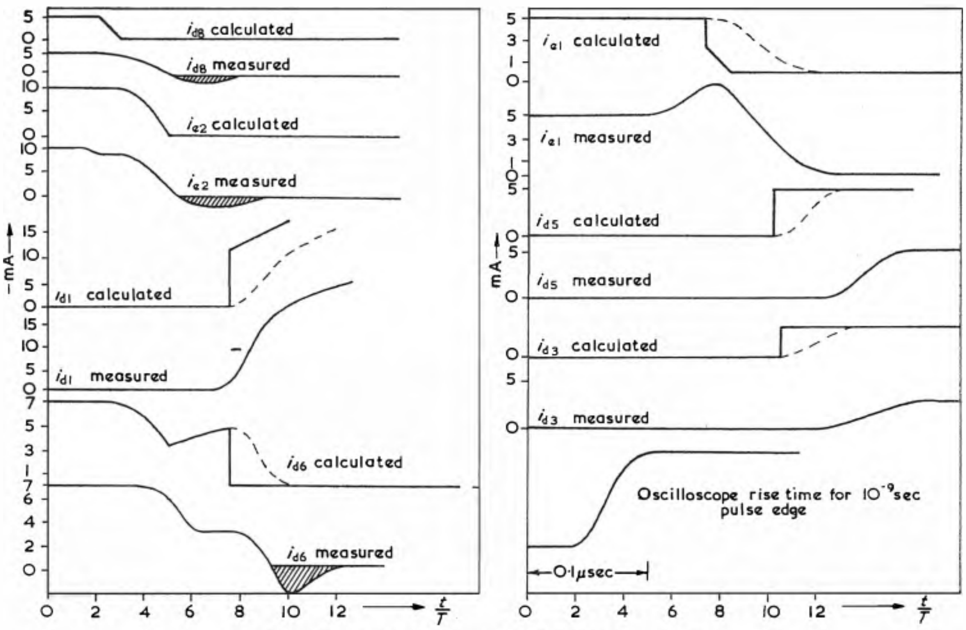

Fig. 10. Calculated and measured switch-off characteristics

Other discrepancies arise because of diode turn-off time, whether the diode is a separate component or part of a transistor. Instances of this, resulting in the flow of reverse current, occur in the curves for i_{d8}, i_{e2}, and i_{d6}, where the reverse currents are shown shaded.

The errors in timing between the end of the i_{e1} waveform and the i_{d5} and i_{d3} waveforms are thought to be due to two causes: first the loss of current because of reverse current in D_6 and, secondly, the necessary voltage drop across these diodes before they pass their allotted currents causes V_{e1} and V_{b1} respectively to fall through greater potentials than those used in the calculations.

Switching Off—Output Waveform

The rate of rise of the output waveform is given by equation (39) of Appendix B. For a value of i_7 of 10mA and a load capacitance of 500pF (equivalent to the load of five similar stages) the rise time is 0·5μsec.

Resolving Time

As it has been shown that both switching-on and switching-off can be accomplished in about 0·2μsec each, it may be thought that the circuit could resolve equally spaced input waveforms 0·5μsec apart. This is not so with the circuit as shown, whose resolving time is about 1·5μsec. The reason for this long time lies in the behaviour of the turn-off transistor X_3 and is analysed in Appendix B under stage 7. A calculated waveform for the collector current of this transistor is shown in Fig. 11 where the turn-off input and output waveforms are also shown. The collector current was calculated for the circuit shown in Fig. B.1, with α_0 of X_3 assumed to be 0·98 and i_9 assumed to be 0·25mA.

It can be seen that with these values the circuit is in a quiescent state after 75 time-constants, giving a resolving time for 8Mc/s transistors of 1·5μsec. This time could be improved by increasing i_9, but this would result in a loss of turn-off sensitivity. A way out of this difficulty is described by N. F. Moody[2] is a companion article.

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† *These designations are defined on Fig. B.1.*

Practical Circuit

So far, the description of the development of the circuit and also the analyses have used constant current sources. For practical purposes these can be simulated by moderate voltages and resistances. Fig. 12 shows such a practical circuit: several hundred laboratory models of this have been built and tested with very satisfactory results.

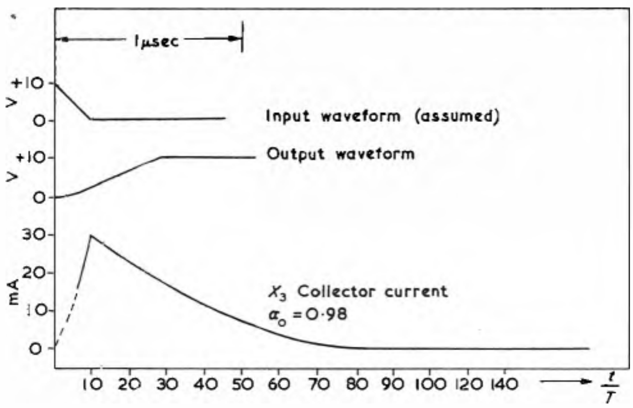

Fig. 11. Calculated collector current waveform

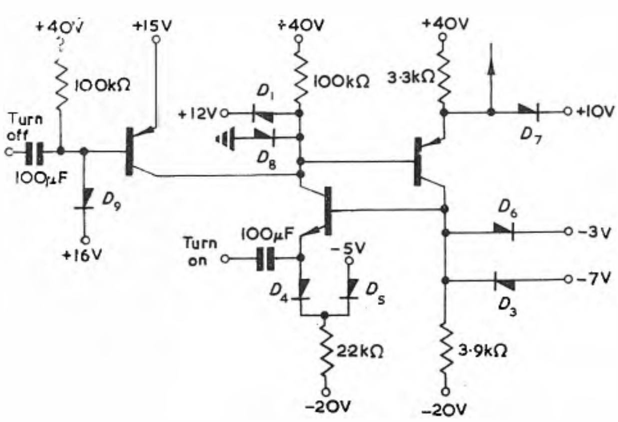

Fig. 12. A practical circuit

Conclusions

It has been shown, both experimentally and theoretically, that the complementary pair trigger circuit, when driving five similar stages, has the following characteristics:

Switch-on time	0·1μsec (approx.)	
Output fall time	0·1μsec	,,
Switch-off time	0·2μsec	,,
Output rise time	0·5μsec	,,
Resolving time	1·5μsec	,,

Also, the current handling capacity in both the on and the off states is excellent, and there are no tight tolerance requirements. The transient response has been shown to be calculable to a good order of accuracy.

Acknowledgments

The author would like to thank Mr. N. F. Moody for constant encouragement and stimulating discussions, Mr. D. P. Henderson for performing the measurements in a very careful manner and the Defence Research Board of Canada for permission to publish this article.

Appendix B

TURN OFF

Fig. B.1 is a reproduction of Fig. 9 of the article, with a 'wetting' current applied to X_3 similar to those described in Appendix A. The base-collector capacitances and the input and load capacitances have also been added and the directions shown for internal currents.

Stage 1

During the first stage of turn-off the base of X_3 falls until it reaches the potential of the emitter diode. It will be assumed that i_{10}, the wetting current, is sufficiently small that it plays no part in the circuit operation except to nullify the effect of emitter/base capacitance.

$$i_{x5} = (V_{b3} - V)pC = i_9 - i_{x3}$$
$$i_{x3} = V_{b3}pk_3C$$

Thus

$$V_{b3}pC(1 + k_3) = i_9 + pVC$$

But $V = -at$ so $pV = \frac{dV}{dt} = -a$

Hence $V_{b3}pC(1 + k_3) - i_9 + aC = 0$

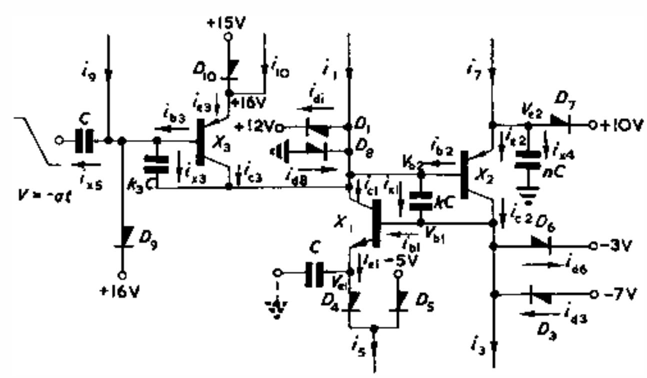

Fig. B1. Reproduction of Fig. 9 with 'wetting' current applied to X_3

Initially $V_{b3} = 0$

$$\text{so } \overline{V}_{b3}pC(1 + k_3) - \overline{i}_9 + \overline{aC} = 0.$$

Now $\overline{i}_9 = (i_9/p)$ and $\overline{a} = (a/p)$

Therefore

$$\overline{V}_{a3} = \frac{i_9 - aC}{p^2C(1 + k_3)}$$

which gives

$$V_{b3} = \frac{T(i_9 - aC)}{C(1 + k_3)}t/T \qquad \ldots\ldots (26)$$

Stage 2

During the second stage of turn-off the collector current of X_3 builds up until D_8 is cut off. Initially the current passed by D_8 is equal to $\alpha_0 i_5 - i_1 - (1 - \alpha_0)i_7$ and these currents remain constant during this phase. Since the base of X_3 is clamped to the emitter, and as D_8 is conducting, no current can pass through k_3C (base to collector capacitance of X_3).

Therefore $i_{c3} = \frac{\alpha}{1 - \alpha}i_{b3} = \frac{\alpha}{1 - \alpha}aC$ if i_9 is neglected.

After substitution for α $\left[\alpha = (\alpha_0/T)\left(\frac{1}{p + 1/T}\right)\right]$ and the initial condition that $i_{e3} = 0$ this leads to

$$i_{c3} = \frac{\alpha_0 aC}{1 - \alpha_0}\left[1 - \exp\left(-\frac{(1 - \alpha_0)t}{T}\right)\right] \ldots\ldots (27)$$

At any time during this phase $i_{d8} = \alpha_0 i_5 - i_1 - (1 - \alpha_0)i_7 - i_{c3}$ so

$$i_{d8} = \alpha_0(i_5 + i_7) - (i_7 + i_1) - \frac{\alpha_0 aC}{1 - \alpha_0}\left[1 - \exp\left(-\frac{(1 - \alpha_0)t}{T}\right)\right]$$

$$\ldots\ldots\ldots\ldots\ldots (28)$$

which equation defines the duration of stage 2, since this ends when $i_{d8} = 0$.

Equation (28) can be approximated as $\alpha_0 \to 1$, by

$$i_{d3} \simeq i_7 - i_1 - aC\,(t/T) \; \ldots\ldots\ldots\ldots \; (29)$$

Stage 3

As D_3 cuts off at the end of stage 2 the base and emitter of X_3 are free to rise in potential during the next stage, which will last until i_{e2} reaches zero. (This statement is true with normally-used values for circuit currents: other values may result in stage 3 terminating when V_{b2} reaches +10V or when D_6 cuts off).

We have

$$i_{e2} = \frac{i_{b2}}{1-\alpha} \; : \; i_{b2} = i_{x1} + i_{c1} - i_1 - i_{x3} - i_{c3}$$

$$i_{x1} = pkCV_{b2} \; : \; i_{x3} = -pkCV_{b2}$$

$$i_{c3} = \frac{\alpha i_{b3}}{1-\alpha} \; : \; i_{b3} = aC - i_{x3}$$

$$\alpha = \frac{\alpha_0}{T(p+(1/T))} \; : \; V_{b2} = \frac{i_7 - i_{e2}}{pnC}$$

Solving for i_{e2},

$$i_{e2}\left[p^2(n+k+k_3) + p\left(\frac{2n(1-\alpha_0)+k(2-\alpha_0)+2k_3}{T}\right) + \frac{n(1-\alpha_0)^2+k(1-\alpha_0)+k_3}{T}\right]$$

$$-i_7\left[p^2(k+k_3) + p\left(\frac{k(2-\alpha_0)+2k_3}{T}\right) + \frac{k(1-\alpha_0)+k_3}{T^2}\right]$$

$$-n(i_{c1}-i_1)\left[p^2 + p\,\frac{2-\alpha_0}{T} + \frac{1-\alpha_0}{T}\right] + \frac{n\alpha_0 aC}{T}$$

$$(p+(1/T)) = 0 \; \ldots\ldots\ldots \; (30)$$

Initially, $i_{e2} = i_7$, and all derivatives = 0.

The r.h.s. of the equation can thus be replaced by

$$pni_7 + \frac{2n(1-\alpha_0)}{T}\,i_7 - (i_{c1}-i_1)\left(p + \frac{2-\alpha_0}{T}\right) + \frac{n\alpha_0 aC}{T}$$

and the currents on the l.h.s. replaced by their transforms.

