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by the exchange of information in these branches of engineering.”*

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INTERNATIONAL JOURNAL — NATIONAL PROCEEDINGS

THERE has been favourable reaction by members and other readers of the Institution's *Journal* to the announcement† that as from January 1963 the *Journal* will bear the title *The Radio and Electronic Engineer*. Whilst remaining the official *Journal* of the Institution, it is now intended that *The Radio and Electronic Engineer* will only contain papers, notices, and articles of technical interest.

Hitherto the international technical reference value of the *Journal* has been lessened by the need to include “domestic” items which are mainly of interest to Institution members. Conversely the inclusion of such internal information as members might reasonably expect has been kept to a minimum in order to give priority to the publication of technical papers.

Such compromise solutions are seldom satisfactory; the Council has given much time to debating the necessity for increasing the technical content of the *Journal* as well as providing the information and services which will foster the corporate spirit of the entire membership. This latter consideration is not easy of solution because the information needed by members in Great Britain is not wholly suitable for promoting the corporate spirit of members in, for example, India or Canada. Hence the Council wishes to promote the extension of *The Canadian Proceedings* the initial publication of which was so warmly welcomed as indicating the way to provide a *national* service to members of an *international* Institution.

This determination of policy does not abrogate the responsibility of the Institution to every member, wherever he resides, to provide technical information on new developments in his profession—conveyed through *The Radio and Electronic Engineer*—or a forum for debate on matters of education, the application of new radio and electronic development to national requirements, elections to membership of the Institution and (a frequently expressed need) to know more of the Institution's international standards for membership through a discourse on examination papers and the reports of examiners. There is also a national need to have ready reference to advance details of papers to be read before the various Sections, reports of local Section meetings, and among a whole list of other national requirements for the radio and electronic engineer is ready access to book reviews, news of members' appointments, etc.

Expression of these needs is to be made with effect from next month in the publication of a new Institution publication entitled *The Proceedings of the British Institution of Radio Engineers*. Circulation will be confined to members in Great Britain, for it will be essentially a publication to encourage the corporate spirit and interest of members in the United Kingdom. Like the first issue of *The Canadian Proceedings*, however, it will provide what is hoped will be a model for national publications for members to produce in their own countries.

The function of *The Proceedings* in giving advance details of Institution meetings will be important since it will provide this information to a far wider group of members than is at present economically possible through individual mailing. The saving in effort, time and money, both at Institution headquarters and through local sections will be considerable.

This policy of providing an international means of disseminating technical and scientific information through *The Radio and Electronic Engineer* whilst simultaneously catering for the national interest of members is a further milestone in the history of the Brit.I.R.E.

G. D. C.

† *J.Brit.I.R.E.*, 24, No. 4, page 261, October 1962.

INSTITUTION NOTICES

Symposium on Radio and Electronics Research in Great Britain

A Symposium sponsored by the Institution's Research Committee on "Radio and Electronics Research in Great Britain" will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, on Monday, 7th January 1963, starting at 2 p.m. (Note: This Symposium has been postponed from the date announced in the October issue of the *Journal*.)

Opening contributions will be made by Professor D. G. Tucker, D.Sc., Mr. L. H. Bedford, C.B.E., M.A., and Dr. Denis Taylor, Ph.D.

Admission to the Symposium will be by ticket only, obtainable from the Institution.

Joint Brit.I.R.E.-I.E.E. Symposium on Automatic Control

A one day Symposium on "Automatic Control" is being organized jointly by the West Midlands Section of the Brit.I.R.E. and the South Midlands Centre (Electronics Section) of the Institution of Electrical Engineers. The Symposium will be held in the Department of Electrical Engineering, University of Birmingham on Thursday, 28th March 1963.

Six papers will be presented, in which contributors will review modern thought and practice in automatic control and discuss recent developments in industrial applications with particular reference to their relationship to increasing productivity.

Further information may be obtained from the Brit.I.R.E. organizer of the Symposium, Mr. P. Huggins, M.Brit.I.R.E., 25 Tamworth Road, Lichfield, Staffs.

South Western Symposium

As part of the contribution made by professional and technical societies in the South West of England to National Productivity Year, the Institution's South Western Section is sponsoring, in co-operation with the local branches of other bodies, a two-day symposium on "Electronics, Instrumentation and Production". The Symposium will be held at the Bristol College of Advanced Technology, Ashley Down, Bristol, on 12th-13th June 1963. Collaborating bodies include the local branches or sections of the Institution of Electrical Engineers, the British Institute of Management, the Society of Instrument Technology and the Institution of Production Engineers, and the Western Operational Research Society. Members of the Brit.I.R.E. may obtain further details from the Honorary Local Secretary: W. C. Henshaw, M.Sc., 3 Northwick Road, Bristol 7.

Symposium on Masers and Lasers

The Symposium on Masers and Lasers will be held at the London School of Hygiene and Tropical Medicine, on Wednesday, 2nd January 1963, and will consist of two sessions: 10 a.m.-12.30 p.m. and 2 p.m.-5 p.m. In addition to the papers listed in the October *Journal* (page 262), two further papers will be presented, making a total of eight. Details of these additional papers are as follows:

"Single Pulse Optical Maser in Communications and Research"—R. H. Christie (Atomic Weapons Research Establishment).

"High Power Laser Development"—R. A. Fitch (Atomic Weapons Research Establishment).

The registration fees for attendance at the Symposium will be £1 for members and £1 10s. for non-members. These charges cover attendance at the Symposium, the provision of preprints or digests of the papers, and light refreshments during the morning and afternoon sessions.

Registration forms and outline programmes may be obtained from the Institution.

The 1963 Physical Society Exhibition

The Annual Exhibition of the Institute of Physics and the Physical Society will be held in the Halls of the Royal Horticultural Society, Vincent Square, London, S.W.1, from 14th-17th January 1963. It will be open at the following time and admission will be by ticket only:

Monday, 14th January 10 a.m.-7 p.m.

Tuesday, 15th January 10 a.m.-7 p.m.

(10 a.m. to 1 p.m. Members of the Institute and Society, and of sister institutions only.)

Wednesday, 16th January 10 a.m.-7 p.m.

Thursday, 17th January 10 a.m.-7 p.m.

Members may obtain tickets from the Institution and are asked to enclose a stamped addressed envelope. In addition special passes for admission on the "Members morning" (Tuesday, 10-1) have kindly been made available by the Institute and Society and those wishing to attend at this time should indicate accordingly.

Completion of Volume 24

This issue completes Volume 24 of the *Journal* which covers the period July-December 1962. An index will be circulated with the January issue (of *The Radio and Electronic Engineer*).

The Control of Nuclear Power Reactors by means of Magnetic Amplifiers

By

M. J. CRADDOCK †

AND

J. A. PURDIE, B.Sc. ‡

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd-4th April 1962.

Summary: Magnetic amplifiers are suitable for use as amplifying elements in the main temperature control loops of nuclear power reactors. The control problems encountered are discussed and a basic temperature controller and three practical designs based on this controller are described. An indication of the expected reliability of these systems is given, based on experience with similar amplifiers in aircraft applications.

1. General Introduction

In a nuclear reactor, the power produced depends on the rate of fission of fuel. Fast neutrons are produced during fission and these bombard nuclei causing further fission. There are two main types of reactor, "fast" reactors, in which the fast neutrons act directly on the fuel and "slow" or "moderated" reactors in which the fast neutrons are first slowed down by a moderating material and then act on the fuel.¹

Generally, control of the process is established in a fast reactor by inserting additional fuel enclosed in movable rods into the core. This increase of fuel in a given volume increases the bombardment by neutrons which would otherwise escape as leakage, further fissions occur and the core activity increases. Withdrawal of the control rods produces a corresponding decrease in activity.

Up to the present, only slow reactors have been constructed for the commercial production of power. These are usually controlled by the withdrawal of absorber rods constructed with materials such as boron or cadmium. On withdrawal of the rods neutrons which would otherwise be absorbed are free to collide with nuclei in the fuel and the core activity increases.

The core state may be subcritical, critical or supercritical. In the subcritical state neutrons tend to be absorbed or leak away without producing further fissions. When criticality is reached new generations of neutrons are produced by fission at the same rate as old ones are absorbed by the fuel and the chain reaction is just self sustaining. In the supercritical state the chain reaction is growing.

To operate the reactor, safety rods are first withdrawn completely. The control rods are then removed slowly and at some point criticality is reached. Further withdrawal causes the fission rate to increase

and this will continue until the rods are restored to their original position of criticality. This rod position corresponds to a steady condition, the actual power output of the reactor being determined by previous positions occupied by the rods. During operation of the reactor the safety rods are held out and are only operated if the reactor safety circuits dictate that a rapid shut-down is required. A proportion of the rods are manually operated, the remainder being on automatic control.

Figure 1 shows a typical control characteristic for a reactor.²

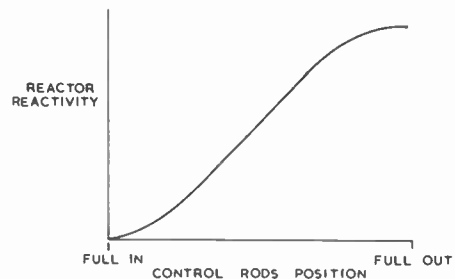


Fig. 1. Typical reactor control rod characteristic.

Since the mechanical life of the fuel elements in the core is adversely affected by temperature cycling, it is highly desirable to operate the core at a constant temperature irrespective of the load the reactor is required to provide. The flow of coolant through the reactor is varied, according to the demand for power, the temperature of the reactor being maintained constant by an automatic closed loop control.

For example, consider a reactor operating at its nominal "datum" temperature and providing a steady power to the coolant system. If a demand for increased power occurs the rate of circulation of coolant is increased to draw more heat per unit time: this provides a greater cooling effect to the reactor and its temperature starts to fall. An error tem-

† Dawe Instruments, Ltd., Western Avenue, London, W.3.

‡ Ultra Electronics Ltd., Western Avenue, London, W.3.

perature occurs (which will be defined as negative). This is detected by thermocouples which initiate the withdrawal of the control rods via amplifiers. As the control rods are removed the reactor goes super-critical and the activity increases, producing power at an increasing rate. The temperature eventually rises above datum and a positive error temperature signal occurs. This automatically returns the control rods to their original position, corresponding to criticality but the rate of fission and therefore the production of power, is now higher and the original datum temperature has been re-established. The system is now balanced and will remain so until a further change in the demanded load occurs. The error temperatures incurred to initiate the control loop are quite small, in the order of 1 deg C and temperature cycling is minimized. A typical datum temperature would be 400°C.

1.1. *Thermocouple Response Lag and Other Delays*

The thermocouples which measure the outlet coolant temperature are normally embedded in grooves at the top of the outlet channels for protection and there is a time lag between a change occurring in the coolant temperature and the resultant change in thermocouple temperature. This lag may be assumed to be an exponential having a time-constant lying between one and fifteen seconds, depending on the design of the reactor.

A further lag exists due to the time taken for the control rod actuator motors to produce a change in rod position, and for the various parts of the reactor fuel and cooling systems to stabilize their temperatures as a result of this. The nature of this delay is complicated but it has been found that the delay can be corrected by a simple anticipation term in the control and if the correct value is chosen a stable system results. The correct value of this term is determined by simulating the reactor on an analogue computer. In the initial stages of the reactor design various assumptions have to be made for this simulation. The

results obtained are only approximate, however; as the design proceeds the assumptions are replaced by more accurate figures so that the best value for the lag can eventually be specified by the reactor engineers. This lag has a typical value in the region of thirty seconds.

It is not always possible to wait until the final value of the lag is known before commencing the design of the electrical control system of the reactor. For this reason some flexibility of design must be retained where practical.

2. **The Basic Temperature Controller**

It is important to minimize the effects of the above lags in order to make the system respond quickly and thus correct errors before they can grow appreciably large. This is achieved by introducing "anticipatory" networks. These have the function of looking at the rate of change of signal into the system and from this, anticipating the final level the signal will reach. (See Appendix 1.)

Figure 2 shows the block diagram of a basic temperature controller suitable for reactor control.

2.1. *Thermocouple Cold Junction Compensation*

The thermocouple measures the difference between the temperatures at its hot and cold junctions. In order to obtain the actual hot junction temperature the effect of variations in cold junction temperature must be compensated. This may be done either by terminating the thermocouple cold junction inside a temperature controlled block of material such as steel or by adding an electrically generated term which is temperature dependent and exactly cancels the effects of temperature changes on the cold junction.

Although the temperature controlled block is often used it has considerable drawbacks. These are:

- (1) Temperature gradients may exist across the block, producing thermocouple errors.

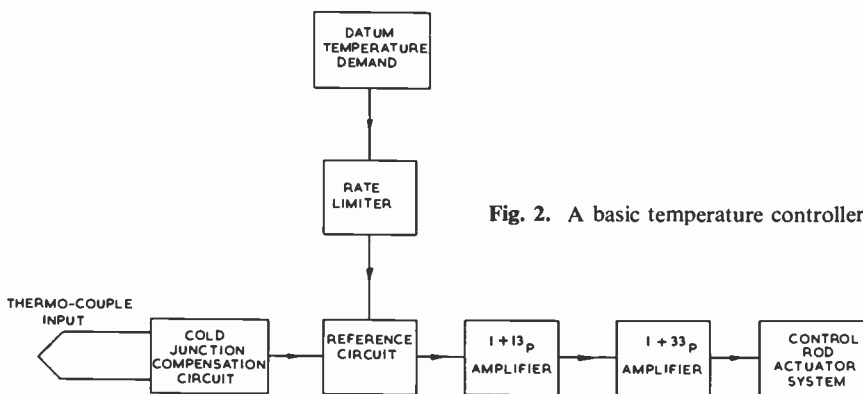


Fig. 2. A basic temperature controller with correction of lags.