Now $\overline{i_7} = (i_7/p)$, $\overline{(i_{c1}-i_1)} = \frac{i_{c1}-i_1}{p}$, $\overline{a} = (a/p)$

so, by substituting and re-arranging,

$$\overline{i_{e2}} = (i_7/p) +$$

$$\frac{n[[i_{c1}-i_1-i_7(1-\alpha_0)](1-\alpha_0)-\alpha_0 aC]}{(n+k+k_3)T^2p\left\{p + \frac{2n(1-\alpha_0)+k(2-\alpha_0)+2k_3+\alpha_0\sqrt{(k^2-4k_3n)}}{2(n+k+k_3)T}\right\}\left\{p + \frac{2n(1-\alpha_0)+k(2-\alpha_0)+2k_3-\alpha_0\sqrt{(k^2-4k_3n)}}{2(n+k+k_3)T}\right\}}$$

which can be written as

$$\overline{i_{e2}} = (i_7/p) + \frac{g}{T^2p\left(p+\frac{\gamma+\delta}{T}\right)\left(p+\frac{\gamma-\delta}{T}\right)}$$

or

$$\overline{i_{e2}} = (i_7/p) + \frac{g}{\gamma^2-\delta^2} \times$$

$$\left[1/p + \frac{\gamma-\delta}{2\delta\left(p+\frac{\gamma+\delta}{T}\right)} - \frac{\gamma-\delta}{2\delta\left(p+\frac{\gamma+\delta}{T}\right)}\right]$$

Now δ is equal to $\frac{\alpha_0\sqrt{(k^2-4k_3n)}}{2(n+k+k_3)}$, which is negative if $k_3n > (k^2/4)$. Since usually $k_3 \sim (k/2)$, δ will be negative if $n > (k/2)$. As it is most unlikely that n will have a value less than $k/2$, we can write jδ in place of δ, then

$$\overline{i_{e2}} = (i_7/p) + \frac{g}{\gamma^2+\delta^2} \times$$

$$\left[1/p + \frac{\gamma - j\delta}{2j\delta\left(p+\frac{\gamma+j\delta}{T}\right)} - \frac{\gamma+j\delta}{2j\delta\left(p+\frac{\gamma-j\delta}{T}\right)}\right]$$

whence

$$\overline{i_{e2}} = i_7 + \frac{g}{\gamma^2+\delta^2} \times$$

$$\left[1 - e^{-\gamma t/T}\left((\cos\delta t/T + \gamma/\delta\sin\delta t/T)\right)\right] \ldots \; (31)$$

If the assumption is made that $k = 2k_3$ and a new constant $k_2 = n/k_3$ is introduced, then equation (31) simplifies when $\alpha_0 \to 1$ to

$$i_{e2} \sim i_7 - k_2aC\left[1 - \exp\left(-\frac{2}{k_2+3}\,t/T\right) \times \left(\cos\frac{k_2-1}{k_2+3}\,t/T + \frac{2}{k_2-1}\sin\frac{k_2-1}{k_3+3}\,t/T\right)\right] \ldots \; (32)$$

For the initial conditions in the next stage we shall need to know the values of V_{b2}, dV_{b2}/dt, and i_{e2} at the end of stage 3.

From $V_{b2} = \frac{i_7 - i_{e2}}{pnC}$ and equation (30),

$$V_{b2}pnC\left[p^2 + p\left(\frac{2n(1-\alpha_0)+k(2-\alpha_0)+2k_3}{(n+k+k_3)T} + \frac{n(1-\alpha_0)^2+k(1-\alpha_0)+k_3}{(n+k+k_3)T^2}\right)\right] -$$

$$ni_7\left[p^2 + p\,\frac{2(1-\alpha_0)}{T} + \frac{(1-\alpha_0)^2}{T^2}\right] +$$

$$n(i_{c1}-i_1)\left[p^2 + p\,\frac{2-\alpha_0}{T} + \frac{1-\alpha_0}{T^2}\right] - (n\alpha_0 aC)/T\,(p+1/T) = 0.$$

Initially $V_{b2} = 0$, and all derivations are zero. Also

$\overline{i_7} = i_7/p$, $\overline{i_{c1} - i_1} = \frac{i_{c1}-i_1}{p}$, $\overline{a} = a/p$.

Inserting the initial conditions and the transforms we get, after re-arranging,

$$\overline{V_{b2}} = -\frac{n[[i_{c1}-i_1-i_7(1-\alpha_0)](1-\alpha_0)-\alpha_0 aC]}{(n+k+k_3)T^2nCp^2\left[p^2 + p\,\frac{(2n(1-\alpha_0)+k(2-\alpha_0)+2k_3}{(n+k+k_3)T} + \frac{n(1-\alpha_0)^2+k(1-\alpha_0)+k_3}{(n+k+k_3)T^2}\right]}$$

which may be written, using the same symbolism as in the derivation of i_{e2}, as

$$\overline{V_{b2}} = -\frac{g}{nCT^2p^2\left(p+\frac{\gamma+j\delta}{T}\right)\left(p+\frac{\gamma-j\delta}{T}\right)}$$

the inverse transform of which is

$$V_{b2} = -\frac{gT}{nC(\gamma^2+\delta^2)^2}\left[-2\gamma+(\gamma^2+\delta^2)\,t/T + e^{-\gamma\,t/T}\left(2\gamma\cos\delta\,t/T + \frac{\gamma^2-\delta^2}{\delta}\sin\delta\,t/T\right)\right] \ldots\ldots \quad (33)$$

Differentiating this expression,

$$\frac{dV_{b2}}{dt} = -\frac{g}{nC(\gamma^2+\delta^2)}\left[1-e^{-\gamma\,t/T}\left(\cos\delta\,t/T + \gamma/\delta\sin\delta\,t/T\right)\right] \ldots\ldots\ldots \quad (34)$$

To obtain i_{c2} at the end of stage 3, we have, during that stage,

$$i_{c2} = \alpha i_{e2}$$

$$i_{c2}\,(p + 1/T) - \frac{\alpha i_{e2}}{T} = 0.$$

Initially $i_{c2} = \alpha_0\, i_7$, so

$$\overline{i_{c2}}\,(p + 1/T) - \frac{\alpha_0 \overline{i_{e2}}}{T} = \alpha_0\, i_7.$$

Substituting for $\overline{i_{e2}}$ from the equation preceding (31),

$$\overline{i_{c2}}(p+1/T) - \frac{\alpha_0\, i_7}{pT} - \frac{\alpha_0\, g}{T(\gamma^2+\delta^2)}\left[1/p + \frac{\gamma-j\delta}{2j\delta\left(p+\frac{\gamma+j\delta}{T}\right)} - \frac{\gamma+j\delta}{2j\delta\left(p+\frac{\gamma-j\delta}{T}\right)}\right] = \alpha_0\, i_7$$

which yields, after re-arranging

$$\overline{i_{c2}} = \alpha_0/p\,\left(i_7 + \frac{g}{\gamma^2+\delta^2}\right) + \frac{\alpha_0\, g}{(\gamma^2+\delta^2)(\gamma^2+\delta^2+1-2\gamma)}\left[-\frac{\gamma^2+\delta^2}{p+1/T} + \frac{(\gamma+j\delta)(\gamma+j\delta-1)}{2j\delta(p+(\gamma-j\delta)/T)} - \frac{(\gamma-j\delta)(\gamma-j\delta-1)}{2j\delta(p+(\gamma+j\delta)/T)}\right]$$

The inverse transform of this expression can be manipulated to give

$$i_{c2} = \alpha_0\left(i_7 + \frac{g}{\gamma^2+\delta^2}\right) + \frac{\alpha_0 g}{(\gamma^2+\delta^2)(\gamma^2+\delta^2+1-2\gamma)}\left[-(\gamma^2+\delta^2)e^{-t/T} + e^{-\gamma\,t/T}\left((2\gamma-1)\cos\delta\,t/T + \frac{\gamma^2-\gamma-\delta^2}{\delta}\sin\delta\,t/T\right)\right] \ldots \quad (35)$$

Using the same assumptions as were used to derive equation (32), equations (33), (34), and (35) simplify to

$$V_{b2} \sim aT/k_3\left[-4+t/T+\exp\left[-\frac{2}{(k_2+3)}\right]\left(4\cos\frac{\sqrt{k_2+1)}}{k_2+3}t/T + \frac{5-k_2}{k_2-1}\sin\frac{\sqrt{k_2-1}}{k_2+3}t/T\right)\right] \quad (36)$$

$$\frac{dV_{b2}}{dt} \sim a/k_3\left[1-\exp\left(-\frac{2}{k_2+3}\,t/T\right)\left(\cos\frac{\sqrt{k_2-1}}{k_2+3}t/T + \frac{2}{k_2-1}\sin\frac{\sqrt{k_2-1}}{k_2+3}t/T\right)\right] \quad (37)$$

and

$$i_{c2} \sim i_7 - k_2 aC + aC\left[e^{-t/T}+\exp\left(-\frac{2}{k_2+3}\,t/T\right)\left((k_2-1)\cos\frac{\sqrt{(k_2-1)}}{k_2+3} + \frac{3k_2+1}{\sqrt{(k_2-1)}}\sin\frac{\sqrt{(k_2-1)}}{k_2+3}t/T\right)\right] \quad (38)$$

Stage 4

When emitter current in X_2 ceases at the end of stage 3 the capacitance nC and the current i_7 become isolated from the rest of the circuit. The emitter of the pnp transistor thus rises in potential at a rate of i_7/nC volts/sec, or

$$V_{e2} = V_{b2}(t_2) + \frac{Ti_7}{nC}\cdot t/T \quad \ldots\ldots\ldots\ldots\ldots \quad (39)$$

where $V_{b2}(t_2)$ represents the value of V_{b2} and consequently of V_{e2} at the end of stage 3.

The potential of the base of X_2 will rise at a faster rate, and we seek to find the law governing this rise during the third stage until the stage is terminated by conduction in D_1.

We have:—

$$i_{c2} + i_{x3} + i_1 - i_{c1} - i_{x1} - i_{b2} = 0$$

i_{c1} = constant during this stage

$i_{b2} = i_{c2}$ during this stage, since $i_{e3} = 0$

$$i_{c3} = \alpha/(1-\alpha)\, i_{b3} : \; i_{b3} = aC + i_{x3} : \; i_{x3} = -pk_3 CV_{b2}$$

$$i_{x1} = pkCV_{b2} : \; \alpha = \frac{\alpha_0/T}{p+1/T}$$

Combining,

$$pCV_{b2}\left[p(k+k_3) + \frac{k_3+k(1-\alpha_0)}{T}\right] - \frac{\alpha_0 aC}{T} - (i_1 - i_{c1} - i_{c2})(p+(1-\alpha_0)/T) = 0$$

Initially, V_{b2} =, say, V_3 (from (33) or (36))

$\frac{dV_{b2}}{dt}$ =, say, V_3' (from (34) or (37))

$$i_1 - i_{c1} - i_{c2} = i_1 - i_{c1} - i_{c2}(t_2)$$

Thus $pC\,\overline{V_{b2}}\,\times$

$$\left[p(k+k_3) + \frac{k_3+k(1-\alpha_0)}{T}\right] - \frac{\alpha_0\bar{a}C}{T} - (\bar{i}_1 - \bar{i}_{c1} - \bar{i}_{c2})(p+(1-\alpha_0)/T)$$

$$= C\;(k+k_3)(pV_3+V_3') + \frac{[k_3+k(1-\alpha_0)]C\,V_3}{T} - [i_1 - i_{c1} - i_{c2}(t_2)]$$

Now $\bar{a} = a/p$, $\bar{i}_1 = i_1/p$, $\bar{i}_{c1} = i_{c1}/p$ since i_{c1} is constant throughout.

For $\overline{i_{c2}}$, we have $i_{c2} = \alpha i_{e2}$

so $i_{e2}(p + 1/T) - \alpha_0/T\; i_{e2} = 0$

Initially $i_{e2} = i_{e2}(t_2)$

hence $\overline{i_{e2}}(p + 1/T) - (\alpha_0/T)\,\overline{i_{e2}} = i_{e2}(t_3)$

$$\text{But } \overline{i_{e2}} = 0, \text{ thus } \overline{i_{c2}} = \frac{i_{c2}(t_3)}{p+(1/T)} \quad \ldots\ldots\ldots\ldots \quad (40)$$

Substituting these transforms and re-arranging,

$$\overline{V_{b2}} = (1/p)\left\{\frac{\alpha_0 i_{c2}(t_3)T}{bC} + \frac{dTV_3'}{b} + V_3 - \left[\alpha_0 a + \left(\frac{(i_1-i_{c1})(1-\alpha_0)}{C}\right)\right]\frac{dT}{b^2}\right\} +$$

$$\frac{\alpha_0 a + \frac{(i_1-i_{c1})(1-\alpha_0)}{C}}{p^2 b} + \frac{1}{p+(b/dT)}\left[\left(\alpha_0 a + \frac{(i_1-i_{c1})(1-\alpha_0)}{C}\right)\frac{dT}{b^2} - \frac{\alpha_0 i_{c2}(t_3)dT}{bC(d-b)} - \frac{dTV_3'}{b}\right] + \left(\frac{1}{p+(1/T)}\cdot\frac{\alpha_0 i_{c2}(t_3)T}{C(d-b)}\right)$$

where $b = k_3 + k(1-\alpha_0)$

and $d = k + k_3$

The inverse transform of this expression leads to

$$V_{b2}=V_3+(T/b)\left[\frac{\alpha_0 i_{e2}(t_3)}{C}+V_3'd-(d/b)\left(\alpha_0 a+\frac{(i_1-i_{e1})(1-\alpha_0)}{C}\right)\right]\left\{1-\exp\left(-\frac{b}{dT}t\right)\right\}+T/b\left(\alpha_0 a+\frac{(i_1-i_{e1})(1-\alpha_0)}{C}\right)t/T+\frac{\alpha_0 i_{e2}(t_3)T}{C(d-b)}\left\{e^{-t/T}-\exp\left(-\frac{b}{dT}t\right)\right\} \quad \ldots (41)$$

This stage terminates when V_{b2} reaches the potential to which D_1 is attached. At this time it will be necessary to know i_{x1} and i_{e2} in order to know the current in D_6.