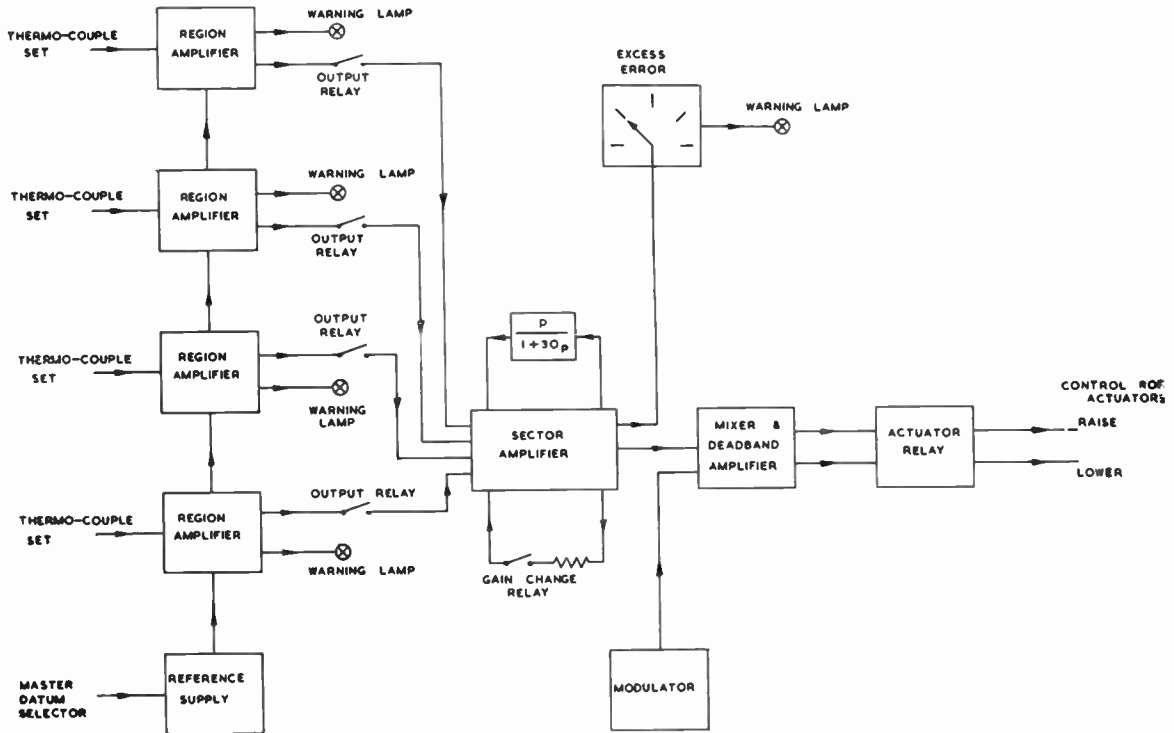


Fig. 3. Sector servo system No. 1.

- (2) A long settling period must be allowed after the thermostatically controlled heater is switched on.
- (3) Correct temperature control of the reactor depends on the thermostat in the block remaining functional and holding its switching temperature accurately.

These disadvantages are avoided if electrical compensation is employed. Appendix 2 describes a practical cold junction compensating bridge which will provide compensation to an accuracy better than 0.25 deg C for a change in cold junction temperature of ± 10 deg C.

2.2. Datum Reference

After cold junction compensation the thermocouple signal is compared with a reference signal, usually a small voltage connected in series and in opposition to the thermocouple voltage. The reference voltage is set to a value electrically equivalent to the temperature at which the reactor is demanded to run.

If the thermocouple voltage is less than the reference a negative voltage occurs. This is amplified and operates on the control rod actuators so as to remove the rods and increase the core temperature. A positive error voltage acts, after amplification, to insert the rods and reduce the temperature.

2.3. Amplifiers

Two amplifiers are used, one of which introduces anticipation, having a time constant of (say) thirteen seconds, to correct for the thermal lag of the thermocouples. The other has anticipation of (say) thirty-three seconds to offset the operating time of the reactor.

Varying the reference voltage provides a convenient way of running the reactor up to its operating temperature and after an operating period, of shutting it down.

2.4. Control of the Reference

For this purpose a manually variable control is made available on the operator's desk. The operator turns this control to the operating temperature he requires. This movement signals a subsidiary servo loop which changes the setting of the potentiometer supplying the reference voltage. The servo loop is arranged to have rate limitation, that is, the rate of change of reference is restricted to one or two degrees Celsius per minute so that the reactor changes its temperature slowly. This is most important as shock temperature changes must be avoided during operation under normal control.

Three recently developed versions of this simple controller will now be described.

3. System No. 1

3.1. General Operation

The reactor is divided into nine sectors, each sector being subdivided into four regions. The temperature of each region is measured by a set of thermocouples. A schematic diagram is shown in Fig. 3.

Magnetic amplifiers are employed as means of amplification. These are chosen because of their high reliability. They are extremely robust and are free from such things as hot filaments or fragile internal structures. They will work over a wide ambient temperature range and, if necessary, under adverse vibration conditions. They provide a very stable means of amplification of low level d.c. signals and have the further advantage that the signal circuit can remain electrically isolated from other circuits.

The magnetic amplifier design and production techniques used in the systems described are based on those used in the temperature and speed control of aircraft engines where the experience of some millions of flying hours has been accumulated.

An outline of the principle of operation of a simple magnetic amplifier is given in Appendix 3. For further details the reader is referred to Bibliography (Section 8).

3.2. Amplifier Layout and Anticipation

The inputs of 4 sets of thermocouples are compared with a set reference datum. The difference signals (errors) are fed into four identical two-stage magnetic amplifiers (regions). These amplifiers each operate a transistor-driven relay, the relays being preset to operate at a given level below the operative datum. As a safety measure a transient negative term is placed over the amplifiers to ensure the relays do not chatter for short term input transients. With the operation of these relays, the amplifiers switch their output into the succeeding amplifier.

This amplifier (sector) is also a two-stage magnetic amplifier and performs four basic functions. The first function is to mix three of the incoming signals (the fourth region being spare).

The second function is to give an excess error indication by meter and warning by lamp. The excess error warning lamp operates at a preset level, the meter continuously monitoring the error.

The third function is to anticipate the rate at which the error changes. The lag in the reactor is approximately 30 seconds at low power and to have a fast acting stable system, a phase advance term of $1 + 30p$ is required. This is obtained by a forward transient network similar to that described in Appendix 1.

The fourth function is to raise the gain of the system in the event of a failure of one operating

region amplifier and the spare. This enables the system to ignore the failure, after having given the appropriate warning.

Following the sector amplifier is the mixer amplifier, again this is a two-stage magnetic amplifier, operating two transistor-driven relays.

The input received from the sector amplifier is mixed with a triangular waveform and operates the output relays according to the polarity of the error. This system provides deadband modulation (see Appendix 4), reducing the error in control which would otherwise result due to the relay deadband.

In a complete system there are nine such units as described above, each operating the control rod actuators for the movement of rods in a given sector in the reactor.

The overall system reference is controlled from one master datum selector in the main control room. This selector operates uniselectors in each sector control, each position of the uniselectors being equal to 2 deg C. A separate datum trim control for each sector is also incorporated, giving a variation of 8 deg C in 2 deg C steps around the selected datum.

A separate temperature trim for every region is also incorporated in each amplifier, in the form of a preset control which is entirely divorced from the master datum selector system. This versatility enables the reactor to be set to a required overall pattern.

3.3. System Functions

The overall system has a drift of not more than 2 deg C, and gives system warnings of:

- (1) excess error, by meter and warning light,
- (2) thermocouple failure, by warning light,
- (3) failure of a region, by warning light.

In the event of a total sector failure the output of the sector controller is open circuited and rods remain at the position they were in at the time of failure. In this case, of course, all warnings are operated.

4. System No. 2

Figure 4 shows the temperature control system used on the reactors in Hinkley Point Power Station.

The reactor is divided into twelve sectors, each sector being separately temperature controlled. The control for one sector is shown. The sector temperature is measured by sixteen thermocouples each of which has separate cold-junction compensation and the current is then fed to a control winding on the first stage of a magnetic amplifier. Thus each thermocouple is electrically isolated, the signals being summed magnetically.

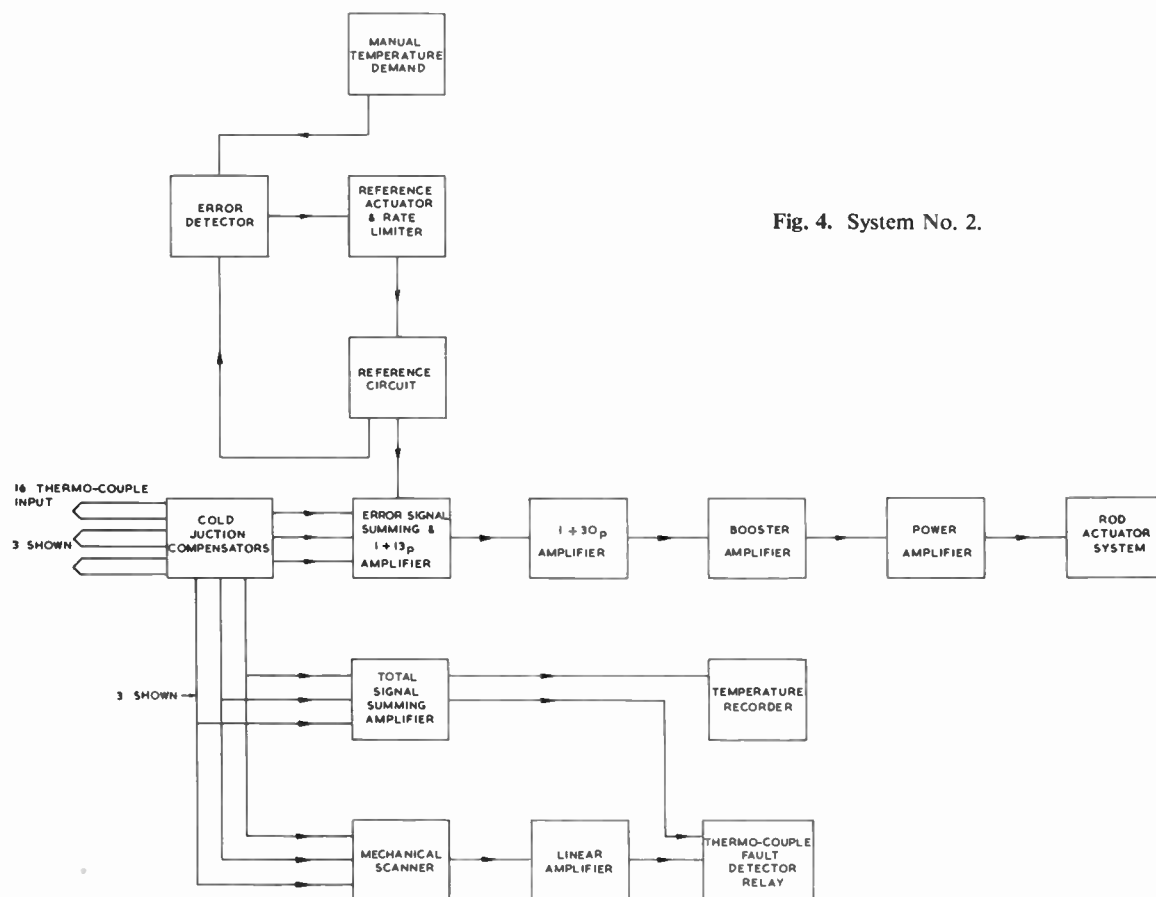


Fig. 4. System No. 2.

4.1. Reference

A reference signal is fed into a further control winding and this subtracts magnetically, from the measured temperature signals; the resultant signal is positive for temperatures in excess of the reference and negative for those lower than the reference. The summing amplifier is a three-stage circuit. Each stage has negative feedback applied, together with negative feedback applied over all three stages. The output is therefore extremely linear and stable.

4.2. Anticipation

A second feature is built into the amplifier in the form of delayed negative feedback. This provides anticipation, having thirteen seconds time constant, which corrects for the thermocouple lag.

Output from the summing amplifier is fed into a transfer function amplifier which produces anticipation, of thirty seconds time constant.

4.3. Other Amplifiers

The amplified signal is then fed through a three-stage booster magnetic amplifier and then a power magnetic amplifier. The output of the power magnetic

amplifier is applied to the field winding of a low frequency generating machine.

4.4. Control Rod Actuator

Three-phase output from the machine is connected to the stator windings of permanent magnet motors which raise and lower the control rods. By suitably signalling the low frequency generator the motors may be driven forward, in reverse, or held stationary electrically.

Thus, if the temperature measured by the thermocouples is higher than the reference, the rods are lowered. If it is lower, the rods are raised. If there is no error the rods remain stationary and the reactor operates at a steady state of criticality.

4.5. Other Functions

Secondary features are supplied by other magnetic amplifiers. One of these is a total-signal summing amplifier which magnetically sums the thermocouple signals and amplifies the sum to a level suitable for operating a pen recorder. This amplifier is a three-stage device with considerable negative feedback to provide good linearity and gain stability.

The possibility of an occasional thermocouple failure cannot be ignored. Spare thermocouples, built into the reactor during its construction, are available as spares. Each thermocouple is scanned every few seconds by a mechanical commutator. The output from the scanner is fed into a magnetic amplifier and amplified to a level suitable for comparison with the output from the total-signal summing amplifier.

If the temperature measured by an individual thermocouple differs by more than 70 deg C from the average of the sixteen, a differential relay operates. This signals a logic circuit which disconnects the faulty thermocouple, registers a failure and changes the gain of the total signal summing amplifier, so that the correct sector temperature is still recorded with fifteen thermocouples.

Provision is made so that the system will continue to control and record the sector temperature correctly if a second thermocouple fails before the first is replaced.

This system will control the temperature at a value dictated by the magnitude of the reference current supplied to the first stage of the error signal summing circuit.

4.6. General Comments on System No. 2

A disadvantage of the magnetic mixing technique employed in the first stages of the summing amplifiers

is the effect of variation in the resistances of the copper windings with varying temperature. This alters the currents in the windings and gives rise to control errors. It is overcome by mounting the summing amplifiers (and also, for convenience, the linear scanning amplifier and booster amplifier), in a thermostatically-controlled rack. The rack temperature is controlled at 52°C which is above the extreme maximum ambient temperature that can occur.

Failure of the thermostatic control is detected by upper and lower warning thermostats which operate a few degrees away from the control temperature.

As part of the production test specification every amplifier is tested at various temperatures to ensure that even with a failure in the thermostatic control, the error introduced into the sector temperature is small.

Thermostatic control eliminates the day and night cycling of 1 deg C in the sector, which would otherwise occur and thereby contributes to extending the life of the reactor fuel elements.

Most of the amplifiers in this system operate from a 400 c/s supply.

5. System No. 3

The control system shown in Fig. 5 is in the design stage at present and will be used in a power station to be built shortly. It has been evolved from ex-

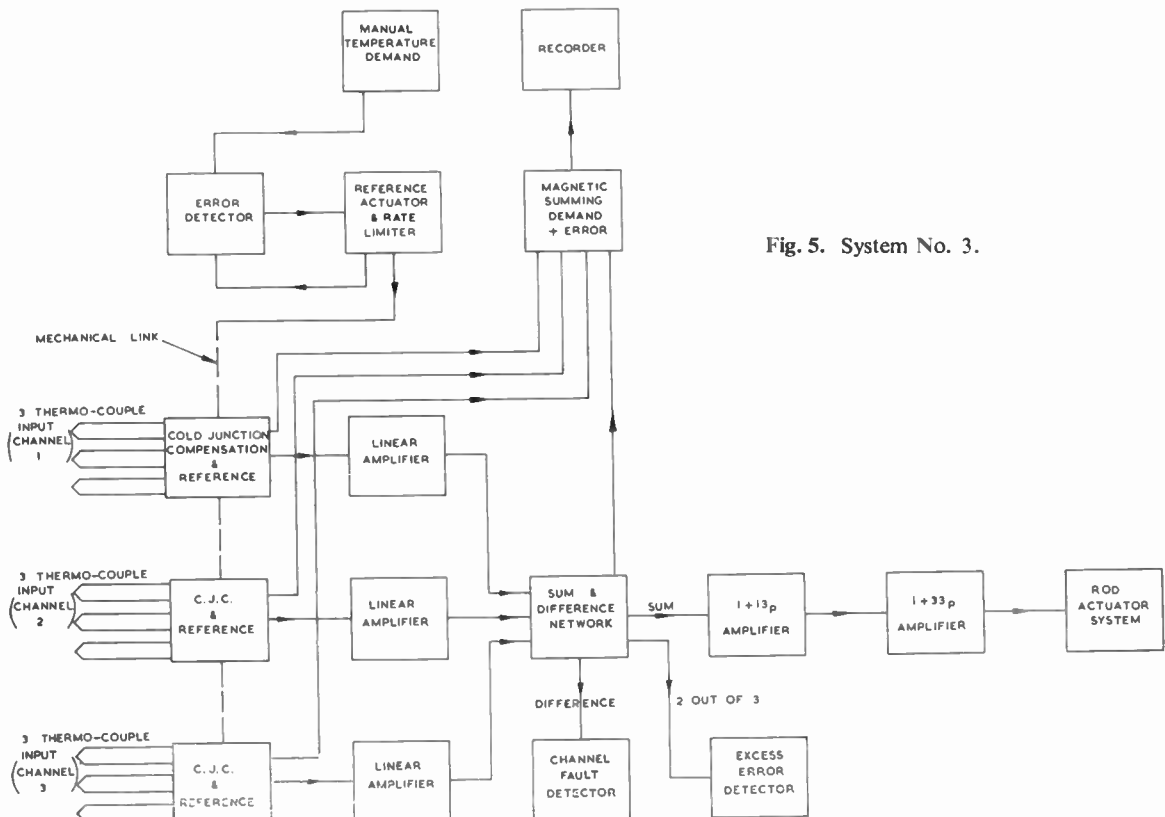


Fig. 5. System No. 3.

perience with System No. 2 and offers more features with greater simplicity. It is also expected to require less routine maintenance, since the mechanical scanner and the temperature-controlled rack of the previous system have been eliminated.

In this new scheme, nine thermocouples are used to detect the sector temperature. These are connected in series in sets of three, so that the cold junction compensation circuitry is reduced. The reference is applied in the form of a voltage connected in series with the thermocouple so that an error signal is produced, before arriving at the input of the magnetic amplifiers. This method eliminates system drift due to changes in the ambient temperature of the amplifiers so that thermostatic control and its auxiliary warning circuits are not required.

5.1. Amplifier Layout

The three signals are fed into three linear magnetic amplifiers which give outputs proportional to the magnitude of the error signal inputs. Under normal conditions these three outputs are virtually identical but if a thermocouple fails, the output of one channel will differ appreciably from the others and by comparing the outputs, the faulty channel can be detected. Thus the need for a thermocouple scanning system is eliminated. It can be seen that any kind of fault whether in the thermocouples, cold junction compensation or linear amplifiers will also be detected.

The outputs of the three amplifiers are fed into a sum and difference network. The sum of the three outputs is fed forward into an anticipation magnetic amplifier which incorporates a time-constant of 13 seconds. The output from this then passes through a second anticipation amplifier which incorporates a thirty-three second time-constant. The signal is now suitable for operating the control rod actuator system.

5.2. Other Functions

The sum of the three linear amplifier outputs is also fed to a further summing amplifier where it is added to the original reference or "demanded" signal. The sum of the demanded temperature plus the error temperature equals the actual temperature and this is fed to the temperature recorder.

Any appreciable difference between one linear amplifier output and the average of the other two is detected as a fault and the faulty channel is switched off; at the same time this loss of one-third of the signals is offset by switching amplifier gain. The detected fault is indicated by a lamp on the front of the control cubicle, as well as by a suitable alarm. The system continues to control the region temperature on a "two out of three" basis, but any further fault results in the control being switched off with

the sector "frozen" at the last control position that applied before the fault.

Finally, the sum of the three outputs is inspected by an "excess error detector" unit. If, for any reason, the actual sector temperature differs by more than a predetermined amount of that demanded, the control system is failing to achieve its function and an excessive output occurs from the three linear amplifiers. This failure is detected by the excess error detector which operates to freeze the sector and indicate a fault condition, provides an alarm and lights an indicator lamp.

5.3. Reference

The reference signal is controlled from the operator's desk in a similar way to that of the previous system and embodies rate limitation of 2 deg C per minute.

5.4. Amplifier Specifications

All magnetic amplifiers are designed to operate from a 50 c/s supply and each amplifier contains a basic standardized three-stage push-pull design which is satisfactory for operation over the ambient temperature range 0°C to +60°C with variations in supply voltage of +8%, -20% and in supply frequency of $\pm 10\%$.

The overall accuracy of control of sector temperature is expected to be better than 2 deg C of error at an operating temperature of 400°C over a period of one year.

6. Reliability

The reliability of the systems can only be compared with that of similar devices which have already had long service. Failures due to actual transducer breakdown are exceedingly rare and no figures can be quoted.

Failures due to associated components like rectifiers, etc., can be quoted, but these again are for conditions far in excess of those ever likely to be met in a power station. There are about 50 rectifiers in a sector control, and these sectors are intended to have a life of twenty years (180 000 hours).

Failure records show an expected rectifier life in excess of 5 million hours, this being computed from airborne equipment working in varying ambient temperatures from -20°C to +50°C at the same time under certain vibration stress. It is reasonable to assume that under the conditions met in power stations that a life in excess of 7.5 million hours could be expected.

Other components such as relays and uniselectors are difficult to assess, the number of operations being uncertain. The life of a standard Post Office relay is in the order of 10 million operations, and that of a uniselector is in excess of 100 million steps.

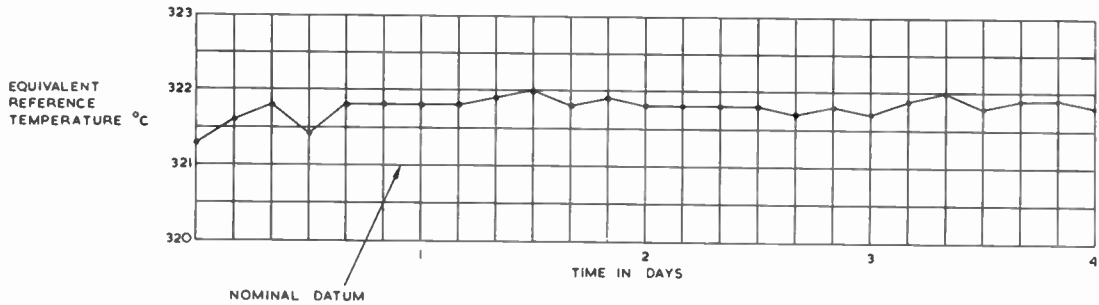


Fig. 6. Reference drift.

A further feature of reliability can be taken as overall drift over a period of time. In the systems described, this must be taken in two forms. The systems are generally zero seeking, and the output for no input ideally should remain at zero. The zero drift is taken as the amount of input required to correct this drift. Zero drift over periods of six months or so has been found to be in the order of ± 1 deg C. A typical drift curve is shown in Fig. 6. The other drift is due to the reference circuit. This is in the order of ± 1 deg C.

7. Acknowledgments

The directors of Ultra Electronics Limited are thanked for their permission to publish this paper.

System No. 1 was developed to the specification of Atomic Power Constructions Limited and due acknowledgment is given to that Company.

The authors wish to thank Mr. S. Tweedy of Ultra Electronics Limited for help in preparing this paper.

8. References and Bibliography

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9. Appendix 1

Analysis of an Anticipation Network

For the purposes of a simplified explanation, it is assumed that the thermocouple is characterized by a single time-constant and is subjected to a sudden step change in the temperature of its surroundings. This of course is a state of affairs which cannot occur in

practice, but nevertheless illustrates the mechanism of the action of the anticipation network.

The voltage of the thermocouple will then rise in an exponential manner from a value V_1 , say, to a value V_2 , with time-constant T , as shown in Fig. 7. Suppose that this voltage is applied to the circuit of Fig. 8, and suppose further that the resistance r is small enough to be neglected. The current i_1 , flowing through R will then be of the same shape as the applied thermocouple voltage and is drawn in Fig. 9. The current i_2 flowing in the capacitor C is proportional to the rate of change of voltage and will therefore jump to a certain value depending on the value of C , then falling to zero exponentially with a time-constant that of the thermocouple, i.e. T . This current is drawn in the same figure.

The current in r is the sum of these currents, and it is seen from the figure that if C is chosen such that the initial jump in current i_2 is equal to the overall change in current i_1 , then the total current $i_1 + i_2$ jumps immediately to its final value and remains

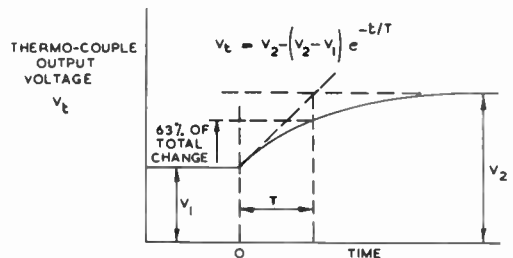


Fig. 7. Response to a step change in temperature from a thermocouple having a thermal lag.

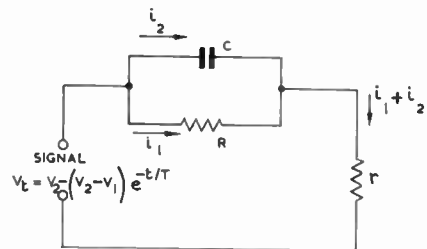


Fig. 8. Suitable anticipation circuit.

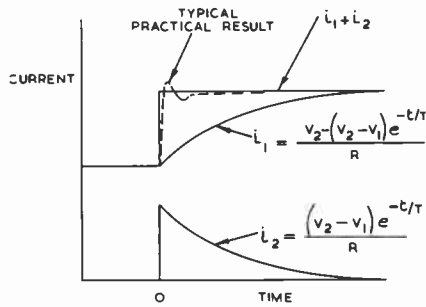


Fig. 9. Currents in an anticipation network.

there, i.e. it exactly follows the applied temperature. Deeper analysis shows that the correct value of C is such as to make the time-constant CR equal to the thermocouple time-constant T .

The output current $i_1 + i_2$ has exactly followed the temperature in the case of a step change. It will also follow the temperature for other types of change. Thus in this simplified case perfect compensation for the effect of the thermocouple time-constant has been achieved.

In actual fact, of course, perfect compensation cannot be achieved; many factors tend to prevent it. Firstly, the resistance r comprising the resistance of the amplifier control winding, thermocouple and leads cannot be entirely ignored, also the capacitor is not perfect, and the assumption that the thermocouple can be characterized by a single time-constant is not absolutely true.

However, by suitable choice of C and R a response is obtainable which adequately compensates for the thermocouple delay, and a typical response is indicated in Fig. 9.

10. Appendix 2

Cold Junction Compensation and Reference Circuits

In all systems operating from thermocouples one of the major problems is to be able to compensate for cold junction temperatures and to obtain a reference for comparison.