Since $i_{x1}=pkCV_{b2}$ then by differentiating equation (41).

$$i_{x1}=k\left[1/b\,[\alpha_0 aC+(i_1-i_{e1})(1-\alpha_0)]\left\{1-\exp\left(-\frac{b}{dT}t\right)\right\}+\left\{\frac{\alpha_0 i_{e2}(t_3)}{d}+CV_3'\right\}\exp\left(-\frac{b}{dT}t\right)-\frac{\alpha_0 i_{e2}(t_3)}{d-b}\left\{e^{-t/T}-(b/d)\exp\left(-\frac{b}{dT}t\right)\right\}\right] \quad \ldots (42)$$

The inverse transform of equation (40) gives

$$i_{e2}=i_{e2}(t_3)\,e^{-t/T} \quad \ldots\ldots\ldots\ldots (43)$$

Substituting, as before, $k=2k_3$, and $\alpha^0\to 1$, equations (41) and (42) become

$$V_{b2}\sim V_3+(T/k)\left[\frac{i_{e2}(t_3)}{C}+3k_3V_3'-3a\right]\left((1-e^{-t/3T}\right)+at/T+\frac{i_{e2}(t_3)}{2C}\left(e^{-t/T}-e^{-t/3T}\right) \quad \ldots\ldots\ldots (44)$$

and

$$i_{x1}\sim 2\left[aC(1-e^{-t/3T})+\left(\frac{i_{e2}(t_3)}{3}+\frac{CV_3'}{10}\right)e^{-t/3T}-\frac{i_{e2}(t_3)}{2}\left((e^{-t/T}-\tfrac{1}{3}e^{-t/3T})\right)\right] \quad \ldots\ldots\ldots (45)$$

Stage 5

Calculation of the current in D_6 at the end of stage 3, using the result just obtained for i_{x1} and i_{e2}, shows that, when using normal component and current values, this current falls to zero at the same time that stage 4 ends. This is because i_{x1} ceases at that time and this current was, towards the end of stage 4, the major component of current in D_6. Current i_3 will now tend to turn off X_1 emitter current in a manner similar to that in which X_2 was turned off by the current i_{e3}. We require to find the time taken for i_{e1} to reduce to zero, the collector currents i_{e1} and i_{e2} flowing at that time, and the potential V_{b1} by which the base of X_1 has been reduced.

To find V_{b1}, we have

$$i_{x1}=i_3+i_{b1}-i_{e2}\ :\ i_{b1}=(1-\alpha)i_{e1}$$

$$i_{e1}=i_5+V_{b1}pC\ :\ V_{b1}=\frac{i_{x1}}{pkC}$$

from which

$$V_{b1}p\left[p+\frac{k+1-\alpha_0}{(k+1)T}\right]+\frac{i_3}{C(k+1)}(p+(1/T))+\frac{i_5}{C(k+1)}\left(p+\frac{1-\alpha_0}{T}\right)-\frac{i_{e2}}{C(k+1)}(p+(1/T))=0$$

Initially $V_{b1}=$, say, 0. $i_{e1}=\alpha_0 i_5$, $i_{e2}=i_{e2}(t_4)$,

Also, by combining the initial equations,

$$i_{x1}=\frac{k[i_3+i_5-i_{e1}-i_{e2}]}{1+k}$$

so inserting the initial values for i_{e1} and i_{e2},

$$i_{x1}=\frac{k[i_3+i_5(1-\alpha_0)-i_{e2}(t_4)]}{1+k}$$

whence $$\frac{dV_{b1}}{dt}=-\frac{k[i_3+i_5(1-\alpha_0)-i_{e2}(t_4)]}{kC(1+k)}$$

Inserting these initial values.

$$\overline{V_{b1}}\,p\left[p+\frac{k+1-\alpha_0}{(k+1)T}\right]+\frac{\bar{i}_3}{C(k+1)}(p+(1/T))+\frac{\bar{i}_5}{C(k+1)}\left(p+\frac{1-\alpha_0}{T}\right)-\frac{\bar{i}_{e2}}{C(k+1)}(p+(1/T))=-\frac{i_3+i_5(1-\alpha_0)-i_{e2}(t_4)}{C(k+1)}+\frac{i_3}{C(k+1)}+\frac{i_5}{C(k+1)}-\frac{i_{e2}(t_4)}{C(k+1)}=\frac{\alpha_0 i_5}{C(k+1)}$$

Now $\bar{i}_3=(i_3/p)$, $\bar{i}_5=(i_5/p)$, $\bar{i}_{e2}=\dfrac{i_{e2}(t_4)}{p+(1/T)}$

$$\text{Thus}\ \overline{V_{b1}}\,p^2\left[p+\frac{k+1-\alpha_0}{(k+1)T}\right]+\frac{i_3}{C(k+1)}(p+(1/T))+\frac{i_5}{C(k+1)}\left(p+\frac{1-\alpha_0}{T}\right)-\frac{i_{e2}(t_4)}{C(k+1)}p=p\frac{\alpha_0 i_5}{C(k+1)}$$

or, after re-arranging,

$$-\overline{V_{b1}}=\frac{T[\alpha_0[i_3+i_5(1-\alpha_0)]+i_{e2}(t_4)(k+1-\alpha_0)]}{C(k+1-\alpha_0)^2}\left\{\frac{1}{p+\left(\dfrac{k+1-\alpha_0}{(k+1)T}\right)}-(1/p)\right\}+\frac{i_3+i_5(1-\alpha_0)}{p^2C(k+1-\alpha_0)}$$

the transform of which is

$$-V_{b1}=\frac{T}{C(k+1-\alpha_0)}\left[[i_3+i_5(1-\alpha_0)]t/T+\frac{\alpha_0[i_3+i_5(1-\alpha_0)]+i_{e2}(t_4)(k+1-\alpha_0)}{k+1-\alpha_0}\left\{\exp\left(-\frac{k+1-\alpha_0}{(k+1)T}t\right)-1\right\}\right] \quad \ldots\ldots (46)$$

To obtain i_{e1},

$i_{e1}=i_5+C\dfrac{dV_{b1}}{dt}$, which, from equation (46) is

$$i_{e1}=i_5-\frac{i_3+i_5(1-\alpha_0)}{k+1-\alpha_0}+\frac{\alpha_0[i_3+i_5(1-\alpha_0)]+i_{e2}(t_4)(k+1-\alpha_0)}{(k+1-\alpha_0)(k+1)}\exp\left(-\frac{k+1-\alpha_0}{k+1}t/T\right)$$

$$=\frac{ki_5-i_3}{k+1-\alpha_0}+\frac{\alpha_0[i_3+i_5(1-\alpha_0)]+i_{e2}(t_4)(k+1-\alpha_0)}{(k+1-\alpha_0)(k+1)}\exp\left(-\frac{k+1-\alpha_0}{k+1}t/T\right) \quad \ldots (47)$$

To obtain i_{c1},

$$i_{c1}=\alpha i_{e1},\ \text{so}\ \bar{i}_{c1}(p+(1/T))-\frac{\alpha_0 i_{e1}}{T}=0$$

Initially $i_{c1}=\alpha_0 i_5$, so $\overline{i_{c1}}(p+(1/T))-(\overline{\alpha_0 i_{e1}}/T)=\alpha_0 i_5$

Transforming equation (47) to give $\bar{i}_{e1}$, and substituting,

$$\bar{i}_{c1}(p+(1/T))-\alpha_0/T\times\left[\frac{ki_5-i_3}{p(k+1-\alpha_0)}+\frac{\alpha_0[i_3+i_5(1-\alpha_0)]+i_{e2}(t_4)(k+1-\alpha_0)}{(k+1)(k+1-\alpha_0)\left(p+\dfrac{k+1-\alpha_0}{(k+1)T}\right)}\right]=\alpha_0 i_5$$

After re-arrangement, this can be written as

$$\bar{i}_{e1} = \frac{ki_5 - i_3}{p(k+1-\alpha_0)} - \frac{i_{e2}(t_4)}{(p+(1/T))} +$$

$$\left[\frac{i_5(1-\alpha_0)}{k+1-\alpha_0} + \frac{i_{e2}(t_4)}{\alpha_0} + \frac{i_3}{k+1-\alpha_0}\right]\left\{\frac{1}{p+\frac{k+1-\alpha_0}{(k+1)T}}\right\}$$

which upon transforming, yields,

$$i_{e1} = \frac{ki_5 - i_3}{k+1-\alpha_0} - i_{e2}(t_4)\,e^{-t/T} + \left[\frac{i_5(1-\alpha_0)}{k+1-\alpha_0} + \frac{i_{e2}(t_4)}{\alpha_0} + \frac{i_3}{k+1-\alpha_0}\right]\exp\left(-\frac{k+1-\alpha_0}{k+1}t/T\right) \quad \ldots\ldots (48)$$

The law governing i_{e2} is already known. It is

$$i_{e2} = i_{e2}(t_4)\,e^{-t/T} \quad \ldots\ldots (49)$$

As $\alpha_0 \to 1$, equations (46), (47) and (48), reduce to

$$V_{b1} \simeq (T/kC)\left[i_3(t/T) + \frac{i_3 - i_{e2}(t_4)k}{k}\left(\exp\left(-\frac{k}{k+1}t/T\right) - 1\right)\right] \quad \ldots\ldots (50)$$

$$i_{e1} \simeq \frac{ki_5 - i_3}{k} + \frac{i_3 + i_{e2}(t_4)k}{k(k+1)}\exp\left(-\frac{k}{k+1}t/T\right) \quad \ldots\ldots (51)$$

and

$$i_{e1} \simeq \frac{ki_5 - i_3}{k} - i_{e2}(t_4)\,e^{-t/T} + (i_{e2} + i_3/k)\exp\left(-\frac{k}{k+1}t/T\right) \quad \ldots\ldots (52)$$

Stage 6

At the end of stage 5, the emitter current of X_1 will have been reduced to zero. X_3 collector current will have a value $i_{e1}(t_5)$ and X_1 collector current a value $i_{e2}(t_5)$. These currents may be calculated from equations (48) (or (52)) and (49) after substituting the value of t/T at which i_{e1} reaches zero from equation (47) or (51). The potential of the base and emitter of X_1 will be at $-V_{b1}(t_5)$ which may be similarly obtained from equation (46) or (50).

During stage 6 the potential of the base of X_1 continues to fall until it is arrested by D_3, and that of the emitter until i_5 is diverted into D_5.

For the base potential we have

$$V_{b1} = -\frac{i_{x1}}{pkC}$$

$$i_{x1} = i_3 - i_{e1} - i_{e2}$$

Thus $V_{b1}p + (i_3/kC) - (i_{e1}/kC) - (i_{e2}/kC) = 0$

Initially $V_{b1} = -V_{b1}(t_5)$, so

$$\overline{V_{b1}}\,p + \frac{\bar{i}_3 - \bar{i}_{e1} - \bar{i}_{e2}}{kC} = -V_{b1}(t_5)$$

Now $\bar{i}_3 = (i_3/p)$, $\bar{i}_{e1} = \frac{i_{e1}(t_5)}{p+(1/T)}$, $\bar{i}_{e2} = \frac{i_{e2}(t_5)}{p+(1/T)}$

Substituting, $\overline{V_{b1}}p^2(p+(1/T)) + \frac{i_3}{kC}(p+(1/T)) - \frac{(i_{e1}+i_{e2})(t_5)}{kC}p = -V_{b1}(t_5)\,p\,(p+(1/T))$

or $\overline{V_{b1}} = -\frac{V_{b1}(t_5)}{p} - \frac{i_3}{p^2kC} + \frac{(i_{e1}+i_{e2})(t_5)}{kC}\left(1/p - \frac{1}{p+(1/T)}\right)$

Transforming,

$$V_{b1} = -V_{b1}(t_5) - \frac{T}{kC}\left[i_3\,t/T - (i_{e1}+i_{e2})(t_5)\,(1-e^{-t/T})\right] \quad \ldots\ldots (53)$$

For the emitter potential we have

$$V_{e1} = -(i_5/pC), \text{ or } V_{e1}p + (i_5/C) = 0$$

Initially $V_{e1} = -V_{b1}(t_5)$, so $\overline{V_{e1}p} + (\bar{i}_5/C) = -V_{b1}(t_5)$

Now $\bar{i}_5 = (i_5/p)$, thus $\overline{V_{e1}}\,p^2 + (i_5/C) = -pV_{b1}(t_5)$

or $\overline{V_{e1}} = -\frac{V_{b1}(t_5)}{p} - i_5/p^2C$, which, upon transformation, gives $V_{e1} = -V_{b1}(t_5) - (Ti_5/C)\,t/T$ ……………… (54)

Stage 7

At the end of stage 4 the collector current of X_3 will have attained a value of, say, $i_{c3}(t_4)$. During stages 5 and 6 this current will continue to increase and this process will go on until some time, say t_6, when the input voltage ramp is terminated. The turn-off transistor will then itself be turned off by the current i_9 of Fig. B.1.

We need to know the value of i_{c3} at time t_6 and the law of decay of current subsequent to that time.

We have: $i_{c3} = \frac{\alpha}{1-\alpha}\,i_{b3}$

Since the collector of X_3 is clamped by D_2 after the end of stage 4, k_3C plays no part, so $i_{b3} = aC$.

Hence $i_{c3}\left(p + \frac{1-\alpha_0}{T}\right) - (\alpha_0/T)\,aC = 0$

Initially $i_{c3} = i_{c3}(t_4) - i_{x3}(t_4) + i_{e1} + i_{x1}(t_4) + i_{e2}(t_4)$

where $i_{x1}(t_4)$ is given by equation (42), $i_{e2}(t_4)$ by (43), i_{e1} is equal to $\alpha_0 i_5$ and $i_{x3}(t_4) = (k_1/k)\,i_{x1}(t_4)$

Thus $\bar{i}_{c3}\left(p + \frac{1-\alpha_0}{T}\right) - (\alpha_0/T)\,\bar{a}C = i_{c3}(t_4)$

Now $\bar{a} = (a/p)$, so

$$\bar{i}_{c3} = \frac{p\,i_{c3}(t_4) + (\alpha_0/T)aC}{p\left(p+\frac{1-\alpha_0}{T}\right)}$$

Whence

$$i_{c3} = i_{c3}(t_4)\exp\left(-\frac{1-\alpha_0}{T}t\right) + \frac{\alpha_0 aC}{1-\alpha_0}\left(1 - \exp\left(-\frac{1-\alpha_0}{T}t\right)\right) \quad \ldots\ldots (55)$$

Let the value of i_{c3} at t_6 be written as $i_{c3}(t_6)$

Then, as before,

$$i_{c3}\left(p + \frac{1-\alpha_0}{T}\right) - (\alpha_0/T)aC = 0$$

But as at t_6 a reduces to zero, and we now consider the small current i_9,

$$i_{c3}\left(p + \frac{1-\alpha_0}{T}\right) + (\alpha_0/T)i_9 = 0$$

Since initially $i_{c3} = i_{c3}(t_6)$,

$$\bar{i}_{c3}\left(p + \frac{1-\alpha_0}{T}\right) + (\alpha_0/T)\,\bar{i}_9 = i_{c3}(t_6)$$

Now $\bar{i}_9 = (i_9/p)$, so $\bar{i}_{c3} = \frac{p\,i_{c3}(t_6) - (\alpha_0/T)\,i_9}{p\left(p + \frac{1-\alpha_0}{T}\right)}$,

whence $i_{c3} = i_{c3}(t_6)\exp\left(-\frac{1-\alpha_0}{T}t\right) - \frac{\alpha_0 i_9}{1-\alpha_0}\left(1 - \exp\left(-\frac{1-\alpha_0}{T}t\right)\right)$ ……………… (56)

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The Design of Function Generators Using Silicon Carbide Non-linear Resistors

By E. Brown*, B.Sc., and P. M. Walker*, B.Sc.