A device of this kind must be extremely stable under adverse conditions, i.e. ambient temperature variations and power supply changes. If it does not comply with these required conditions, then the whole system cannot maintain its required accuracy. Several methods are employed in an attempt to overcome these problems.

One of the most successful methods is to use a simple Wheatstone bridge supplied from a constant current source. In the two opposing arms of the bridge are two relatively high resistances and two low resistances, one of the low resistances being wound in

copper. This copper resistance has a temperature coefficient and is chosen so that for a given change of ambient temperature, the voltage seen across the resistor changes by an amount equal to the change in output of a thermocouple when its cold junction undergoes the same ambient temperature change. By connecting the bridge in series with the thermocouple its cold junction becomes compensated.

Example. Suppose it is required to compensate a Chromel/Alumel thermocouple, having a sensitivity of $40 \mu\text{V}$ per deg C and that a 2 mA stabilized supply is available. The circuit shown in Fig. 10 would compensate for ambient temperature and relate the thermocouple to 0°C .

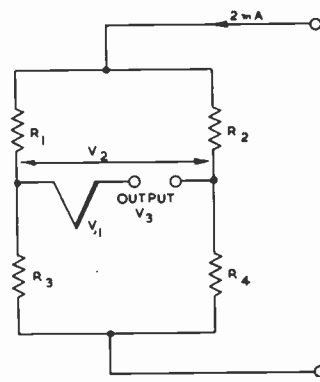


Fig. 10. Simple cold junction compensation circuit.

R_1, R_2 are relatively high resistances, say about 1.5 kilohms, and matched to 0.5%. R_3 is a copper resistance calculated as follows: the coefficient of copper is 0.00393, the potential required across R_3 for a change of 1 deg C ambient is 0.040 mV. This means a resistance change of 0.04 ohms (1 mA in the arm), so that this resistance will be $0.04/0.00393 = 10.17$ ohms at 0°C .

R_4 is made equal to this resistance but is wound in manganin or another material having negligible temperature coefficient (coefficient of manganin is 0.00001).

The effect on the bridge is such that at an ambient temperature of 0°C , V_2 is zero, V_3 is equal to the temperature input V_1 . At an ambient temperature of 20°C , V_2 is equivalent to 20 deg C electrically, V_1 is equal to the input temperature minus the cold junction temperature (ambient 20°C), V_2 therefore remains constant and the cold junction is compensated.

It is possible to incorporate a reference as well by only a slight alteration to the bridge. There are three main ways this can be done, two of them very similar in approach, the third a little more complex.

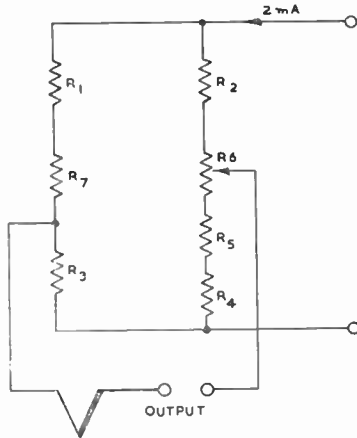


Fig. 11. Cold junction compensation incorporating reference.

- R_1 and R_2 High value matched resistances.
- R_3 Copper resistance.
- R_4 Fixed resistor equal to R_3 at 0°C .
- R_5 Basic reference resistance.
- R_6 Potentiometer or a set series of resistances.
- R_7 Equal to R_5 and R_6 .

The first two ways are by adding further resistance in series with R_4 , either by a potentiometer or by a series of small resistances giving an equal amount of reference (Fig. 11). The potentiometer can be calibrated, and used satisfactorily as a trim or a major reference selector. The series of resistors can be selected in steps by switches or a uniselector, the latter giving a simple system of remote control.

Both the above systems suffer from the disadvantage that they introduce mechanical contacts into the system. This problem only arises, of course, due to the low potentials involved, and in actual use may only make a difference of perhaps 1 deg C, but in a highly accurate system this may be the whole of the allowable tolerance.

To appreciate the third way, the simple operation of the bridge should be explained. The thermocouple and the input winding of the following amplifier are in series across the bridge. The compensation due to the copper bobbin is added to the incoming signal. The reference acts in opposition to the incoming signal and therefore causes the difference signal to be injected into the input of the amplifier. The system will adjust itself so that the error (difference signal) is zero, in other words the incoming signal is equal to the reference.

If a variable potential is introduced into the input line, the same effect of a varying reference will be achieved. A Hall generator is used to achieve this. When a wafer of a semiconductor material has a current passed through its axis, and the wafer is placed at 90 deg to a magnetic field, then a voltage

appears across the opposite axis. This is known as the Hall effect. As the wafer is rotated in the field, the voltage decreases to zero. The pole pieces of the magnet producing the field, can be shaped to make this voltage follow a required curve.

The Hall effect is quoted as:

$$V_H = \frac{R \times H \times I}{t} \times 10^{-8}$$

where V_H = Hall voltage

R = Hall constant

H = field in gauss

I = exciter current

t = thickness of the element in cm.

Thus the Hall effect gives a means of inserting a varying potential into the input line, without using a mechanical wiping contact.

Usually the high reference temperatures are the most critical and the Hall generator can be conveniently connected so as to back off a high, constant reference. This gives improved accuracy at high temperature since the Hall generator contributes an error only due to its change of zero.

A suitable circuit incorporating a Hall generator is shown in Fig. 12.

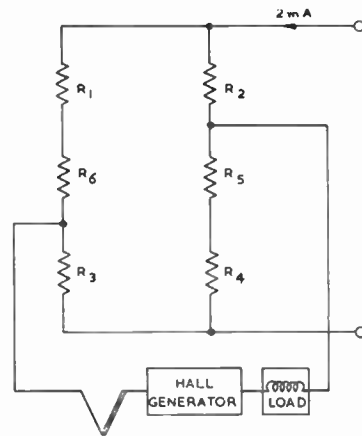


Fig. 12. Cold junction compensation and reference using a Hall generator.

R_6 equals R_5 , R_3 being in this case the high reference level.

V_2 in this case represents the high reference less ambient temperature (voltage across R_3). This requires that to obtain zero error, V_1 must equal V_2 . To maintain a lower reference, the Hall voltage V_3 is increased and added to V_1 so that $V_1 + V_3 = V_2$. The variation then in this case is from the high reference to the maximum backing-off voltage of the Hall generator.

As explained previously the sense can be reversed so that $V_1 - V_3 = V_2$ where V_2 is the low reference.

The use of the Hall effect as a variable reference has several advantages, it is contactless, its resistance in zero field is low, its output can be shaped readily to suit a given curve. It is easily compensated for wide variations of ambient temperature.

A cold junction compensation and reference system such as this is accurate to $\pm 80 \mu\text{V}$ over the ambient range of -40°C to $+100^\circ\text{C}$, so that for this application in power stations, possible ambients of $+10^\circ\text{C}$ to $+40^\circ\text{C}$ are more than adequately compensated.

11. Appendix 3

The Magnetic Amplifier

There are many kinds of magnetic amplifier, each type having a particular principle of operation. However, all types have one feature in common, namely, the degree of saturation of a ferro-magnetic core is varied by a current flowing in a winding on the core (a "signal" or "control" current) and this degree of saturation determines the effective impedance of a second winding, connected in series with a load, across an a.c. supply. It can be arranged so that the

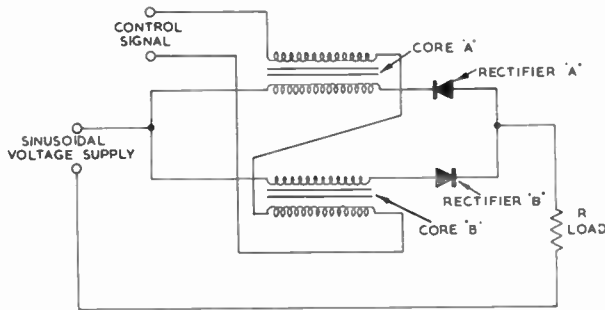


Fig. 13. A simple magnetic amplifier.

power change developed in the load is greater than the change in power dissipated in the control circuit, thereby achieving amplification. The type of magnetic amplifier described below is greatly favoured by designers and is generally known as the "auto-self-excited magnetic amplifier".

11.1. Principle of Operation

The circuit diagram of the simple magnetic amplifier is shown in Fig. 13. The circuit consists of two cores on which are wound a control winding and a main winding.

In designing a magnetic amplifier a ferro-magnetic core material is chosen which has a high remanence, low coercivity and sharp saturation. The main windings are connected in parallel via rectifiers and this network is connected in series with a load across an a.c. supply. The supply voltage is just sufficient to saturate the cores. The control windings are connected in series to a source of control current.

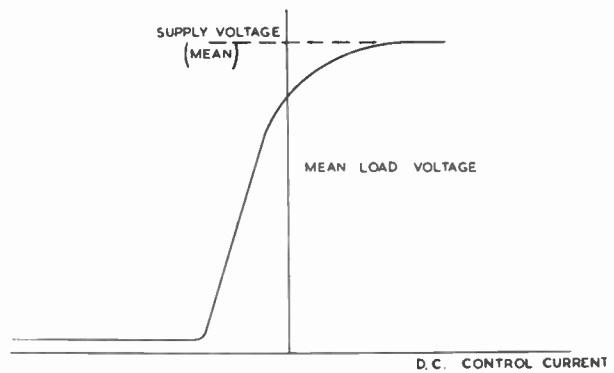


Fig. 14. Load voltage characteristic.

If the control current is varied a load voltage characteristic is obtained which takes the form shown in Fig. 14. Consider the amplifier with zero current flowing in the control winding. Because of the blocking effect of rectifier A, the current through the main winding on core A is unidirectional. Thus after a few cycles of supply the core will be operating on the minor hysteresis loop shown in Fig. 15.

After a conducting half cycle the core is left at a flux density equal to its remanence, i.e. near saturation. When the next conducting half cycle is applied, the core is able to support only a small portion of the voltage before it completely saturates. After saturation, the supply voltage appears in full, across the load (voltage drops due to the d.c. resistance of coil windings have been neglected). On opposite half cycles core B traverses a similar minor hysteresis loop. Thus at zero control signal, the voltage appearing on the load is high and is determined by the remanence of the cores.

It has been seen that at zero control the cores are always operating near saturation. Any additional positive control current can only drive the cores further towards saturation. Increasing positive control therefore increases the load voltage slightly until it

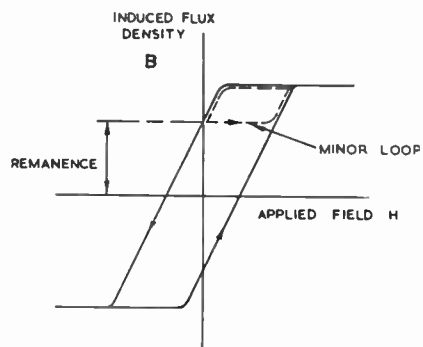


Fig. 15. Hysteresis loop of a magnetic core.

equals the supply voltage (neglecting ohmic voltage drops across the main windings).

When negative control current flows, the cores reset to lower levels of flux, depending on the magnitude of the current. When the next half cycle of supply voltage occurs the core must move from this lower level to saturation and in doing this it supports a greater proportion of the supply half cycle before it saturates. Thus an increasing negative signal produces decreasing load voltage.

In considering the magnetic amplifier so far, the circulation of harmonic voltages in the control circuit, which can occur, has been neglected. These have the effect of somewhat modifying the characteristic.

Milnes⁴ gives the mean ampere-turns gain of this kind of magnetic amplifier as

$$\frac{\delta(I_a N_a)}{\delta(I_c N_c)} = \frac{1}{\pi} \cdot \frac{X}{R_a}$$

where X is the reactance of an a.c. winding.

Figure 16 shows the circuit diagram of a pair of simple transducers connected in push-pull to give a reversible d.c. output. The two halves of the push-pull unit are biased to about halfway up the linear portion of their characteristics by a constant current flowing in a separate control winding.

This type of amplifier exhibits a very high zero stability, that is, the output variation at zero signal due to changes in supply voltage, supply frequency, ambient temperature and time is small.

Typically, the power which must be dissipated in the control circuit to correct for these small drifts lies in the region of 10^{-11} to 10^{-12} watts. Thus an error signal of 1 deg C detected by, say, three thermocouples each with a coefficient of 0.04 mV per deg C connected in series with a control winding, and having a total control circuit resistance of 150 ohms, will dissipate 10^{-10} watts in the control circuit and this will be readily detected as an error signal. The power

gain of a single stage push-pull amplifier operating from a 50-c/s supply is in the region of 1000.

It is usually preferred to operate magnetic amplifiers at supply frequencies in the range 400 to 1600 c/s where good sensitivity is obtained. Designs for operation at a frequency below this band result in an increased size of transducer and some sacrifice in stability, particularly where feedback techniques are used. However there is usually a good argument for operation at 50 c/s since guaranteed 50-c/s supplies are normally available. Operation at a higher frequency necessitates the introduction of a frequency converter and this extra device must itself be reliable so that several converters must be incorporated to protect against loss of the control system due to the failure of a single converter. For this reason the trend is to operate magnetic amplifiers from 50-c/s supplies.

12. Appendix 4 Deadband Control

In a deadband control system, corrective action is initiated, for example by operating a relay, when the system error exceeds a predetermined level. Two relays are used, one operating for positive errors, the other for negative errors.

At low levels of error no correction takes place. The system error can then be at least the magnitude of the deadband setting. The deadband error can be removed by mixing the error signal from the amplifier with a "dither" signal, the resultant being taken to a deadband actuator amplifier. The dither signal is a triangular waveform of magnitude just less than that required to overcome the deadband. At zero error signal the triangular modulation has no effect (Fig. 17). When an error occurs the actuator is pulsed at the modulation frequency with a pulse proportional to the magnitude of the error (Fig. 18). The system becomes quasi-proportional and the rate of correction is proportional to the error. For stability the modulation frequency is chosen to be high compared with the natural frequency of the system.

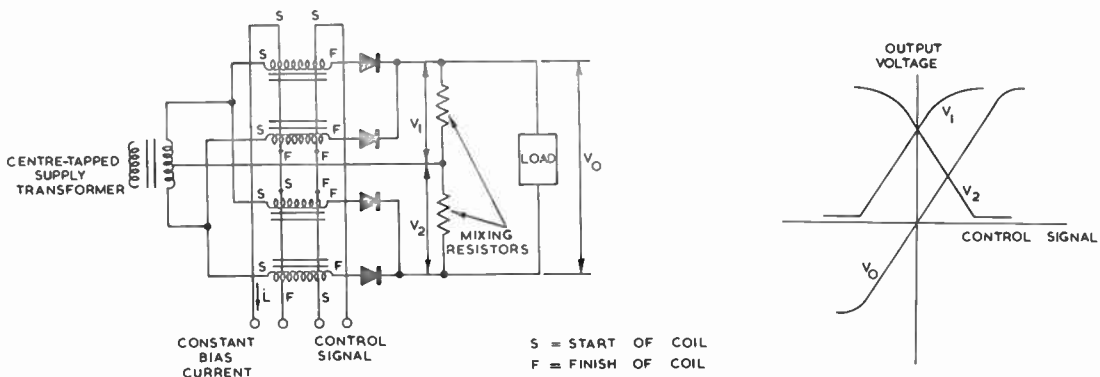


Fig. 16. Push-pull magnetic amplifier with d.c. output.

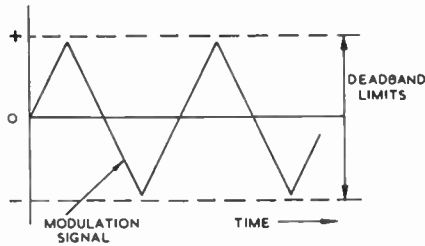


Fig. 17. Zero error.

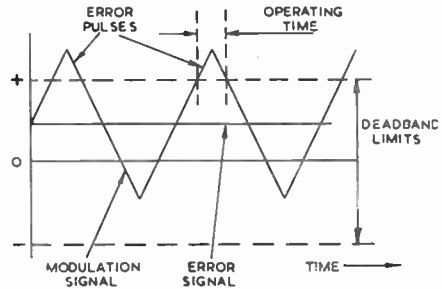


Fig. 18. System in error.

The following is a practical application of the system. The actuator is a magnetic amplifier the output of which operates two transistor-driven relays. Diodes are inserted so that the relays operate on signals of opposite polarity. The bias to the transistors is variable so that their operation can be controlled, these controls setting the deadband of the system.

The inputs of the magnetic amplifier are (1) the error signal, and (2) the modulator waveform, the latter having a controlled amplitude.

Thus it is possible to set the deadband (at zero error) over a wide range, and maintain the modulation within that deadband. In fact it can be arranged that both these controls are ganged and act as the system gain.

Manuscript received by the Institution on 8th February 1962 (Paper No. 772).

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POINTS FROM THE DISCUSSION

Mr. J. O. Green (Graduate): Mention was made of the use of an anticipatory circuit to allow for the inherent lag of the temperature measuring devices as well as the system as a whole. This assumed that the total lag is constant over fairly long periods; however, at least one other system employed on a power reactor does not recognise this as being so. In this system terms are continuously fed to the controller to modify the amount of phase advance and overall gain in sympathy with coolant gas flow and rod effectiveness, the latter being dependant upon the position in the reactor core. Would the authors care to comment on whether the efficiency gained from running closer to the demanded temperature for the longest possible time has been sacrificed for equipment simplification.

Mr. Purdie (in reply): One of the most important factors in the operation of a nuclear power reactor is to ensure that the maximum life is obtained from the fuel elements. This life is likely to be reduced if the elements are allowed to undergo frequent changes in temperature. The object of the control systems described is to maintain a temperature constant to within one or two degrees Celsius irrespective of variations in coolant gas flow through the reactor. This accuracy is maintained for both continuous conditions and during transient changes, by the correct choice of anticipatory circuits.

The anticipation requirements are determined during the design stage, by simulating the reactor working at

various conditions, on an analogue computer. The anticipation terms required to stabilize the reactor are obtained and then simplified as far as possible consistent with maintaining a very stable temperature, at the high power levels (i.e. the normal condition of the reactor, acting as a base load) at the same time, giving satisfactory stability at low power levels (i.e. during run-up).

The object is to take advantage of equipment simplification without sacrificing the accuracy of temperature control under normal conditions and tolerating only a minimal error for other conditions which occur infrequently. This simplicity gives better reliability, ease of servicing and economy.

Mr. R. Scott: Will the authors please comment on the overall deadband of the controller described. Is the deadband variable? If so what is the minimum value in degrees Celsius?

Mr. Craddock (in reply): The deadband is variable from 2 deg C to 12 deg C in 1 deg C steps. The deadband modulation expressed as a percentage of the deadband, is variable from 50% to 100% in 10% steps. The modulation frequency may be selected either as 0.2 c/s or 0.02 c/s.

These latitudes have been built into the design by simple circuitry to enable the best operating conditions to be obtained during commissioning, consistent with system noise and minimum rod movement.

OBITUARY

The Council has learned with regret of the deaths of the following members:

Basil Montefiore Joseph Davis (Member) was born in London in 1892. After service in the Royal Naval Air Service during the first World War he returned to join the Gaumont British Picture Corporation and was concerned with installation of sound equipment in cinemas. He became chief engineer of G.B. Picture Corporation in 1927, and subsequently held other similar appointments in cinematograph engineering. During the Second World War he served in the Royal Air Force as a technical signals officer both in this country at the Air Ministry and in the Ministry of Aircraft Production, and in Canada.

Mr. Davis was elected to Full Membership of the Institution in 1944. He served on the Finance Committee from 1945 to 1950 and on the Membership Committee from 1950 to 1953.

Ill-health in the post-war years led to his retirement from active professional work. He died during August whilst on holiday in Torquay, as a result of cerebral haemorrhage.

Thomas Silas Woodget (Member) was born in Streatham, London, in 1913 and received his technical education in General and Electrical Engineering at the Regent Street Polytechnic. He served his apprenticeship in the Research Laboratories of the General Electric Company at Wembley and joined the M-O Valve Company in 1934, advancing through various technical grades to technical superintendent by May 1941. From 1945 onwards he further progressed through various technical and production appointments, being made works manager in April 1959 and production director in April 1962. Mr. Woodget was elected to Full Membership of the Institution in June 1961; he was also a Member of the British Institute of Management.

He died suddenly and unexpectedly on 25th May last while playing golf. He leaves a widow and two daughters.
W.G.N.

Gordon Wills (Associate) joined the East African Posts & Telecommunications Administration in 1949 from Rediffusion Ltd., following service during the war with the R.A.F.

In his early days in East Africa he was engaged on the installation and maintenance of remote police radio stations in the Northern Frontier District of Kenya and then on maintenance of the Administration's v.h.f. multi-channel routes. In recent years he was employed on planning of trunk telecommunications in the office of the Engineer-in-Chief.

He served in Nairobi and Mombasa and his quiet pleasant personality made many friends. He is well

remembered in Mombasa for the concerts of classical music which he organized in the Cathedral.

He had the inquiring mind of the true engineer and was constantly improving his knowledge of the telecommunications field. His death on 22nd July 1962 at the age of 40 years came as a great blow to all who knew him. He leaves a widow and six children.

Mr. Wills was elected to Associateship of the Institution in 1959; his exemption from the Graduate-ship examination was unfortunately completed too late to enable him to seek transfer to Corporate Membership.
D.H.A.S.

Frederick Edwin Aplin (Associate Member) who died on 11th March 1962 aged 43 years, was a technical administrator with Amalgamated Electric Services Ltd. He served with the Royal Signals as Foreman of Signals during the War and subsequently held a number of posts in the receiver manufacturing and servicing branches of the radio industry. Mr. Aplin was elected an Associate of the Institution in 1946 and was transferred to Associate Membership in 1951.

Anthony Cornwallis Petherick (Associate) who died on 13th April 1962 after a long illness, served in the Royal Navy during the war, initially as ship's radar officer and, for eighteen months prior to demobilization, as an instructor. He then joined Marconi's Wireless Telegraph Company as a technical representative overseas, working mainly in Sweden. He was subsequently associated with Swedish electronics organizations until 1958 when worsening health led to his relinquishing regular employment. Mr. Petherick was elected an Associate of the Institution in 1946.

John Preedy (Graduate) who was born in London in 1934, died on 29th July 1962 as a result of a road accident. Mr. Preedy first joined the Institution as a registered student in 1957 and was transferred to Graduate Membership in 1958 after completing the Institution's examination requirements.

At the time of his death he was with the Ever Ready Company, being concerned with the development of transistor television receivers; from 1956-60 he held an appointment as an engineer with the British Broadcasting Corporation. He was 28 years of age.

James Hugh Paterson (Graduate) died on 13th January 1962. He was born and educated in Paisley and was 30 years of age at the time of his death. Elected in 1958, Mr. Paterson had held an appointment as an engineer with the British Scientific Instrument Research Association since 1957 until shortly before his death.

A High-speed Echo-sounder Recorder having Seabed Lock

By

R. W. G. HASLETT, Ph.D. †

Presented at the Symposium on Sonar Systems in Birmingham on 9th-11th July 1962.

Summary: An electro-mechanical recorder giving an intensity-modulated record on dry paper is described which indicates the presence of fish within $1\frac{1}{2}$ fathoms of the seabed (i.e. under the headline). Experiments were made with several possible types of triggered recorder to determine the best design, also with various signal delays required for seabed lock. Details are given of the moving-iron recorder, finally adopted, which displays a range of 4 fathoms with high accuracy across paper 6 in. wide. A magnetic drum storage system (described in an associated paper concerning the complementary c.r.t. display), is employed.

These displays form part of a comprehensive fish-detection equipment of improved acoustic performance, for use on distant-water trawlers.

1. Introduction

An expanded c.r.t. display (as previously described)¹ gives details of the shapes and amplitudes of the fish echoes and indicates fish and seabed echoes in stable positions, unaffected by vertical motion of the ship and changes in contour. Also, the use of repeated scan enhances the brightness and facilitates observation. It is of considerable assistance, however, to have a permanent record when making an estimate of the quantity of fish caught by the trawl over a long period (e.g. $2\frac{1}{2}$ hours).

The present paper describes the development of a triggered electro-mechanical recorder providing a greatly-expanded permanent record in the form of intensity modulations on dry paper. This high-speed recorder is used in conjunction with the c.r.t. display, both being triggered under the control of the seabed echo. The resulting records have the further advantages of accurate trace-to-trace correlation (allowing the characteristic traces of single fish and shoals to be recognized) and of "white line"² (so that fish-echoes very near the seabed echo can be distinguished).

Although the major improvement lies in the display, which can be used with existing types of echo-sounder, the performance of the acoustic system has also been improved.

The investigations which led to the choice of the final design may be divided into two categories: (a) the form of electro-mechanical recorder and (b) the type of signal delay required for seabed lock.

† Kelvin Hughes Division, S. Smith & Sons (England) Ltd., Ilford, Essex.

2. Requirements for the Electro-mechanical Recorder

The requirements of a seabed-locked recording scale-expander for use on trawlers are stringent, namely:

- (a) very high speed of stylus across paper (e.g. about 50 ft/s),
- (b) to use wide dry electro-sensitive paper (6 in. wide),
- (c) precise electrical triggering of mechanical movement,
- (d) the origin from which the stylus starts its movement must be well defined,
- (e) each stroke of the stylus must repeat accurately, and
- (f) the mechanism to be rugged and reliable over a long period.

3. Some Possible Types of Triggered Recorder

A number of different methods of obtaining the triggered sweep of the stylus across the paper were considered:

1. electro-magnetic clutch,
2. rotary solenoid,
3. friction clutch and stop,
4. moving-coil motor, and
5. moving-iron motor.

To determine the accuracy of repetition of stylus movement (under 2 (e) above), a special calibration signal was applied between stylus and platen, consisting of a series of d.c. pulses (with a 50/50 mark/space ratio) 1 ms apart, initiated by the same trigger pulse

as the recorder mechanism. With paper movement, this signal should give a series of parallel black and white "bars", under ideal conditions.

3.1. Electro-magnetic Clutch

A Warner electric clutch (Type SF-250)[†] was used as the basis of an experimental recorder. One side of the clutch was connected to the shaft of an electric motor rotating at constant speed, the other was joined to the stylus-arm. The clutch (2 $\frac{5}{8}$ in. dia.) normally requires a power of 15 watts at 28 V to energize it continuously, but in the experiments, a capacitor of 10 μ F was charged to 280 V and discharged through the clutch winding. Thus, on an intermittent basis, the speed of operation was greatly enhanced without overheating and a 4 in. stylus-arm could be accelerated to full speed in 3.5 ms, crossing the paper 6 in. wide in 17.5 ms. At each side of the paper, a spring "buffer" was mounted and the stylus was returned to the origin (resting against one of these buffers) by a weak spring.

Although robust, this device was extremely noisy and it was apparent that considerable wear would occur in time. Also, repetitions of the stylus movement varied by as much as ± 0.3 ms (measured at the right-hand side of the paper) and the speed of the stylus was affected by stylus pressure.

3.2. Rotary Solenoid

This experimental model employed a Ledex rotary solenoid type 5S \ddagger having a total angular movement of 35 deg and a nominal working voltage of 29 V. A capacitor (120 μ F) was charged to 280 V and discharged through the solenoid. The operating delay was 1 ms and the time of transit over 35 deg arc was 3.5 ms when attached to a stylus arm 4 in. long. Thus, the maximum deflection (2.4 in.) would be equivalent to 1.4 echo-fathom.