Silicon carbide resistors, the effective current-voltage characteristics of which have been modified by series and parallel linear resistors, are employed as input or feedback components of an operational amplifier to obtain non-linear function generators. The design of square law and sine units having a high accuracy is described in detail.

THE basic techniques of electronic analogue computers have been described in a recent series of articles[1,2]. The input grid of a high gain amplifier to which feedback is applied, as shown in Fig. 1, may be regarded as a virtual earth; and if the grid current is negligible, then

$$V_2/V_1 = -(R_2/R_1) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

and

$$V_2 = -I_1R_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Thus, when R_1 and R_2 are linear resistors, scaling by a constant factor is achieved. The use of more than one input terminal and resistor makes possible the addition of several input voltages.

It is frequently necessary in analogue computer work to evaluate one variable (V_2) which is a non-linear function of another (V_1). One way of achieving this result is to replace either R_1 or R_2 by a network of resistors and biased diodes[3]. The output voltage of such a unit takes the form of a series of straight lines which approximate to the required curve. This type of function generator will be unsuitable in applications where the derivative of the output is required, since the derivative will be discontinuous. To obtain high accuracy many diodes are required, and the function generator units in consequence become bulky.

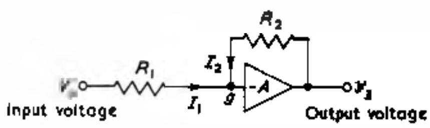

Fig. 1. Computer amplifier with feedback network

This article describes a simpler and more compact type of function generator, using silicon-carbide non-linear resistance elements, which do not possess the above disadvantages.

Theory

It can be seen from Fig. 1 and equation (2) that the problem of obtaining an output voltage V_2 which is the required non-linear function of the input voltage V_1 is essentially that of producing a network to replace R_1 for which I_1 is the required function of V_1. The modification of the current-voltage characteristic of a silicon carbide resistor is the basis of the present method[4].

The current-voltage relationship of silicon-carbide may be represented by

$$I = kV^{\alpha} \quad \ldots\ldots\ldots\ldots \quad (3)$$

At high voltages the value of α is reasonably constant but falls as the voltage decreases. The resistors used are in disk form, supplied by Morgan Crucible Company Limited, and the Automatic Telephone Company Limited. Three different types have been used and are shown in Fig. 2. For convenience they are designated D_1, D_2 and D_3; the characteristics of these disks are given in Table 1, and current-voltage curves are shown in Fig. 3. In Fig. 4 the effect of series and parallel linear resistance on the current-voltage curve for a type D_1 disk is illustrated. It will be observed that, in the case of the series linear resistance, as the voltage is increased the index of the effective power law varies from that typical of the disk to unity; for the parallel resistance, this change of power law occurs in the reverse direction for increasing voltage. It will be appreciated therefore that combined series and parallel resistances will enable a wide range of power laws to be obtained between given voltage limits.

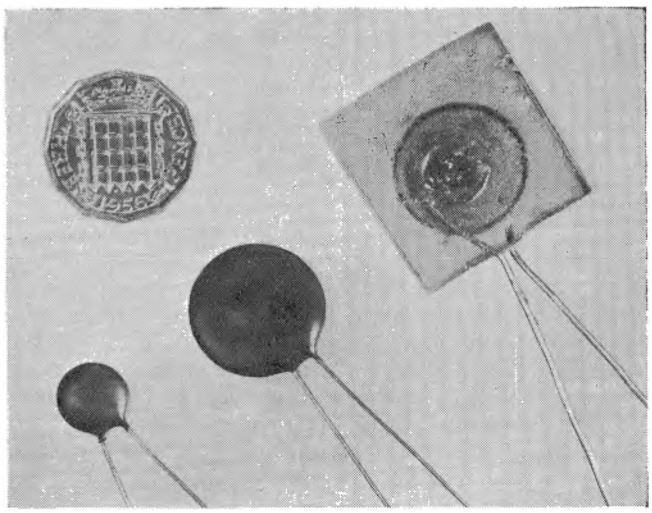

Fig. 2. Non-linear resistors

Trigonometrical functions may be expressed as a power series, of which the first term is linear or a constant. The higher order terms may be obtained by means of a non-linear circuit, and the dominant terms of such series expansions are the cube and the square. The square law is of particular importance since it also arises in certain geometrical relationships and physical laws; further a multiplier can be made based on square law generators.

TABLE 1

	RESISTANCE (at 50V)	k (at 50V)	α (at 50V)	α (at 5V)
Disk type D_1	100kΩ	$8{\cdot}4 . 10^{-10}$	3·4	2·0
Disk type D_2	75kΩ	$3{\cdot}2 . 10^{-11}$	4·3	3·0
Disk type D_3	70kΩ	$2{\cdot}4 . 10^{-12}$	5·0	3·5

* *The General Electric Company Limited, Applied Electronics Laboratories, Stanmore.*

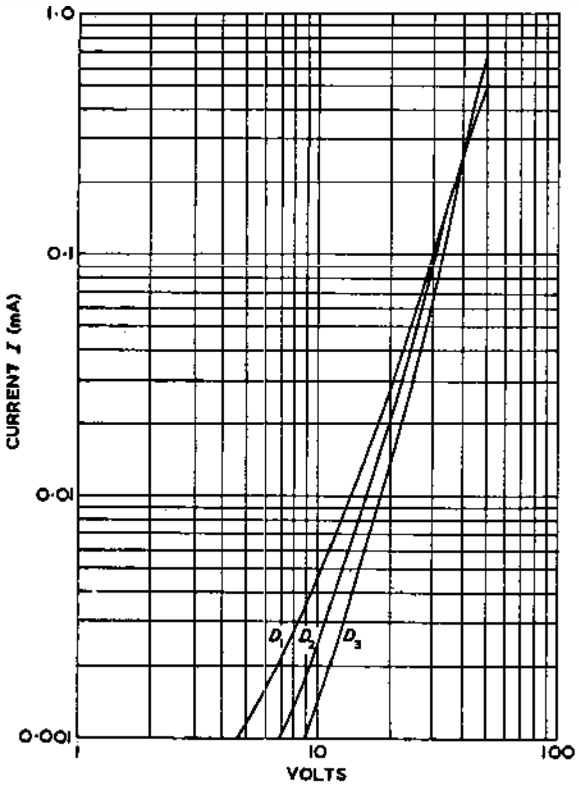

Fig. 3. Current voltage curves for type D_1 D_2 and D_3 non-linear resistors

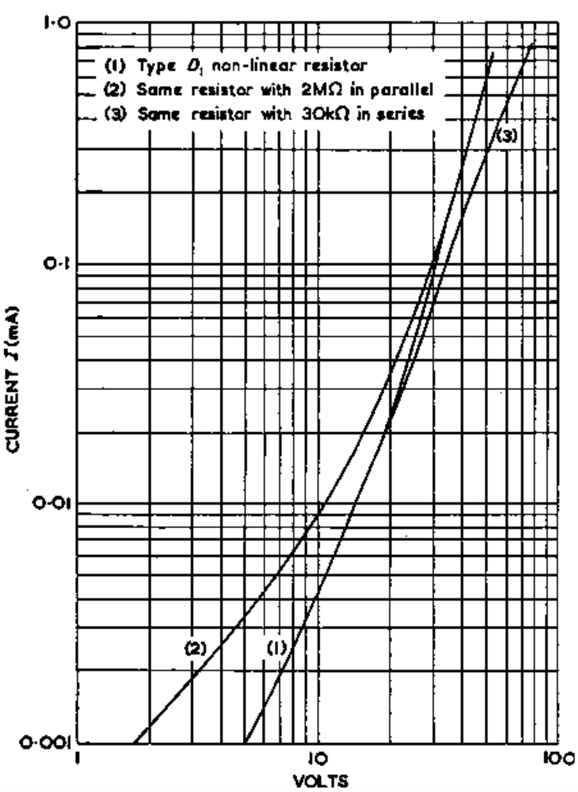

Fig. 4. Current voltage curves for a non-linear resistor with series and parallel linear resistors

A Square Law Function Generator

A first approximation to a square law characteristic is achieved with an input network consisting of one non-linear resistor in series with a linear resistance. A type D_1 disk is most suitable since α for this disk is closest to two. As α is dependent on the voltage gradient in the disk, a further degree of freedom can be obtained by feeding the disk from a potential divider, so that the operating voltage range of the disk may be changed independently of the input scale. It is also convenient to be able to adjust the scale of the output voltage without altering the feedback resistor. This necessitates a current divider. A circuit with both of these facilities is shown in Fig. 5; Fig. 6 depicts two typical

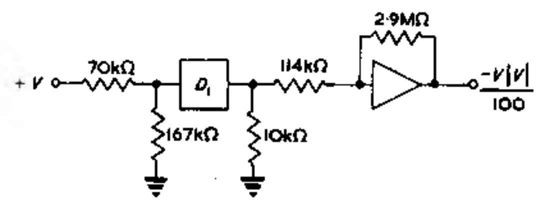

Fig. 5. Simple square law function generator showing typical component values

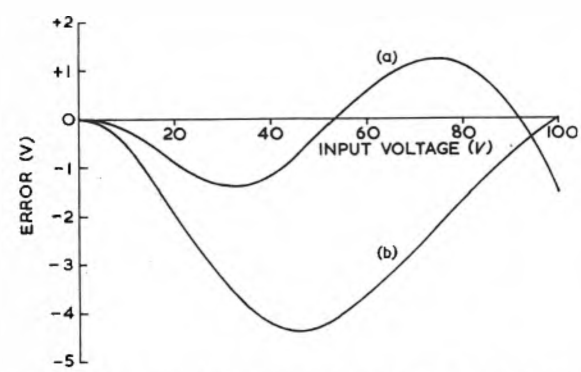

Fig. 6. Typical errors curves for simple square law function generator

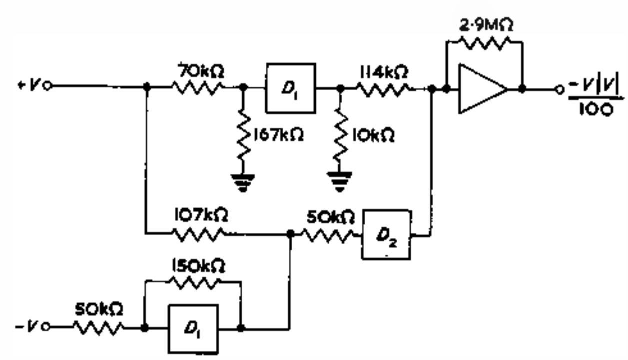

Fig. 7. Square law function generator and correction circuit, with typical component values

error curves for this type of circuit. Fig. 6(a) shows the closest approximation to a square law that can be obtained. All the units described are scaled so that an input of 100V gives rise to 100V output.

Several types of circuit have been investigated; the most accurate type developed sums the output current from the network of Fig. 5 and the current from a correction circuit. In this case a basic unit is built which has an error curve as shown in Fig. 6(b), since this form of error is most easily corrected.

The circuit of such a unit is shown in Fig. 7: it can be seen that an input voltage of $\pm V$ is required, but if the unit is required to operate in two quadrants, this is no disadvantage. The correction circuit consists essentially of a potential divider chain between V_1 and $-V$. The lower arm contains a non-linear disk, type D_1, and so its resistance decreases with applied voltage. Initially therefore the potential at the centre point of the chain will increase positively with V. As V rises further still the voltage will drop, until when the resistances of both arms are equal, the voltage will be zero. The second disk, type D_2, serves further to shape the correction curve, and ensures zero slope at the origin. A typical error curve for such a square law unit is shown in Fig. 8; it should be noted that the error is small near the origin.

The units described generate $-V|V|$. If units are required to operate as square law generators in two quadrants, two diode switches must be inserted which operate at the

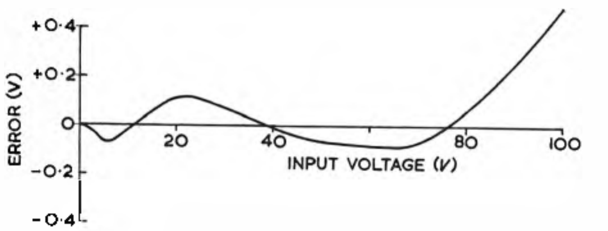

Fig. 8. Error curves for square law function generator shown in Fig. 7

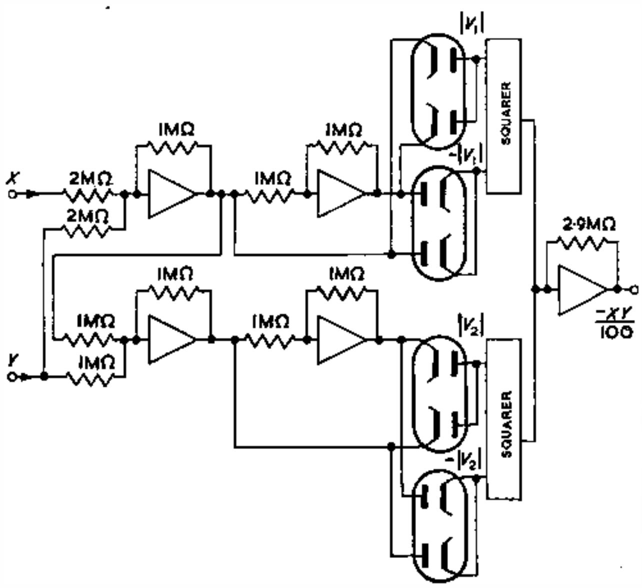

Fig. 9. Quarter-square multiplier

origin. In adjusting the circuit values, the back voltage and impedance of the diodes need to be taken into account while, to ensure smooth switching, selection of pairs of diodes with low and equal back voltage is necessary. The back voltage can be further reduced by applying a bias current to the diodes.