The stability of the origin was found to be reasonably good and the accuracy of repetition of the stroke was high. Unfortunately, operation was very noisy and vibration was excessive so that it was clear that wear would be serious in due course.

3.3. Friction Clutch and Stop

In this device, a stop held the stylus which was attached to a small, very light carriage, which slipped around the edge of a continuously-rotating disc. When the stop was withdrawn, the carriage clutched to the disc and moved across the paper at constant speed.

Due to the uneven wear which occurred around the disc and the time taken to withdraw the stop, this method was not promising.

[†] Manufactured by Westool Ltd., St. Helens Auckland, Co. Durham.

[‡] Made by N.S.F. Ltd., Keighley, Yorks.

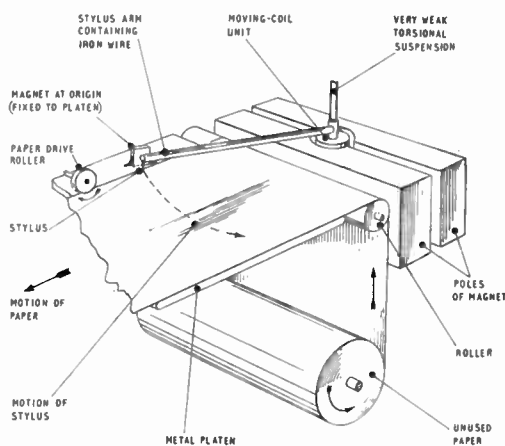


Fig. 1. The basic features of an early experimental triggered-recorder using the moving-coil principle.

3.4. Moving-coil Motor

This used a modified Kelvin-Hughes moving-coil pen recorder (see Fig. 1)³ in which the stylus arm is attached to a coil placed in a strong magnetic field. This arm (made of aluminium tube) was 6 in. long and the width of the paper was 3 in. A heavy-current pulse (lasting about 1 ms) was applied to the coil to accelerate the arm to the required speed. Thereafter, the stylus moved across the paper at approximately constant velocity. Finally, a second pulse was applied to the coil in the reverse direction and halted the motion. A weak torsional suspension returned the stylus to the origin. To ensure accurate alignment, a small magnet was mounted on the platen at the origin to attract a piece of iron wire inserted in the arm near the tip. The stylus was pressed against the paper by a spring at the root of the arm. The signals were delayed by means of a magnetic drum (see Sect. 4.3) and a gate in the amplifier feeding the stylus, allowed signals to pass only during the forward stroke of the stylus time-base (to prevent burning of paper when the stylus was stationary and to eliminate marking on the flyback). This recorder gave ranges of either 15 or 8 fathoms across 3 in. paper and was quite accurate.

Sea-trials were conducted in 1956. Typical results are seen in Fig. 2 in which (A) is the expanded seabed locked record and (B) is the normal echo-sounder chart taken at the same time. The paper speed is 0.6 in./min in both cases. In (A), the vertical motion of the vessel has been eliminated and groups of fish-echoes are clearly shown at ranges just less than that of the seabed echo.

The stylus movement was linear and repeatability of sweep was high, but the origin was not sufficiently stable and the stylus tended to lift off the paper before completing the time-base stroke. However, valuable

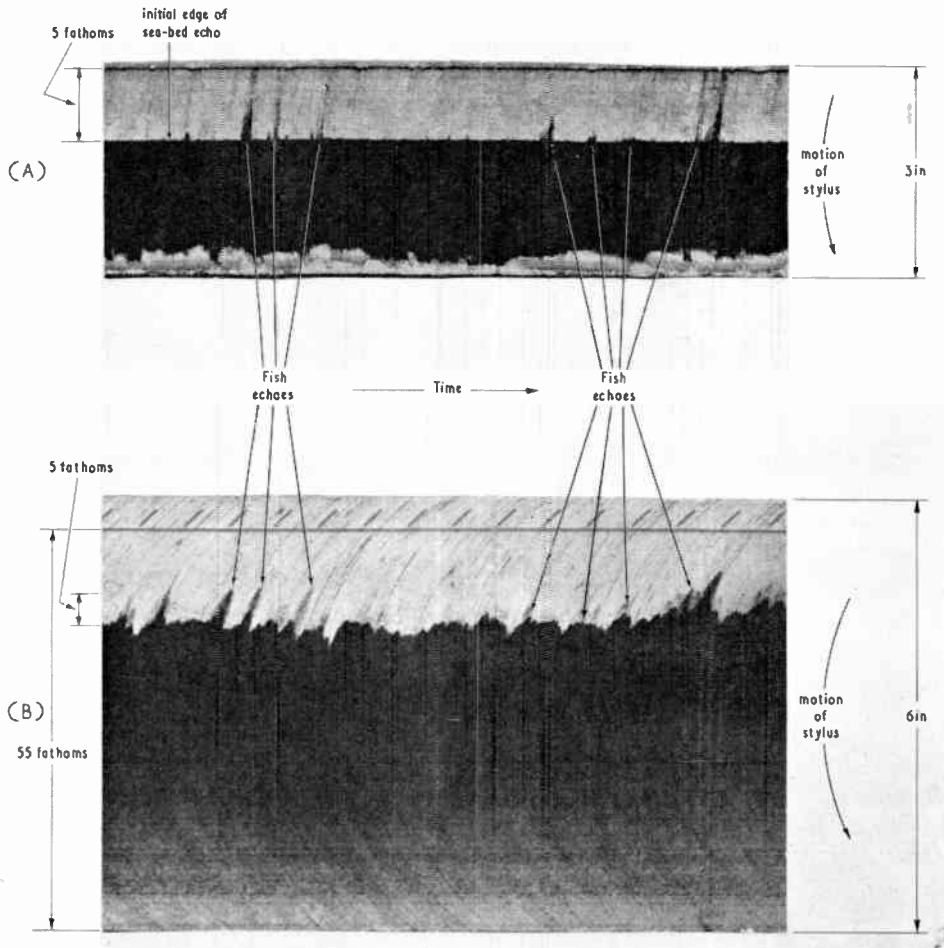


Fig. 2. (A) The record produced by an early experimental seabed stabilized recording scale-expander, compared with (B) the normal echo-sounder chart taken at the same time. (S.T. *Lancella* steaming at 13 knots over a rocky seabed off the Norwegian coast, 23rd September 1956. Depth: 160 fathoms; fish—herring. Transmission phased 145 fathoms in (B).)

experience was gained in the use of this type of display at sea but it was considered that the mechanism was somewhat too delicate for use by fishermen. In particular, the stylus arm and stylus were easily damaged but difficult to replace.

3.5. Moving-iron Motor

The movement which was finally adopted³ is now described. This uses a two-pole moving-iron motor (Fig. 3) embodying a stack of laminations 1 3/4 in. high. The stylus arm is very robust and is attached to the rotor of laminated soft-iron. When the stylus is at the left-hand side of the paper, the plane of the rotor is at an angle of 50 deg to the plane of the two poles. On passing a heavy-current pulse through the stator windings, the rotor accelerates so as to reduce the gap in the magnetic field. Before the plane of the rotor has reached the plane of the poles, the current pulse ceases, but rotation continues due to the momentum

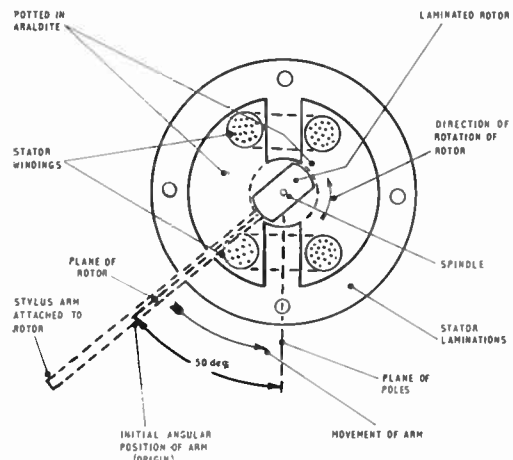


Fig. 3. Basic features of moving-iron motor to which a current impulse is applied.

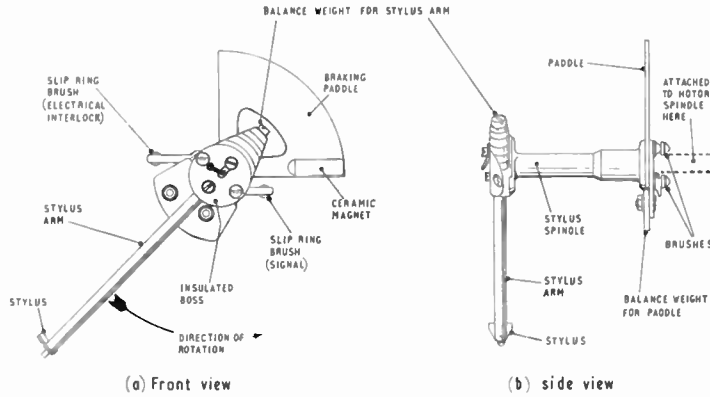


Fig. 4. Final recorder—stylus arm and braking-paddle assembly.

of the rotating assembly (Fig. 4). As the rotor is mounted in ball-bearings, friction is low and the velocity of the stylus across the paper is almost constant.

The trailing stylus returns to the origin by completing one rotation. The aluminium paddle (0.060 in. thick), attached to the stylus assembly, passes between the poles of a strong permanent magnet (flux: 14 000 maxwells) (Fig. 5) during the last quarter-rotation. The paddle has a high electrical conductivity and hence heavy braking occurs (in 300 ms). Finally, the small ceramic magnet, mounted in the trailing edge of the paddle, is strongly attracted to the larger magnet, without touching. The ceramic magnet is designed to fit the poles of the larger magnet leaving a clearance of 0.009 in., and the position of the stylus at the origin is very accurately determined; strong forces act to restore the stylus to the correct position should there be a temporary deviation (a magnetic “detent”). When the stylus is “caught” at the origin in this way, a small amount of damped oscillatory motion occurs, but at this point the stylus is off the paper so that no damage results. An adjustment is provided (Fig. 5)

for the amount of eddy-current braking. A Mumetal screen surrounds the braking magnet as far as possible, reducing the safe compass distance to 7 ft.

The stylus/rotor assembly (Fig. 4) is very carefully balanced, both statically and dynamically. It is essential to balance the paddle and stylus arm (4 in. long) individually by separate balance weights to ensure long life for the ball-bearings (Fig. 5).

The inner slip-ring (Figs. 4 and 6) supplies the signal to the stylus arm which is insulated from the rotor spindle. The outer slip-ring operates an electrical interlock circuit, which has two functions: to disconnect the stylus drive if the stylus is not at the

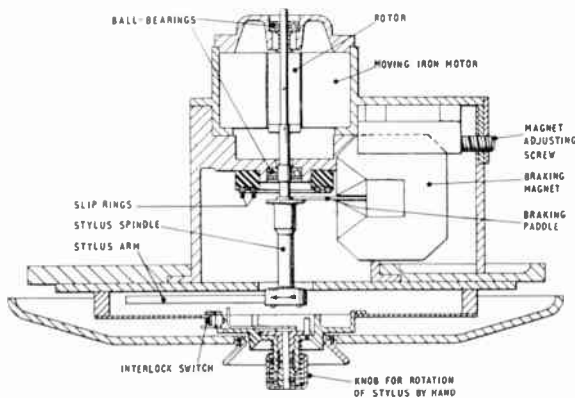


Fig. 5. Sectional plan view of final triggered-recorder, used as a recording scale-expander.

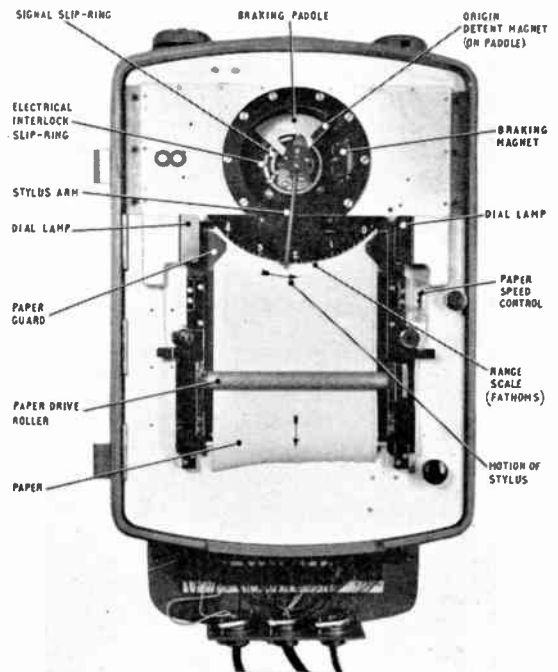


Fig. 6. Final recorder—front view (with cover and manual knob removed to show braking paddle and slip rings).

origin and to generate a pulse when the stylus has begun to move away from the origin (see Sect. 5.1 and Fig. 9(g)).

The stylus consists of short length of tungsten wire (0.010 in. dia.) in the form of a single-loop spring and to replace it, the arm may be rotated by hand (Fig. 5). When this is done, the electrical interlock is operated so that the stylus drive current cannot be generated and the signal voltage on the stylus is disconnected. The mechanism for hand rotation is designed to spring clear of the stylus when not in use.

The origin of movement is off the left-hand side of the paper. After triggering, the stylus takes 4 ms to reach full speed (i.e. an acceleration of about 400 g), by which time it is over the left-hand edge of the paper (Fig. 6). The width of the paper (6 in.) is crossed in 10 ms (corresponding to a range of 4 fathoms). The range scale is calibrated in terms of distance above the seabed and is slightly non-linear (but does not depart by more than $\pm 4.6\%$ from linearity). The seabed echo is recorded on the paper at "zero" on the range scale.

The recorder may be triggered at rates up to 200 scans/min but is normally used at 100 scans/min. The accuracy of stylus position at the origin is ± 0.004 in. and repetition of stylus movement is within ± 0.006 in. of the mean position measured at the right-hand

side of the paper by means of the bar calibration signal. (The latter corresponds to a change of range of 0.3 in., equivalent to 10 microseconds echo-time.)

The stylus moves over an insulated platen and contact is made to the paper via a conducting portion of the platen $1\frac{1}{2}$ in. lower down and through the top paper roller (under the range scale in Fig. 6). This roller has insulated sections at each end under the arc of stylus movement. In this way damage to platen and roller by excessive signal levels is prevented.

This recorder has been found to be very robust.

4. Some Possible Types of Signal Delay

Three different methods of obtaining the signal delay (necessary to display fish-echoes which arrive before the seabed echo) were investigated:

1. acoustic delay-line,
2. electrical delay-line, and
3. magnetic-drum delay.

4.1. Acoustic Delay-line

It was decided to approach the matter empirically. Experiments were made with 19 different delay lines using the apparatus seen in Fig. 7. Two signal sources were available, either from apparatus which simulated the actual shape of a fish-echo or from the

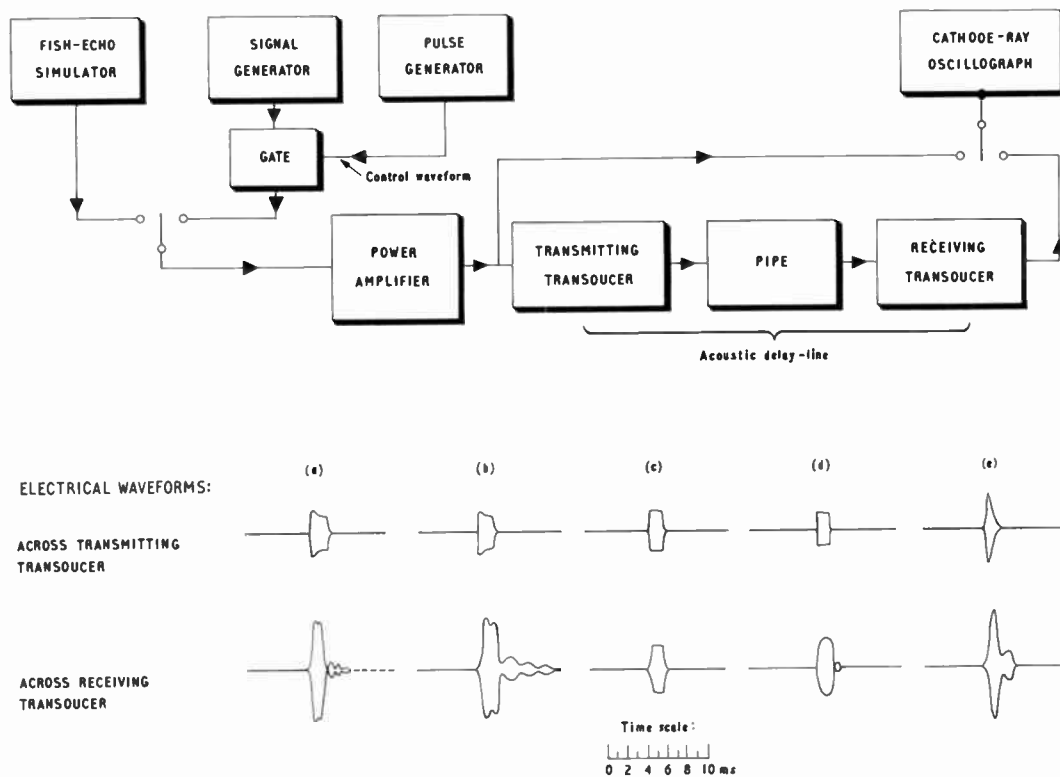


Fig. 7. Some of the better results of the experiments with various acoustic delay-lines. (Further details are given in Table 1.)

gated signal generator which gave a pulse of high-frequency waves having an approximately-square envelope. After amplification, the signal was applied to the transmitting transducer at one end of the pipe (constituting the acoustic delay line). At the far end, the receiving transducer re-converted the acoustic waves to electrical signals. The waveforms across the transducers at input and output of the line were observed on the cathode-ray oscillograph.

In most cases, the medium used for delay was air, in which the velocity of sound is approximately 1 ft/ms. This is readily obtainable and cannot leak as can other fluids, also the velocity is reasonably low so that a delay line to give a delay of 10–15 ms is not too big physically. The following "pipes" were tried (the internal dimensions are given in each case):—

- (i) rubber or p.v.c. hose-pipes, dia. $\frac{1}{2}$ in. or $\frac{5}{8}$ in.,
- (ii) steel conduit, dia. $\frac{1}{2}$ in. or $1\frac{1}{8}$ in.,
- (iii) 3 cm-radar copper waveguide, $1 \times \frac{1}{2}$ in.,
- (iv) copper tube, dia. 1 in.,
- (v) 9 × 9 in. steel box (5 ft long), padded internally with cotton wool, leaving $4\frac{1}{2} \times 4\frac{1}{2}$ in. clear inside,
- (vi) no pipe at all (in laboratory),
- (vii) a solid block of p.v.c. foam, $\frac{1}{2} \times \frac{1}{2}$ in., and
- (viii) a solid block of butyl rubber (Polysar butyl 100), $\frac{1}{2} \times \frac{1}{2}$ in.

Also various transducers were used:—

- (i) Acos sound cell, type MIC.25–1†
- (ii) an electrostatic loudspeaker, type LSH75,‡
- (iii) Goodmans Trebax loudspeaker,§ and
- (iv) a small 30 kc/s magneto-strictive transducer.

These were selected because they had some output in the range of frequency up to 30 kc/s. The Acos sound cell is piezo-electric (Rochelle salt 0.8 × 0.4 in.). The electrostatic loudspeaker consists of a plastic film (2 × 2 in.), metallized on one side, pressed against a perforated metal plate and is externally polarized. The Trebax is a high-frequency moving-coil unit with a horn.

Where air was the delaying medium, lengths of pipe between 6 ft and 12 ft were used. Experiments were made on the flexible pipes both straight and coiled in helices of diameter 8 in. The received pulse-shape did not seem to alter much between these two conditions. When the output pulse-shape depended critically on the frequency, the frequency which gave the best results in the range 19–33 kc/s was chosen.

† Made by Cosmocord Ltd., Waltham Cross, Herts.

‡ From Technical Suppliers Ltd., 63 Goldhawk Road, London, W.12.

§ Made by Goodmans Industries Ltd., Wembley, Middx.

Table 1

Details of Transducers, Acoustic Delay-line Pipes and Measurements for which the Waveforms are shown in Fig. 7. (see Sect. 4.1)

Transducers	Pipe	Fre- quency (kc/s)	Overall attenua- tion (dB)	Signal/ noise ratio	Delay approx. (ms)
(a) Acos sound cell	12 ft p.v.c. hose, $\frac{5}{8}$ in. dia. bore, coiled.	30	120	15/1	11.2
(b) Acos sound cell	8 ft steel pipe, $\frac{1}{2}$ in. dia. bore straight.	30	110	30/1	7.8
(c) Electro- static loud- speaker	6 ft p.v.c. hose, $\frac{1}{2}$ in. dia. bore, coiled.	30	108	100/1	6.0
(d) Trebax loud- speaker	12 ft p.v.c. hose, $\frac{5}{8}$ in. dia. bore, coiled.	19.6	58	50/1	11.5
(e) Electro- static loud- speaker	No pipe, trans- ducers 5 ft apart	30	100	16/1	5.0

When the frequency was not too critical, 30 kc/s was used, as this was the frequency of operation of the echo sounder. The resultant delays varied between 4.5 and 11.5 ms. The mean velocity of sound in the air-filled pipes was 1.04 ft/ms.

Blocks of p.v.c. foam and butyl rubber were considered because the velocities of sound through these materials were so low (approx. 1 in/ms and 0.4 in/ms respectively) that the delay-line would be very small physically and therefore more convenient than air. Unfortunately, in both cases the received pulse was highly distorted.

All the delayed signals had appreciable tails. Five of the better examples are seen in Fig. 7. (Further details of these are given in Table 1.) The input pulse-length was approx. 1.5 ms. Results were difficult to repeat and were sensitive to small changes of frequency and to slight changes in the positions of the transducers relative to the ends of the pipe. Most of the attenuation occurred in the transducers.

In one case only was reasonable fidelity obtained (Fig. 7(c)).

4.2. Electrical Delay-line

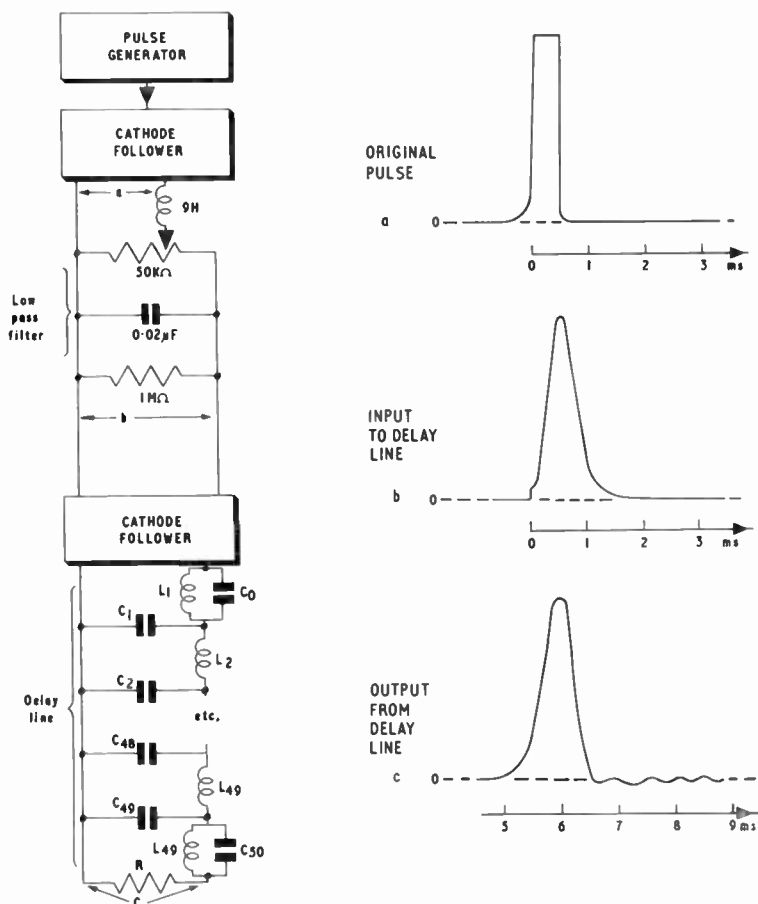
These experiments are designed to explore the feasibility of using a simple delay-line which could be manufactured with comparative ease.

The high-frequency bandwidth of the echo sounder is approx. ± 2 kc/s (centred on 30 kc/s). After

Fig. 8. The experimental electrical delay-line and the wave-forms observed at various points.

$L_1 = 32.7$ mH; L_2 to L_{49} —descending values between 109.0 mH and 105.5 mH; $L_{50} = 31.65$ mH.

$C_0 = 0.0634$ μ F; $C_1 = 0.0952$ μ F; C_2 to C_{48} —descending values between 0.119 μ F and 0.114 μ F; $C_{49} = 0.0912$ μ F; $C_{50} = 0.0608$ μ F; $R = 959\Omega$.



detection, the bandwidth is 0.2 kc/s so that the minimum cut-off frequency (f_c) of the delay-line would be $2\sqrt{2}$ kc/s = 2.83 kc/s. Using constant- k sections:

$$Z_0 = L^{\frac{1}{2}}/C^{\frac{1}{2}}; f_c = 1/[\pi(LC)^{\frac{1}{2}}]$$

and $\delta \simeq (LC)^{\frac{1}{2}} = 1/(\pi f_c)$ (when losses are negligible), where Z_0 = characteristic impedance of line (ohms), L and C are series inductance (H) and shunt capacitance (F) of each lumped element and δ = time delay per section (seconds). Substituting for f_c , $\delta = 0.112 \times 10^{-3}$ and 48 sections would give a delay of 5.4 ms.

In the electrical delay-line which was actually made up (Fig. 8), the values of L and C varied by $\pm 1.6\%$ and $\pm 2.1\%$ respectively and $f_c = 2.84$ kc/s. Two m -derived matching half-sections ($m = 0.6$) were added at the input and output of the line. The calculated total delay at low frequencies is 5.43 ms. The components were arranged in descending order of inductance and capacitor values so that the impedance (Z_0) remained more-nearly constant along the line. The size of the line was $6 \times 10 \times 14$ in. and the weight 20 lb.

Continuous-wave measurements indicated that the attenuation varied from 3 to 7 dB over the working

band (0.2 kc/s); it was 17 dB at the cut-off frequency (2.84 kc/s) and 65 dB at 3.15 kc/s. The phase response began to deviate from the ideal straight line when the frequency exceeded 1.6 kc/s. The delay at the cut-off frequency was 1.5 ms greater than that calculated above. (The phase response could have been improved by providing inductive coupling between the inductances but this would increase the cost of manufacture.)

For pulse measurements, a rectangular pulse lasting 0.5 ms (Fig. 8(a)) was generated. The frequency components contained in this pulse were restricted as in an echo sounder by means of the low-pass filter which was adjusted to give the rounded waveform (b). This pulse, having a duration of 0.6 ms (50% amplitude), was applied to the delay line. The output waveform (c) is a replica except that an oscillation of amplitude 7% (peak to peak) of the main pulse is added to the tail which decays to 3.5% (peak to peak) in 1 ms.