The setting up of square law units can be facilitated by means of a test unit consisting of two ganged precision potentiometers. A fixed voltage is applied to the winding of the first potentiometer the wiper of which is taken to a unity gain amplifier. The amplifier output is applied to the second potentiometer. A further unity gain amplifier gives an output proportional to V^2, where V is the output from the first amplifier. Using potentiometers of accuracy better than 0·1 per cent, and taking care over their loading, a test unit of accuracy better than 0·03 per cent of full scale has been made.

The identity

$$\left|\frac{X+Y}{2}\right|^2 - \left|\frac{X-Y}{2}\right|^2 \equiv XY$$

shows how two square law function generators may be used to make a multiplier. Fig. 9 gives the circuit diagram of

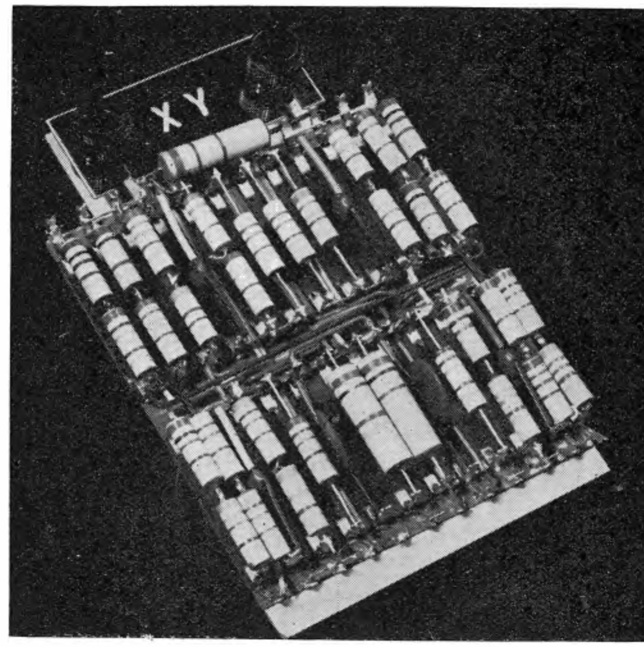

Fig. 10. Quarter-square multiplier plate

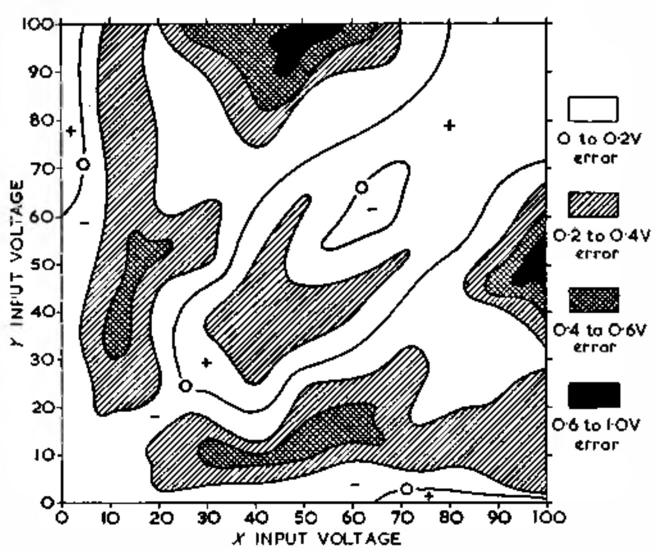

Fig. 11. Quarter-square multiplier error curves contour diagram

such a multiplier, in which the blocks marked 'squarer' consist of the input circuit of Fig. 7. The resistors and non-linear elements associated with a complete multiplier are shown mounted on a plate in Fig. 10. The accuracy of a multiplier may be deduced by supposing that the errors in each square in the above expressions are ϵ_1 and ϵ_2 respectively, then the output is $XY + \epsilon_1 - \epsilon_2$ instead of XY. The quantity $(\epsilon_1 - \epsilon_2)$ is the error of the multiplier and for this to be small, the squarer errors need to be as small as possible. Of particular interest is the case when X or Y

is zero. The two squarers will then have the same inputs and so, if exactly matched, ϵ_1 and ϵ_2 will be the same and cancel. Matching the two squarers therefore reduces the error when X or Y is small.

With the unit described above, an overall accuracy of better than 1·0V can be obtained and, over most of the range, the accuracy is considerably better; for X or Y zero, the maximum output is 0·1V. Fig. 11 shows contours giving lines of constant error, the contours being in 0·2V steps.

A multiplier may be connected in a feedback circuit to provide a divider. A quantity Z is multiplied by one input X, and the product $-XZ$ is equated to a second input Y in a high gain amplifier. Hence $Z = -(Y/X)$. If errors ϵ_1 and ϵ_2 are assumed for the squares of

$$\left|\frac{X + Z}{2}\right| \text{ and } \left|\frac{Y - Z}{2}\right|$$

respectively, then the quantity $\epsilon_1 - \epsilon_2$ is equivalent to an error in Y of this magnitude. This enables the accuracy of the divider to be readily assessed.

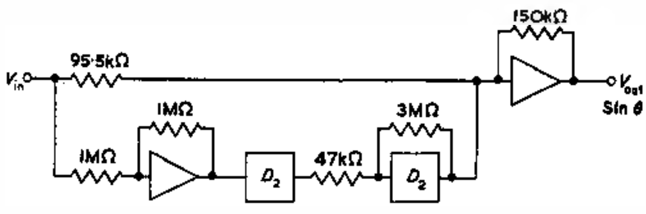

Fig. 12. Sine function generator with typical component values

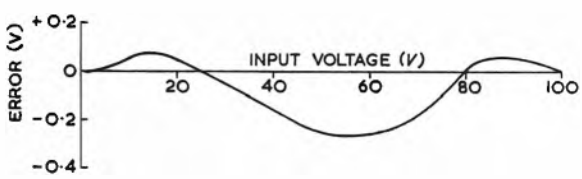

Fig. 13. Typical error curve for sine function generator

Trigonometrical Function Generators

For trigonometrical functions a non-linear circuit is used to obtain the higher order terms of a series expansion. In the case of a sine unit

$$\sin\theta = \theta - (\theta^3/3!) + (\theta^5/5!) - + \ldots.$$

The linear term is readily obtained, and the dominant term in the remainder is the cubic one. For this reason type D_2 disks are employed, the best result being obtained from a circuit using two disks, as shown in Fig. 12. A typical error curve is shown in Fig. 13. The normal full scale output voltage of 100V is used for all trigonometrical units, and for the sine the peak error is approximately 0·3V for a unit with range $(-\pi/2) \leqslant \theta \leqslant (\pi/2)$.

A sine unit may be converted into a cosine unit quite easily, by forming $|(\pi/2) - \theta|$, but it is more economical in amplifiers to design a special cosine unit. A circuit similar to that of the sine unit using type D_1 disks gives good results for the non-linear term. For an input of $\pm(\pi/2)$ an accuracy of 0·3V is obtained, and if the range is restricted to $\pm(\pi/3)$, this can be improved to 0·1V in 100V output.

The tangent function requires a disk of higher power law than the sine. A tangent unit has been constructed obtaining the non-linear terms of the expansion from a type D_3 disk with a single series resistance. In the range $\pm 50°$, it is accurate to 0·2V. A function generator for the secant has been constructed, using similar principles.

In each case the interchange of input and feedback networks gives rise to the inverse function. A satisfactory performance has been obtained from square root, inverse tangent and inverse sine units.

Accuracy and Stability

The figures quoted show that several useful functions may be generated with an accuracy of 0·3 per cent of full scale. The frequency response of the elements does not impair the accuracy of the function generators within the bandwidth of the amplifiers used, and it is believed that satisfactory operation up to 20kc/s is possible.

The effect of temperature on the current-voltage relation of silicon carbide is given by

$$V = C_0(1 + \gamma T) I^{1/\alpha}$$

where γ lies between $-0·0012/°C$ and $-0·0018/°C$. When a series resistor is present, the change of current with temperature alters the voltage across the disk in such a way as to maintain the original current. The overall temperature coefficient of the units described is therefore very much less than is suggested by that of the disk, and normal changes in laboratory temperature have have been found to have only a small effect on the accuracy of the units described.

The long term stability of the units is good. A small adjustment in the scaling of non-linear terms has been occasionally necessary after the units have been in use several months. It should be pointed out that with some forms of silicon-carbide disks an increase in resistance with time takes place on application of a constant voltage. This is initially quite rapid and may be as large as 2 per cent after one minute, but increases to only 5 per cent after several hours. The change is in the opposite direction to that expected from coulomb heating, and is reversible, but only occurs to a significant degree above certain voltage levels, and only for disks D_3.

The characteristics of silicon-carbide are affected by humidity, and those disks which were not already sealed have been dried, and potted in araldite. Since silicon-carbide resistors are normally manufactured for application as spark and surge suppressors where tolerances are large, each function generator must be adjusted individually and the component values given in the circuit diagrams are only typical. In the present application a more stringent tolerance is of considerable help in production of identical units. It is suggested that it is useful to quote a tolerance on the conductivity k at a particular applied voltage which should be within the operating range of the voltage across the disks in the particular circuit.

In conclusion it can be seen from the foregoing discussion that simple, compact and inexpensive function generators can be made using silicon carbide resistors. A wide range of functions can be obtained to good accuracy, which is maintained over long periods without adjustment.

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BOOK REVIEWS

Three Steps to Victory

By Sir Robert Watson-Watt. 474 pp. Large 8vo. Udhams Press Ltd. 1958. Price 30s.

THERE is much of interest and value in this book, but unfortunately, there is also much which is intolerably boring and, long before the reviewer had got half-way through its pages, he had begun heartily to regret that he had ever undertaken to review it. The book is at once an autobiography of its author and an account of the genesis of radar in 1935, its development during the succeeding years and its many applications throughout the war.

The 'three steps' were respectively the invention of the cathode-ray direction finder, the development of radar and the institution of systematic Operational Research but much of the story of these great achievements is written in a style so pompous, so pedantic and so full of involved circumlocutions as to become most tedious. Much of the matter is irrelevant and unnecessary, much of the text can only be described as exuberant verbosity. Many of the sentences are not only far too long but they are broken up by such a prolixity of qualifying adjectives and phrases and by such frequent use of explanatory parentheses as to leave the reader in doubt as to which is the main subject verb and object until he has read the sentence two or three times—and even then the meaning is often obscure. For example, on page 133:—

"Hence our insistence on 'supplementary and not integral'—integral in the special sense that the emergency system must be something that would work after a fashion, providing such information as could be obtained (a) without a direction finder, and (b) without the particular high-speed computer (which we visualised by requirement but not by a design to meet the requirement) that would eliminate the effect of mis-association in range-cutting; this it would do by finding such solutions as were compatible from computations based on the changes in range between successive observations spaced closely in time."

Apart from the exaggerated verbiage which mars so much of this book, the reader's sentiments are badly shaken every few pages by the unashamed and boastful egotism of its author. Why, oh why, need we be told that one of his schoolteachers had told him (in 1938!) that he had been "her most brilliant scholar ever" or that she had recommended him for a special prize for "general excellence"? And why need we be irritated by stupid and puerile puns on almost every alternate page? It may amuse a childish mind when reading, for example, the chapter dealing with the need for higher powers on centimetre wavelengths to note the title of the chapter—"Wanted! Watts on Watts!" [*"Watts 'n' watts" would have been funnier! Ed.*]—but it is little short of infuriating to the adult reader.

The second half of the book is far more interesting and the style far less irritating than the first. Sir Robert's factual account of the early days of radar, the difficulties, administrative as well as technical, with which he and his team had so often to contend, his descriptions of the brilliant successes of radar and the true background to one or two of the tragic failures (e.g. Pearl Harbour and the "Scharnhorst"/"Gneisenau" incident) all form a valuable contribution to the history of our times. To Sir Robert Watson-Watt, more than any other single man, we owe the winning of the Battle of Britain; few will deny after reading this book, written though it is by his own hand, the tremendous part he played in the final winning of the war.

There are many lessons to be learned from this story but their learning would have been far easier—and far more certain—if only some experienced journalist, armed with a plentiful supply of blue pencils, had been commissioned to eliminate the verbiage, the irrelevant, and to reduce by half the length of the text.

G. R. M. GARRATT

Symposium on Vacuum Technology 1956

Edited by E. S. Perry and J. H. Durant. 257 pp. 50 figs. Royal 8vo. Pergamon Press Ltd. 1957. Price £4 10s.

ONE cannot but envy the lively interchange of information between pure and applied physicists at the many specialist conferences in America. In many fields these specialist conferences are annual events and the conference reports published by the American Institute of Physics show how wide is the range. Of course, many conferences on specialists' subjects are convened here, but one gets the impression that the Americans place a greater emphasis on the conference as a medium for developing a subject, particularly in technology. Understandably, its greater population permits the organizing of conferences devoted to topics which would not be practical here. Also, one must recognize the adverse effects that military secrecy must have had on the free interchange of knowledge generally, and that it has apparently become difficult for American scientists to organize *International* conference in their country. This apart, the reviewer believes that an American technologist has a better opportunity of learning and submitting his work to criticism under conditions of live debate than many of his counterparts here; this undoubtedly is the situation in vacuum engineering.

In 1953 a committee on vacuum techniques came into being in the U.S.A. "to sponsor symposia on vacuum technology and to assist in the dissemination of related knowledge." Since then, three annual symposiums have been held and the contributed papers published. The work reviewed here is for the 1956 conference at which some forty papers were read and grouped in the published material as follows: — Fundamental Developments in Vacuum Technology; Methods and Techniques for obtaining High Vacuum; Instrumentation, Controls and other Vacuum Devices, Vacuum Distillation; Metallurgical and Chemical Applications; Technical Subjects of Current Interest.

Such a wide range of subjects sets a task beyond the critical capacity of the reviewer, who will confine himself to discussing a few of the papers and the general presentation of the reports.

One or two papers set a high standard. For example on the theoretical side that by Dayton on 'Gas Flow Patterns at Entrance and Exit of Cylindrical Tubes' develops earlier work by Clausing, and that by Cooke on 'An Inherent Error in the Knudsen Effusion Manometer and Method of Correction' in which it is shown that the effusion method only gives a correct value for the vapour pressure when the diameter of the effusion hole is reduced to zero, a result one can obtain by extrapolating data obtained with different hole diameters. One can see here the correctness of the idea that the means of measurement interferes, however imperceptibly, with the normal equilibrium of the system being studied, but as Cooke shows, many workers using different size effusion holes have attributed the effect to experimental error and plotted their results to obtain a mean value. Perhaps it was an oversight that made him use the symbol 'u' for viscosity in a book on vacuum technique!

In the papers on vacuum pumps it is of interest to read of the experiences of Gale using a titanium getter/ion pump for exhausting X-ray tubes, and Gould's pioneering intention to use larger, but similar, pumps on the particle accelerator half a mile in diameter at Brookhaven. Gale's paper is marred by over condensed writing, but the effect he and other workers have observed on the increased pumping rate of titanium in the presence of ionized gases differs with the results obtained by Wagner for barium, and there is an interesting study of the effects of an ionized atmosphere on the pumping rate of getters awaiting the attention of a post graduate worker. Attention is given to mechanical boosters (Rootes type blowers) by German and American workers, and to the steam and water ejector by Blatchley. Such pumps are intended for operation at intermediate pressure and are used in large scale industrial processes. However, there is no discussion of the vapour booster pump, which using conventional vacuum fluids may have an exceptionally high backing pressure and provide high pumping speeds from 2mm down to 10^{-4}mm Hg. These pumps have been well developed in this country, and because of their pumping range have proved valuable for metallurgical purposes such as melting and casting in vacuum. The analysis by Hickman and Kinsella of the factors influencing the design of a diffusion pump is of interest, and one wonders whether the resultant pump with its novel boiler could possibly be the final and ultimate form.

Of interest in the section on instrument design is the brief description of the 'acoustical' vacuum gauge by Schwarz. Such gauges depend on the energy lost by a vibrator to the surrounding gas which is a function of pressure. Schwarz does not deal with the theory of his gauge which, depending on the pressure range, may lose energy by direct molecular impulse of viscous flow (an interesting theoretical paper on the acoustic absorption coefficient of gas has been recently published by Edmond & Lamb in *Proc. Phys. Soc.*, *71*, 1958, 17).

Nerken's article on the flow of gases through leaks is of practical value in showing the range of tube dimensions and pressure over which the laws of either molecular or viscous flow apply, but it could have been improved by relating the results to the general expression developed by Guthrie and Wakerling and Oehert and Steckelmacher which was not mentioned.

The article by Preuss on evaporation of radioactive materials requires the insertion of a square root sign in the formula for evaporation of alloys on page 145.

An interesting paper showing the growing relation between the work of the vacuum engineer and high altitude flight is that by Maslach on 'Design and Operation of a Hypersonic Low Density Wind Tunnel'. He states that "the development of low density wind tunnels for the simulation of high speed, high altitude flights is probably one of the least publicized use of modern vacuum technology". He describes the tunnel at Berkeley in which a pressure of 85 microns Hg. (40 miles equivalent altitude) can be obtained with an air current of 3in diameter, streaming at a velocity of 2 380ft/sec at a temperature of 65°K.

Now to a general view of the collected papers. The editors have obviously been accommodating with some 25 per cent of the contributions and accepted work which should have been more thoroughly edited. This charge refers to many papers from workers who obviously have not the harassed existence of the industrial scientist and thus, one can only believe, did not take the publication as serious as its motives warranted. Undoubtedly, the common language is becoming more than ever cleaved by new idiomatic usages. Surely, however, there must be bad American as there is bad English? Thus one learns that equations, like drinks, are 'set up' and that an experimental result may "... show dramatically ... and points up the importance". These criticisms apart, the committee on Vacuum Technology is doing valuable work, and they should consider publishing their symposia in less costly format, so that it is within the reach of workers who want to be informed of general advances.

L. HOLLAND

Industrial Electronics

By A. W. Kramer. 311 pp. 60 figs. Royal 8vo. Sir Isaac Pitman & Sons Ltd. 1958. Price 36s.

THE material for this book has been drawn from many sources; from the Transactions and Proceedings of technical societies such as the American Institute of Electrical Engineers and the Institute of Radio Engineers from articles in electronic and radio magazines from manufacturers' data and from various other sources.

It is primarily a book for those who have a good knowledge of general physics and engineering but who have had very little training or experience in electronics. To a certain extent it is a beginner's book. It explains, step by step, in simple and concise language and without the use of mathematics, what electron tubes are, how they work, how they evolved and how they can be used.

Process Instruments and Controls Handbook

Edited by M. Considine. 552 pp. 250 figs. Royal 8vo. McGraw-Hill Publishing Co. Ltd, New York and London. 1957. Price £7 6s. 6d.

TAKEN in its entirety, this book gives a comprehensive and detailed compilation of descriptions and data relating to instrumentation and control. It gives numerous examples of application to show how the devices and principles described can be used in a variety of industries.

Among the subjects covered are measurement standards, primary elements, measurement systems, indicators and recorders, automatic controllers, timers, electric and pneumatic telemetering, final control elements, fundamental principles of process control, and mathematical techniques for solving automatic control problems.

Kempe's Engineers' Year Book 1958

3 000 pp. Crown 8vo. (Two volumes in case.) 63rd edition. Morgan Bros (Publishers) Ltd. Price 82s. 6d.

AMONG the alterations and additions to this edition are those under the headings of 'Railway Engineering', 'Aerodynamics' 'Aircraft Propulsion', 'Electronic Engineering', and 'Atomic Power'. Every section and sub-section of the book has received the attention of a specialist in that field, and details of established developments in modern practice are included.

The 79 chapters of these volumes cover practically every branch of engineering.

ELECTRONIC EQUIPMENT

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

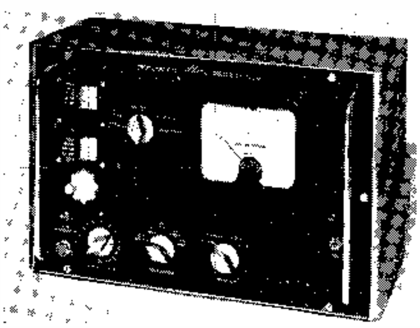
A.M.-F.M. MODULATOR METER

(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire

THE modulation meter type 210 may be used to measure the percentage modulation of amplitude modulated signals and the peak deviation of frequency modulated signals in the carrier frequency range 2·25Mc/s to 300Mc/s. Operation with reduced sensitivity is possible up to a frequency of 600Mc/s. Modulation depths up to 100 per cent and frequency deviations up to ±100kc/s, may be measured in the modulation frequency range 30c/s to 10kc/s.

Outputs at the intermediate frequency of 750kc/s and at low frequency are available from terminals on the front panel. These outputs enable the modulated envelope of the input signal and



the demodulated signals to be observed on an oscilloscope.

The limiting action of the instrument is so effective that it can be used to measure spurious frequency modulation occuring on amplitude modulated signals. Furthermore, changes of mean carrier level when amplitude modulation is applied can be measured to an accuracy of better than ±1 per cent.

One of the most attractive features of the instrument is its simplicity of operation. The tuning control is adjusted until a meter reading is obtained, and the input attenuator adjusted for full scale deflexion. It is then only necessary to switch to the a.m. or one of the f.m. positions to obtain a direct reading of modulation.

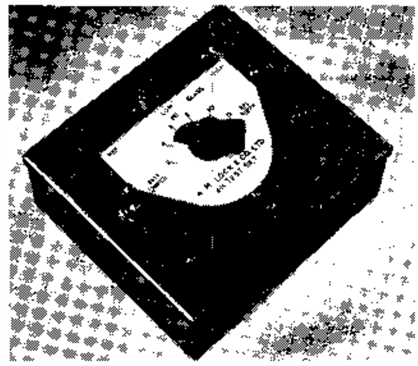
pH METER TEST SET

(Illustrated above right)

A. M. Lock & Co. Ltd, Prudential Buildings, 79 Union Street, Oldham, Lancashire

THE Lock universal pH meter test set is a simple instrument for quickly checking and localizing faults on pH meters, and on pH electrode systems and for assisting in the checking and setting up of industrial pH measuring, recording and control gear.

This pH test set consists basically of a long life dry cell across which is connected a number of high stability resistors forming a potentiometer. These resis-



tors connect to a multi-position selector switch which selects an increasing voltage as it is rotated in a clockwise direction, each step being equal to approximately to 2pH units.

A number of interconnecting leads are provided for connecting to the pH meter under test. Two output connexions to the pH meter are provided, a low impedance one, and a high impedance one of approximately 250MΩ to simulate the high resistance of normal glass electrodes.

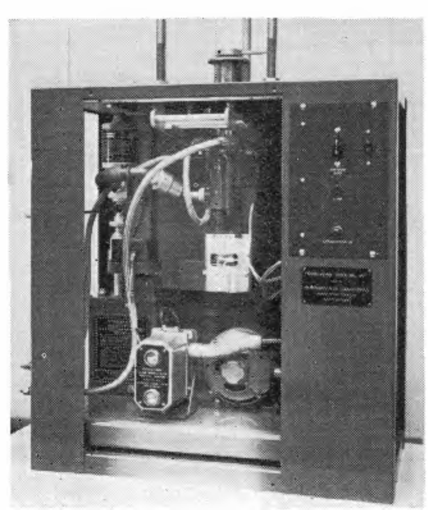
C.R.T. RE-PUMPING EQUIPMENT

(Illustrated below)

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

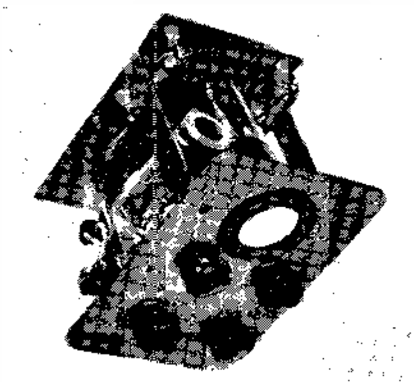
THERE is a considerable interest throughout the country in reconditioning of cathode-ray tubes for subsequent resale, and many electrical companies are successfully undertaking this type of work. One of the most important processes is, of course, the actual repumping of the tube and Edwards are manufacturing a special pumping unit for this purpose.

This outfit consists of a "Speedivac" 1S50 rotary vacuum pump displacing approximately 1·7ft^3/min, backing a type 203 oil diffusion pump capable of attaining a final vacuum of 5×10^{-6}mm Hg



The diffusion pump is fitted with a thermosnap switch for protection if the cooling water supplies should fail and a fluid economizer is mounted in the lacking line to reduce the amount of oil lost from the rest of the system, thus enabling the system to be left under vacuum when the unit is switched off. Other safety switches and non return valves are supplied for the complete protection of the plant in the event of power failure.

The top surface of the cabinet is made from heat resistant material and the neck gripper of the cathode ray tube is water cooled so that the tube may be baked in an oven prior to sealing. The complete plant is entirely self-contained and occupies a floor space of approximately 16in × 28in the cabinet being 30in high.



MINIATURE MONITOR OSCILLOSCOPE

(Illustrated above)

Bourne Engineering Co. Ltd, Abbey Mill, St. Albans, Hertfordshire

THE Uniscope has been designed as a building 'brick' for monitoring waveforms, noise or any other suitable parameters in electronic equipments.

It is sufficiently small to be built into even the smaller panels without absorbing more than a minor part of the available space, and is sufficiently cheap to be used in all important parts of a large installation. By this means, most convenient monitoring facilities may be provided completely independently of external apparatus.

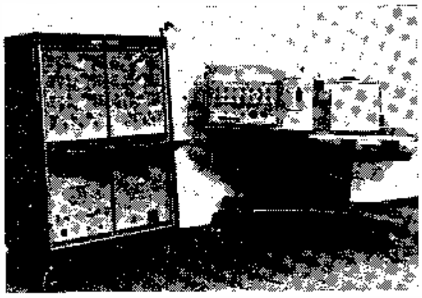
The Y amplifier sensitivity is up to 1V/cm with a bandwidth of 10c/s to 300kc/s. The time-base is continuously variable from 25c/s to 15kc/s. The overall dimensions are $4\frac{1}{4}$in × $4\frac{3}{4}$in × $3\frac{1}{2}$in.

UNIVERSAL OSCILLOGRAPH

(Illustrated above right)

Southern Instruments Ltd, Frimley Road, Camberley, Surrey

THIS new type of cathode-ray oscillograph has been developed to provide a multi-channel recording equipment of great flexibility. The 3in diameter recording tubes are separate from the main



equipment, housed in a cabinet to which the camera and optical system are fitted. Power supplies and driving voltages are supplied to a master distribution panel in this unit by means of flexible cables from an amplifier console.

Within the recording unit the tubes are mounted in adjustable clamps on a short steel column. This is fixed to a base sliding on rails and enables the tubes to be moved towards or away from the camera lens in a precise manner. Thus, the user can alter the conditions with ease and photograph as many of the tube faces as he needs. It is obviously possible to install a simple equipment in the first place and subsequently to add tubes and amplifiers to provide further channels with a minimum of expense and inconvenience.

Two models are available, providing up to six and twelve channels respectively with time marking.

TACHOMETER

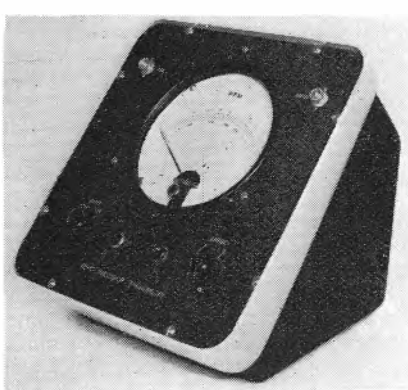
(Illustrated below)

Revo Electric Co. Ltd, Tipton, Staffordshire

THIS is a general purpose instrument for the measurement of speed and frequency. It may be used to measure almost any rotational speed by means of a photo-electric transducer supplied with the instrument. The tachometer is useful for measuring speeds of small mechanisms as no load is imposed on the object whose speed is being measured.

The light source in the transducer is provided from batteries so that no trouble is experienced from light ripple. The transistor amplifier is also contained in the transducer and operated from batteries so that the transducer is completely self contained and may be used with other instruments capable of accepting a 2V peak waveform. Individual transducers can be supplied for specific applications.

The minimum frequency of the instrument is 1c/s but by employing more than one pulse per revolution, lower shaft speeds can be measured. The range covered being 60 to 10 000rev/min with an accuracy of 2 per cent.



PRESS-NUTS

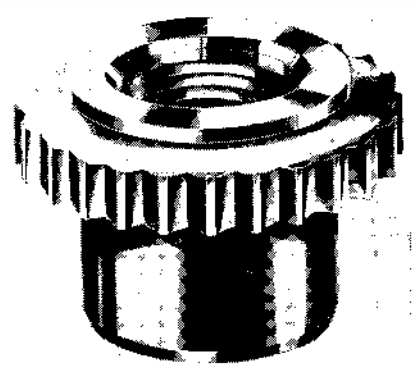
(Illustrated below)

Instrument Screw Co. Ltd, Northolt Road, South Harrow, Middlesex

THE Rosan patent press-nut provides a new and faster way to obtain deep tapped holes in thin sheet metal. This nut is lighter and smaller than most of the existing devices, needs only one hole, is shake-proof and, requires access from one side of the sheet only for fixing.

The Rosan press-nut is manufactured in this country under licence from its American designers by the Instrument Screw Co. Ltd.

The nut is being manufactured in all threads up to ¾in and can be supplied with any thread to special requirement.



Fixing the nut is simple. First, drill or punch a hole in the sheet to the correct size. Insert the smaller flange into the hole. Place a hollow punch over the nut and strike sharply once or twice. The impact forces the sheet metal to flow into the space between the flanges, locking the press-nut securely in place. The teeth of the serrated flange broach themselves into the parent metal preventing rotation.

L.F. PHASE METER

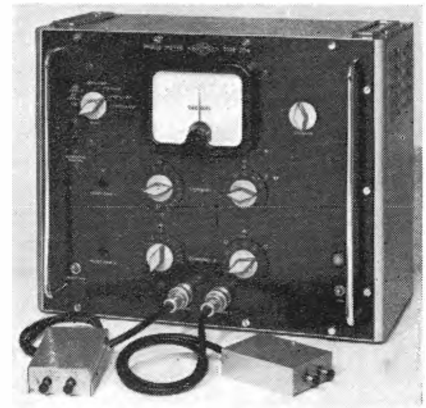
(Illustrated above right)

Airmec Ltd, High Wycombe, Buckinghamshire

THE phase meter type 206 has been designed to enable measurements of gain, attenuation and phase shift to be made on any four-terminal network operating in the frequency range 20c/s to 100kc/s.

Phase is indicated directly on a 6in meter having four scales, and gain or loss values are indicated by the difference between two attenuator settings.

Input signals are connected to the instrument via two cathode follower probes having input impedances of 12MΩ. Measurements are made by adjusting the attenuators in each channel to give standard meter deflexion, phase displacement then being indicated directly on the meter when the 'select' switch is turned to 'phase.' The range of input level on either channel is 1mV to 1V and insertion losses and gains of up to 60dB can therefore be measured.



The circuit consists of two identical amplifiers incorporating cathode follower inputs and calibrated attenuators. Each amplifier is followed by an identical trigger circuit which generates a pulse at the point where the signal crosses the axis. These pulses are then passed to the metering circuit where they determine the relative duty cycles of two valves forming a flip-flop circuit. The differential of these duty cycles is indicated on the meter as phase measurement. Since the instrument indicates the difference in phase where the two signals cross their axes the instrument is relatively insensitive to waveform distortion.

NON-DESTRUCTIVE FLASH TESTER

(Illustrated below)

Labgear Ltd, Willow Place, Cambridge

THIS instrument is designed for the non destructive batch testing of components or other breakdown applications at voltages of 250, 500V, 1kV, 1·5kV and 2kV a.c. These voltages are selected by a high grade ceramic switch controlled from the front panel. Owing to the negligible amount of current flowing in the circuit under test, this instrument is completely non-lethal even at maximum voltage. Indication of breakdown is observed on a moving-coil meter, the pointer of which is adjusted to a red mark on the dial before the circuit under test is connected. Indication of breakdown is

shown when the meter pointer falls below the red mark. Two heavily insulated terminals on the front panel are provided for the attachment of a test jig for the components under test. The main feature of this instrument is that a component or circuit may be tested to breakdown point, and beyond, without damage. The whole instrument is housed in a strong steel case. Other test voltages up to 10kV can be supplied to special order.

WIDE BAND ATTENUATOR

(Illustrated below)

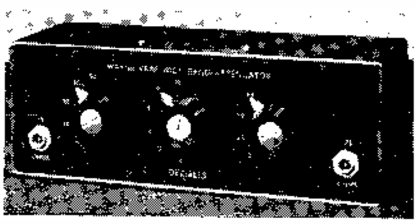
Wayne Kerr Laboratories Ltd, Roebuck Road, Chessington, Surrey

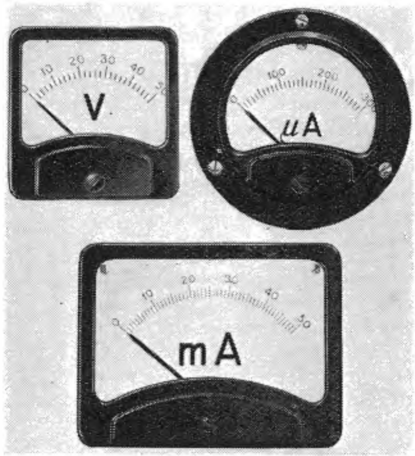
THE type Q.251 is a light, compact and portable 75Ω attenuator of high accuracy and stability for use in the frequency range 0 to 60Mc/s.

The instrument is symmetrical, and is constructed in three well screened sections giving an attenuation range of 0 to 61·5dB in 10dB, 2dB and 0·5dB steps. The accuracy up to 20Mc/s is 0·1dB and 0·3dB at 60Mc/s.

The attenuator is of robust construction and is housed in a light alloy case. If required, the instrument can be supplied without the outer case for incorporation in existing equipment.

The overall size of the attenuator is 8½in × 3¼in × 4¼in weight 4½lb.





MINIATURE PANEL INSTRUMENTS

(Illustrated above right)

The English Electric Co. Ltd, Stafford

A COMPREHENSIVE range of 'miniature' a.c. and d.c. panel instruments is the latest addition to the products of the Instrument Division of The English Electric Company. These are d.c. permanent magnet moving-coil instruments with rectifiers for operation on a.c., frequency range 20c/s to 10kc/s, and with thermocouples for high-frequency applications up to 1Mc/s. All types are supplied in black Bakelite phenolic cases and have nominal scale lengths of 2in, 2½in and 3½in in square pattern cases, 2½in and 3½in round pattern cases, and 3in and 4½in in 3in by 4in by 5in by 6in rectangular cases.

The design incorporates a sintered composite anisotropic magnet with mild-steel pole pieces. Bearing centres are jig bored in relation to the centre pole of the magnet system, the jewelled bearings being located in the die-cast movement holder. This construction gives good mechanical stability and high flux density with low leakage flux, ensuring high sensitivity, low power consumption and a high torque/weight ratio. Sword shaped pointers are fitted as standard, but knife edge (with anti-parallax mirror if required) and spade pointers are also available.

Accuracy of all instruments conforms to B.S.89-1954, typical performance figures for 5mA sensitivity at full scale being:

Torque/weight ratio	87 × 10−4 Newton m/kg (87 Dyne cm/g)
Flux density	0·3090 Webers/m2 (3090 gauss)
Power consumption	47·μW

Laboratory tests show that the instruments can withstand shocks of up to 200*g* in any plane and that the movements are not affected by repeated oscillation and severe vibration conditions.

VIBRATION INDICATOR

(Illustrated below)

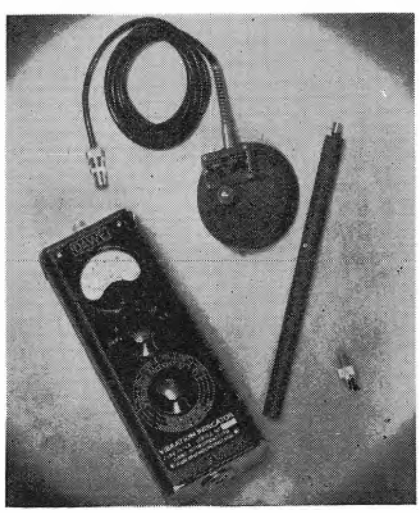
Dawe Instruments Ltd, 99 Uxbridge Road, London, W.5

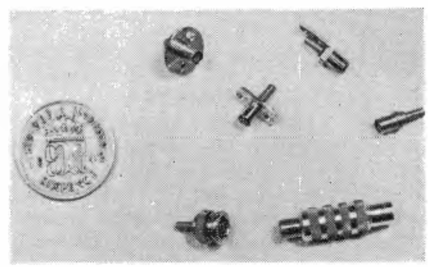
AN inexpensive, simple yet accurate vibration indicator has become available for routine testing and vibration surveys, for which the full scope of elaborate equipment is not required.

Known as the type 1414 vibration indicator, the unit consists of the instrument itself and of a probe, which are connected by means of a cable. The probe, held in contact with the object under investigation, generates an electrical signal which is amplified and indicated on a meter.

This self-contained indicator measures 8¾in × 3⅜in × 2⅞in and weighs only 2½lb, so that it can be held in one hand. It is capable of giving a direct reading of peak acceleration, velocity or displacement of vibration. The required measurement is instantly obtained by turning the control knob to position *A*, *V* or *D*. Two additional positions provide for the periodic checking on the meter of the condition of the built-in h.t. and l.t. batteries. Turning a second control knob cuts successive stages of the resistive attenuator into the amplifier input circuit to provide six ranges in steps of approximately 3 : 1.

The vibration pick-up is of the inertia-operated crystal type delivering a voltage proportional to the acceleration of the applied motion up to a maximum value of some 10*g*. Velocity and displacement readings are obtained by the use of single- and double-integrating networks. The pick-up housing is 3in diameter by 2½in high, the weight with the 8in long extension probe shown in the illustration being 1lb. The extension probe makes it possible to check vibrations in locations which are difficult to reach other wise.





SUB-MINIATURE COAXIAL CONNECTORS

(Illustrated above)

Sales Services, Pathway, Radlett, Hertfordshire

THIS new range of 'micro' electrical connectors has been designed for high grade electronic equipment, the components being particularly serviceable in compact and miniaturized assemblies. With suitable cable, the connectors can be used for linking purposes in connexion with single and multi-channel instruments, and can be mounted close together to form a concentrated test panel: they are ideal for portable instruments and piezo-electric devices.

The plugs and sockets are sufficiently robust to withstand many industrial applications in addition to the more exacting demands of the laboratory; they are therefore of considerable interest to designers engaged in all aspects of electronic engineering.

For chassis mounting, the connectors are designed in different forms for special types of fitting, and can be screwed, bolted, riveted, or connected to the structure. Cable terminations may be screw-clamped or moulded, and 3mm cable, suitable for use with these connectors, is available from the cable manufacturers.

Short News Items

The Radio Industry Council has announced the award of five premiums of 25 guineas each for articles published in the public technical press during 1957. The premiums to be presented at a luncheon of the Public Relations Committee of the Radio Industry Council on Thursday, 10 April, were awarded to the following.

"Heat Control in Electronic Equipment" by E. N. Shaw, (R.R.E. Malvern), *Electronic Engineering*, January, February, March 1957.

"An Alternative Colour TV System" by E. J. Gargini, (E.M.I. Research Department), *Wireless World*, August, 1957.

"Cleaning by Ultrasonics" by L. Atherton, (Mullard Ltd, Equipment Division), *British Communications and Electronics*, March, 1957.

"Speed Control in Industry" by E. J. P. Long, *British Communications and Electronics*, November, December, 1957, January, 1958.

"Components for Printed Circuits" by L. W. D. Sharp, (The Plessey Co, Components Division), *British Communications and Electronics*, April, 1957.

A total of 53 articles published during 1957 were submitted together with five late entries from 1956. Up to six premiums of 25 guineas each have been awarded annually since this scheme was inaugurated in 1952. The object of the scheme is to make more widely known throughout the world British achievements in electronics development.

The 15th Annual Exhibition of The Television Society will be held at The Royal Hotel, Woburn Place, London, W.C.1, from 4-6 March at the following hours: 4 March, 11.30 a.m.-8 p.m.; 5 March, 12 noon-8 p.m.; 6 March, 12 noon-7 p.m. A wide range of equipment and new apparatus will be demonstrated, including colour television receivers. The exhibition covers many facets of the television field. Admission will be by ticket only, available free from the Secretary, Television Society, 166 Shaftesbury Avenue, London, W.C.2, or from any exhibitor.

The British Sound Recording Association will not be holding an exhibition this year, for it is felt that the Audio Fair is successfully catering for the demand for such an exhibition. The Audio Fair Committee have kindly given them a demonstration room at the Waldorf Hotel, Aldwych, London, for the period of the Audio Fair (18-22 April), and it is intended to use this to publicize the Association to a greater number of interested people and to attract new members, as well as to give a service to the existing membership. Further information may be obtained from Mr. S. W. Stevens-Straten, 3 Coombe Gardens, New Malden, Surrey.

The Managing Director of B. & K. Laboratories will not be issuing personal invitations to previous years' visitors to the International Instrument Show. Unfortunately this is no longer practicable owing to the large increase in attendance, but those concerned are invited to write to 57 Union Street or telephone HOP 4567 for tickets.

The Second International Conference for Analog Computations will be held in Strasbourg from 1-7 September. This conference will be devoted to the various systems of analog computations, to their numerous applications, and to the connexions between the analogical methods and the arithmetical methods of computations in electronics. Further information may be obtained from Monsieur F. H. Raymond, Secrétaire Délégué Général de l'Association Internationale pour le Calcul Analogique, c/o S.E.A., 138 Boulevard de Verdun, Courbevoie (Seine), France.

The Thirteenth Annual Electronics Exhibition and Convention organized by the Northern Division of The Institution of Electronics will be held during the periods 10-12 and 14-16 July, at the Manchester College of Science and Technology. The exhibition will consist of a Manufacturers' Section and a Scientific and Industrial Research Section, including exhibits of interest to all branches of science and industry. The convention will include a series of lectures and film shows on electronic topics. Admission will, as formerly, be free of charge. Further particulars may be obtained from the Honorary Exhibition Organizer, Mr. W. Birtwistle, 78 Shaw Road, Rochdale, Lancs.

The University of Liverpool, Department of Extra-Mural Studies, are arranging an Easter vacation course on Basic Principles of Nuclear Engineering from 13-19 April. It is a residential course of sixteen lectures, together with demonstrations, discussions, and a tour of the work at the University Nuclear Physics Laboratory. The inclusive fee for tuition and residence will be £15 15s. Further information and enrolment forms may be obtained on application to the Director of Extra-Mural Studies, 9 Abercromby Square, Liverpool 7.

Evershed and Vignoles Ltd have found it necessary to open a branch office in Belgium. It will be under the management of Mr C. Samyn and will open at the beginning of April. The whole range of electrical instruments, instrumentation and control equipment, will be made available through this office for Belgium, the Belgian Congo and the Grand Duchy of Luxemburg. Mr. Samyn will be responsible to the Company's Overseas Instrument Division and International Instrumentation and Controls Division. The address of this new office is Evershed & Vignoles Ltd, Succursale, 142 Rue Gallait, Brussels.

Multimusic Ltd, a newly-formed wholly-owned subsidiary company of Multicore Solders Ltd, are to manufacture and market specialized magnetic tape recording equipment. The new company have acquired the patents, goodwill and trade marks of Rudman Darlington & Co Ltd, and Rudman Darlington (Electronics) Ltd of Wednesfield, Staffordshire. The manufacture of the professional type "Reflectograph" tape recorder is being transferred to a new factory at Hemel Hempstead adjacent to Multicore Works.

Cawkell Research and Electronics Ltd is a new company which has been formed to extend the activities of the former A. E. Cawkell, Electronic Engineers. The Directors are Mr. A. E. Cawkell (managing) and Mr. J. Rotheroe. The main products will be, as formerly, electronic instruments. By associating with Rotheroe & Mitchell Ltd, precision mechanical engineers, the company may also manufacture equipments to satisfy the increasing trend of combining electronic and mechanical devices in precision control systems. The address for the time being is 6-8 Victory Arcade, Southall, Middlesex, but the company will shortly be moving to new premises.

The Broadcasting Corporation of the Federation of Rhodesia and Nyasaland recently came into being. The new Corporation has its headquarters in Salisbury, and the responsibility for running the Federation's broadcasting services is now in the hands of a Board of Governors, headed by the Chairman, Sir Robert Hudson. The Director-General of the Corporation is Mr. J. McClurg, recently of the South African Broadcasting Corporation, and there are three Controllers in charge of the English and African Services of Administration. The Chief Engineer is Mr. J. M. Reeves who was previously Broadcasting Engineer to the Northern Rhodesia Government.

Letters to the Editor have had to be omitted from this issue on account of the space which has been devoted to ZETA and its instrumentation.

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 26 March. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Electronics in Medicine.
By: R. F. Farr.

South Wales Section

Date: 26 March. Time: 6.30 p.m.
Held at: The Department of Physics, University College of South Wales, Cardiff.
Lecture: Television Lighting Effects.
By: A. E. Robertson.

South Midlands Section

Date: 28 March. Time: 7 p.m.
Held at: The Winter Gardens, Malvern.
Lecture: Transistors in Reception.
By: L. E. Jansson.

West Midlands Section

Date: 12 March. Time: 7.15 p.m.
Held at: Wolverhampton Technical College, Wulfruna Street, Wolverhampton.
Lecture: Analogue Computers.
By: K. C. Garner.

North Eastern Section

Date: 12 March. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Lecture: Industrial Television.
By: S. G. Gobbi.

Merseyside Section

Date: 21 March. Time: 7 p.m.
Held at: Birkenhead Technical College
Lecture: Radio Exploration of the Galaxy.
By: J. E. Baldwin.

Scottish Section

Date: 13 March. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Lecture: Type Reading Equipment.
By: C. E. G. Bailey.

THE INSTITUTE OF PHYSICS

Liverpool and North Wales Branch

Date: 7 March. Time: 7 p.m.
Held at: The Department of Electrical Engineering, University of Liverpool.
Lecture: High Resolution Mass Spectrometry.
By: J. H. Beynon.
Date: 28 March. (Time and place as above).
Lecture: New Methods of Studying the Electronic Structure of Metals.
By: A. B. Pippard.

Midland Branch

Date: 12 March. Time: 7 p.m.
Held at: The Birmingham Exchange and Engineering Centre.
Lecture: Applications of High Speed Photography.
By: E. W. Walker.

South Western Branch

Date: 14 March. Time: 7 p.m.
Held at: The Physics Department, University of Bristol.
Lecture: The Birth of the Nuclear Atom.
By: E. N. da C. Andrade.

Yorkshire Branch

Date: 19 March. Time: 5.15 p.m.
Held at: The Physics Department, University of Leeds.
Lecture: Research by Rockets and Satellites.
By: R. L. F. Boyd.

Electronics Group

Date: 18 March. Time: 5.30 p.m.
Held at: The Institute's House, 47 Belgrave Square, London, S.W.1.
Group Annual General Meeting followed by
Lecture: Microwave Ferrites.
By: L. A. Thomas.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meetings

Date: 20 March.
Lecture: Storage and Manipulation of Information in the Brain.
By: R. L. Beurle.

Supply and Measurement Sections

Date: 12 March.
Lecture: A Method of Analysis of Transformer Impulse Voltage Distribution using a Digital Computer.
By: Beryl M. Dent, E. R. Hartill and J. G. Miles.

Radio and Telecommunication Section

Date: 19 March.
Lectures: Efficiency and Reciprocity in Pulse-Amplitude Modulation. Part 1—Principles.
By: K. W. Cattermole.
Efficiency and Reciprocity in Pulse-Amplitude Modulation. Part 2—Testing and Applications.
By: J. C. Price.
Transistor Pulse Generators for Time-Division Multiplex.
By: K. W. Cattermole.
Date: 31 March.
Informal Evening on Future Radiocommunication Methods for Civil Aircraft.
Talk by: W. E. Brunt.
Date: 27-28 March.
Convention on Radio Aids to Aeronautical and Marine Navigation. (Registration forms available on application).

Supply Section

Date: 26 March.
Lecture: The Objective for Euratom.
By: F. Giordani.

Cambridge Radio and Telecommunication Group

Date: 25 March. Time: 8 p.m.
Held at: University Arms Hotel, Cambridge.
Informal Evening on: Domestic High-Fidelity Reproduction.
By: J. Moir.

East Anglian Sub-Centre

Date: 3 March. Time: 6 p.m.
Held at: The Crown and Anchor Hotel, Ipswich
Lecture: The BBC Sound Broadcasting on Very-High Frequencies.
By: E. W. Hayes and H. Page.
Date: 17 March. Time: 7.30 p.m.
Held at: The Assembly House, Norwich.
Lecture: The Human Operator in Closed-Loop Control Systems.
By: K. A. Hayes.

Mersey and North Wales Centre

Date: 17 March. Time: 6.30 p.m
Held at: The Town Hall, Chester.
Lecture: Recent Uses of Ultrasonics in Investigating the Characteristics of Materials.
By: J. Lamb.
Date: 31 March. Time: 6.30 p.m.
Held at: The Royal Institution, Colquitt Street, Liverpool.
Annual General Meeting and Informal Evening on: Colour Television.
By: L. C. Jesty and E. L. C. White.

North-Eastern Radio and Measurements Group

Date: 3 March. Time: 6.15 p.m.
Held at: King's College, Newcastle upon Tyne.
Lecture: Radio in Air-Sea Rescue.
By: G. W. Hosie, D. Kerr and W. Kiryluk.
Date: 17 March. (Time and place as above).
Lectures: Some Transistor Input Stages for High-Gain D.C. Amplifiers and A Transistor High-Gain Chopper-Type D.C. Amplifier.
By: G. B. B. Chaplin and A. R. Owens.
Date: 31 March. Time: 6.15 p.m.
Held at: The Rutherford College of Technology, Northumberland Road, Newcastle upon Tyne.
Annual General Meeting, followed by a series of demonstrations and displays in the laboratories.

North-Western Radio and Telecommunication Group

Date: 19 March. Time: 6.45 p.m.
Held at: The College of Science and Technology, Manchester.
Informal Lecture: Domestic High-Fidelity Reproduction.
By: J. Moir.

South-East Scotland Sub-Centre

Date: 4 March. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
Lecture: The Digital Computer Applied to the Design of Large Power Transformers.
By: W. A. Sharpley and J. V. Oldfield.
Date: 18 March. (Time and place as above).
Lecture: The BBC Sound Broadcasting Service on Very-High Frequencies.
By: E. W. Hayes and H. Page.

South-West Scotland Sub-Centre

Date: 5 March. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: The Digital Computer Applied to the Design of Large Power Transformers.
By: W. A. Sharpley and J. V. Oldfield.

South Midland Centre

Date: 10 March. Time: 7.30 p.m.
Held at: The Winter Gardens, Malvern.
Lectures: Measurement of Magnetic Fields by Proton Resonance Method.
By: G. C. Lowe.
Tracking the Earth Satellites with the aid of Radio Interferometry.
By: D. B. Yale.

South Midland Radio and Measurements Group

Date: 24 March. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Colour Television
By: D. A. Maurice.

South Midland Supply and Utilization Group

Date: 10 March. Time: 6 p.m.
Held at: The College of Technology, Birmingham.
Informal Lecture: Operational Experience at a Nuclear Power Station.
By: K. L. Stretch.

Southern Centre

Date: 19 March. Time: 6.30 p.m
Held at: The Brighton Technical College.
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.

West Wales (Swansea) Sub-Centre

Date: 13 March. Time: 6 p.m.
Held at: The Conference Rooms, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: The Use of Nuclear Devices.
By: D. Taylor.

Bedford District

Date: 11 March. Time: 7.15 p.m
Held at: The Swan Hotel, Bedford.
Lecture: The Transatlantic Telephone Cable
By: R. J. Halsey.

Hatfield District

Date: 27 March. Time: 7 p.m
Held at: The Hatfield Technical College.
Lecture: Electrical Synthesis of Music.
By: A. Douglas.

Maidstone District

Date: 3 March. Time: 7 p.m.
Held at: The Medway College of Technology, Maidstone Road, Chatham.
Informal Evening on Electronics and Automation.
Talk by: H. A. Thomas on Some Industrial Applications.

Oxford District

Date: 12 March. Time: 7 p.m.
Held at: The Southern Electricity Board, 37 George Street, Oxford.
Lecture: High-Voltage D.C. Transmission, with special reference to Cross Channel Cable.
By: F. J. Lane, O.B.E.

THE TELEVISION SOCIETY

Date: 13 March. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Transistors in Television Receivers.
By: B Overton.

PUBLICATIONS RECEIVED

STEREOPHONIC SOUND by Norman H. Crowhurst is a book which provides information needed to add this dimension to high-fidelity listening. The author explains the theory of stereophonic sound, the difference it can make in listening pleasure and the various systems available or likely to become available. John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $2·25.

ADVANCES IN ELECTRONICS AND ELECTRON PHYSICS VOLUME IX edited by L. Marton of the National Bureau of Standards, Washington, D.C., has recently been published by Academic Press, Inc. Publishers, 111 Fifth Avenue, New York 3. Price $9·00. The main theme is geophysical and originates from the dedication of 1957-58 as the International Geophysical Year.

RADIOISOTOPE LABORATORY TECHNIQUES by R. A. Faires and B. H. Parks is published by George Newnes Ltd, Tower House, Southampton Street, London, W.C.2. Price 25s. The first few chapters are devoted to basic considerations of nuclear physics, isotope production and radiological protection. These are followed by chapters on laboratory design, hazard control and waste disposal. The various methods of detection and measurement are treated from a practical aspect and sections are included on statistics and on the choice of equipment. In the final chapters, a number of radioisotope techniques and applications are reviewed and a chapter is included on the calculation of the feasibility of using isotopes in a particular system.

BAILEY TRANSISTORIZED ELECTRONIC SOLIDS FLOW FAILURE DEVICE is the subject of a recent specification sheet produced by Bailey Meters & Controls Ltd, Purley Way, Croydon, Surrey. This is an instrument which has been developed to give adequate warning on the failure of coal or solids flow in pipes or chutes.