The dynamic range of the delay line was very high (180 dB) and as totally-screened Mullard LA4 pot-cores were used for the inductors, the hum pick-up level was very low.

Re-cycling the pulse in the delay line in order to obtain repeated-playback for the c.r.t. display, would not be possible owing to the cumulative distortion. (Also re-modulation would have been required.) However, the line might have been suitable for the recording scale-expander, alone, although the tail from a shoal echo might appear to be a smaller group of fish nearer the seabed. It was clear that manufacture in quantity would be difficult.

4.3. Magnetic Drum Delay

A magnetic drum delay had been developed in parallel with this work, for use with a seabed stabilized c.r.t. scale-expander. This type of delay was chosen for the c.r.t. because a storage facility can be obtained readily and provides an important advantage, namely repeated play-back of the signals on to the c.r.t.

TRIGGER SELECTION UNIT :

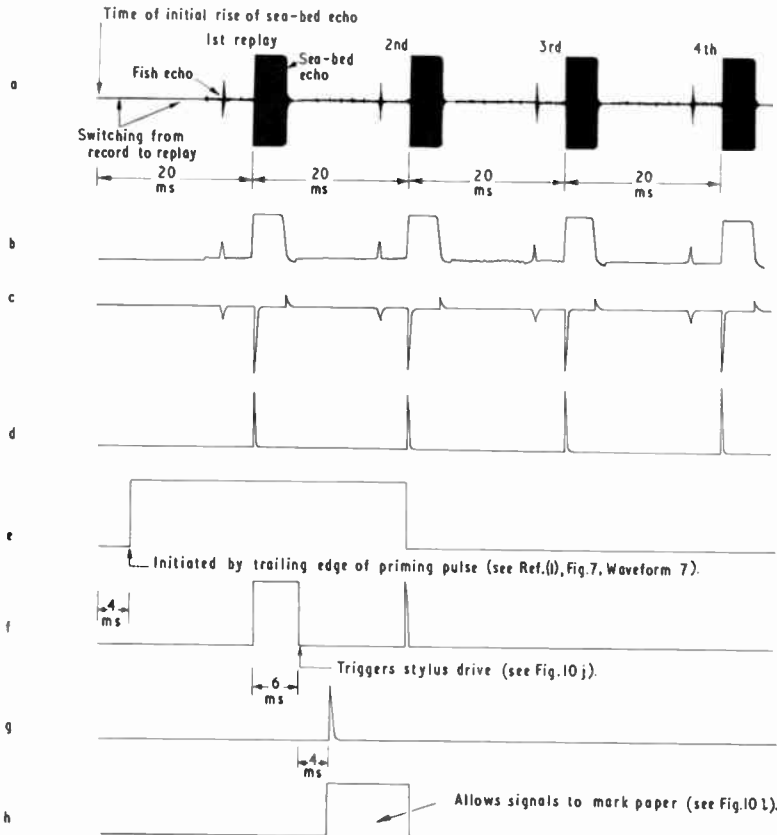
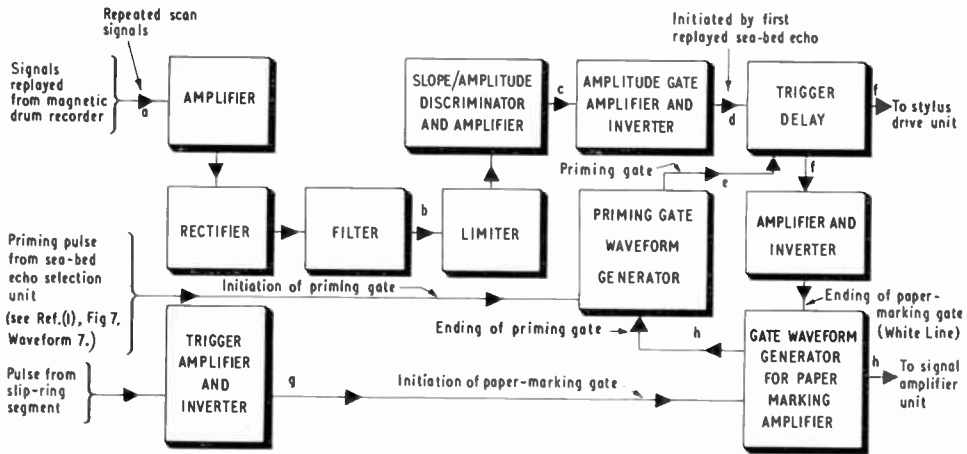


Fig. 9. Final recording scale-expander—block diagram of trigger selection unit and associated waveforms.

This recording drum coated with magnetic material, rotating at 3000 rev/min, has been described in detail elsewhere.¹ The recording/replay and erase heads are not in contact with the drum and the dynamic range is about 50 dB.

The decision was taken to use this device in a comprehensive fish-detection equipment embodying both the cathode-ray display and the recording scale-expander as its performance is ample for the latter. The drum has the following additional merits:

- (i) it records direct at 30 kc/s,
- (ii) fidelity is high,
- (iii) repeated playback is available for the c.r.t. display without further distortion, and
- (iv) it is compact and reliable.

5. Final Circuits Associated with Recording Scale-expander

These may be divided into three parts:

- (i) trigger selection unit,
- (ii) stylus drive unit, and
- (iii) signal amplifier unit.

The total power consumption is 150 W.

5.1. Trigger Selection Unit (Fig. 9)

Signals replayed from the magnetic drum recorder (waveform *(a)*) are amplified, rectified and smoothed to give *(b)*. This wave is differentiated and amplified in the slope/amplitude discriminator⁴ (waveform *(c)*) so that trigger edges from the seabed echoes, only, are allowed to pass through the amplitude gate (waveform *(d)*) and then only when the seabed echo is large and has a steep initial rise. (Accuracy of triggering from the seabed echo is normally within $\pm 15 \mu\text{s}$.) Large fish echoes are ignored. These selected signals can be effective in initiating the trigger delay only when the latter is primed by the priming gate waveform.

The "control waveform" which is generated in the seabed echo selection unit (as described in the associated paper¹), is used as a priming pulse which rises 4 ms after the initial edge of the seabed echo (and recovers 550 ms later). This rise triggers the priming gate waveform generator (*(e)*) so that the trigger delay is initiated by the first replayed sea-bed echo of waveform *(d)*. The trigger delay recovers after 6 ms (*(f)*) (but remains primed by waveform *(e)*). The trailing edge of *(f)* triggers the stylus drive. This edge, however, produces no effect at this time on the gate waveform generator for the paper marking amplifier (wrong polarity).

Four milliseconds after the stylus has started to move (see *(g)*), a pulse is received from the interlock slip-ring (corresponding to the stylus passing over the

left-hand edge of the paper) and this pulse triggers the gate waveform generator for paper marking (*(h)*).

When the second replayed sea-bed echo arrives (*(d)*), the trigger delay (which is still primed) starts to execute a second cycle of operation (*(f)*). This initial edge shuts down the paper-marking gate waveform (*(h)*), which, in turn, ends the priming gate waveform (*(e)*).⁵ The trigger delay is no longer primed and its second cycle is cut short abruptly (*(f)*) so that the resultant "pip" is only 100 μs long (corresponding to 3 in. echo-range). This results in an accurate "white line"² on the recording scale-expander chart, the initial rise of the sea-bed echo giving a mark only 0.040 in. long. The third and later replays of the seabed echo have no effect on the trigger delay.

5.2. Stylus Drive Unit (Fig. 10)

The trigger waveform for the stylus drive (*(f)* in Fig. 9) is only allowed to pass through the trigger inverter and gate when the electrical interlock is complete and the equipment has warmed up. This gate is shut if the stylus is not at the origin, when the manual knob is pressed and when the paper tank is opened.

When triggered, a thyratron (CV2253) discharges a capacitor (10 μF charged to 520 V stabilized) to produce a heavy-current pulse (Fig. 10(j)) through the stator of the moving-iron motor. The resulting deflection of the stylus is seen in graph *(k)*. The stylus takes 4 ms to accelerate and pass over the left-hand edge of the paper (when the slip-ring segment generates the pulse *(g)* in Fig. 9).

5.3. Signal Amplifier Unit

The signals replayed from the magnetic drum are mixed with the time and fix marks (when required) but are only allowed to pass to the paper (*(l)*₁) when the paper-marking gate (*(h)*) is open, i.e. when the stylus is passing over the paper. Thus a range within four fathoms (10 ms) of the seabed echo is recorded on the second replay from the magnetic drum. The output stage supplying signal to the stylus, consists of a pentode valve (type EL360) with an h.t. of 900 V. The power required to saturate the paper is about 100 W as the stylus velocity is so high. The stylus pressure is 0.9 oz (27 g).

Signals for time and fix marks are generated by the gated oscillator (frequency 1 kc/s), triggered by waveform *(g)* from the slip-ring segment. This waveform is arranged to be either short for time marks so that one cycle, only, is produced, or long for fix marks so that a large number of oscillations occur (*(m)*) (decaying slowly). The time marks are switched on once a minute by a segment on the paper-drive motor, whilst the fix marks are controlled by a push-button. Waveforms *(m)*, *(n)* and *(l)*₂ illustrate the effect of depressing this button. Half-cycles of 1 kc/s are produced

(m). These are added to the h.f. signals (n) and a dotted fix-mark results.

6. Improvements to Acoustic System

Although the major improvement to the comprehensive fish-detection equipment for use on trawlers is the use of seabed locked displays, steps have been

taken to improve the acoustic performance of the system as well.

First, a valve transmitter was designed to deliver a peak power of 8.2 kW to the transducer (pulse-length: 0.5 to 2.5 ms). Secondly, the acoustic beam was made narrower than usual by increasing the size of the radiating face of the transducer to $6\lambda \times 4\lambda$ at

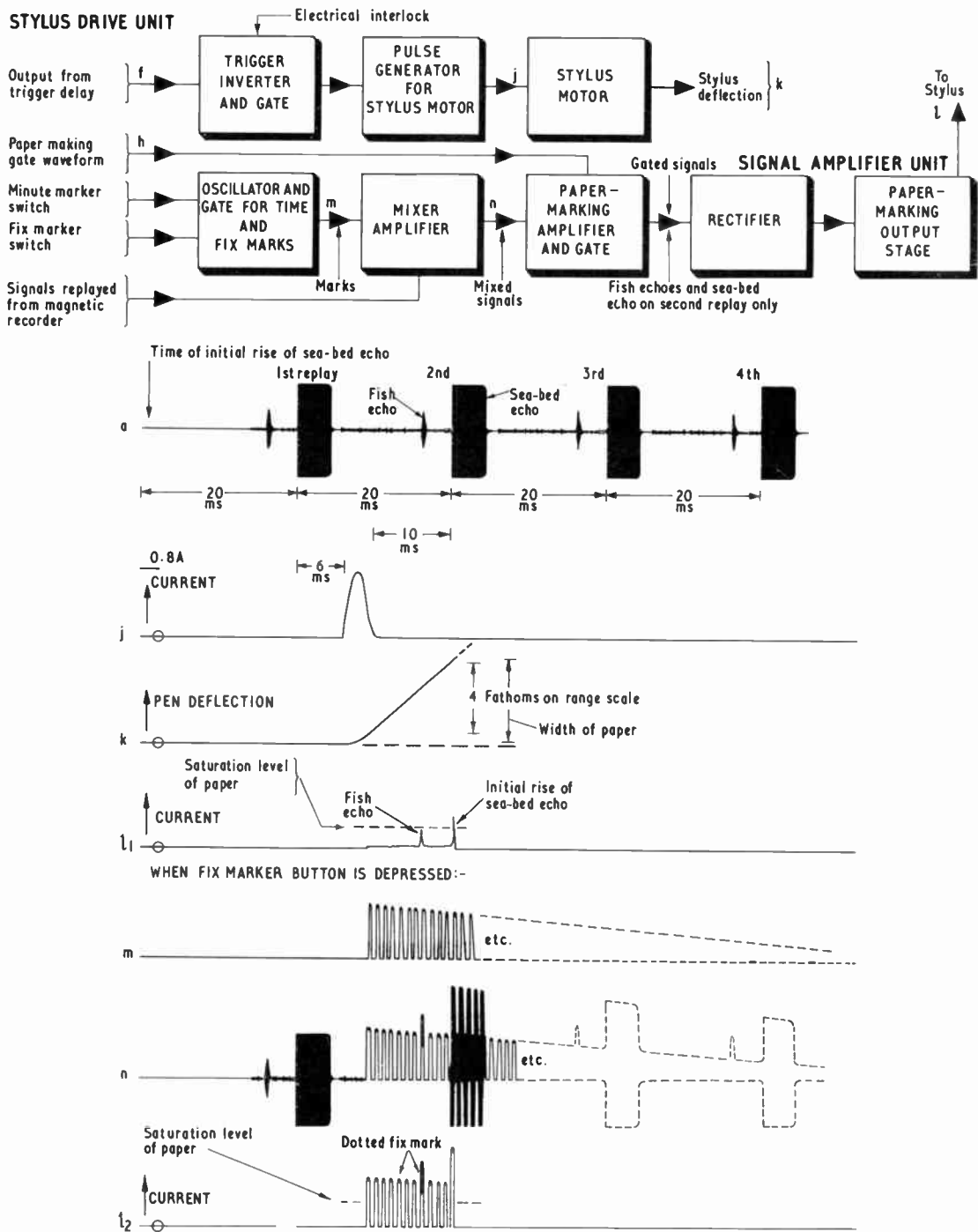


Fig. 10. Final recording scale-expander—block diagrams of stylus drive and signal amplifier units and relevant waveforms.

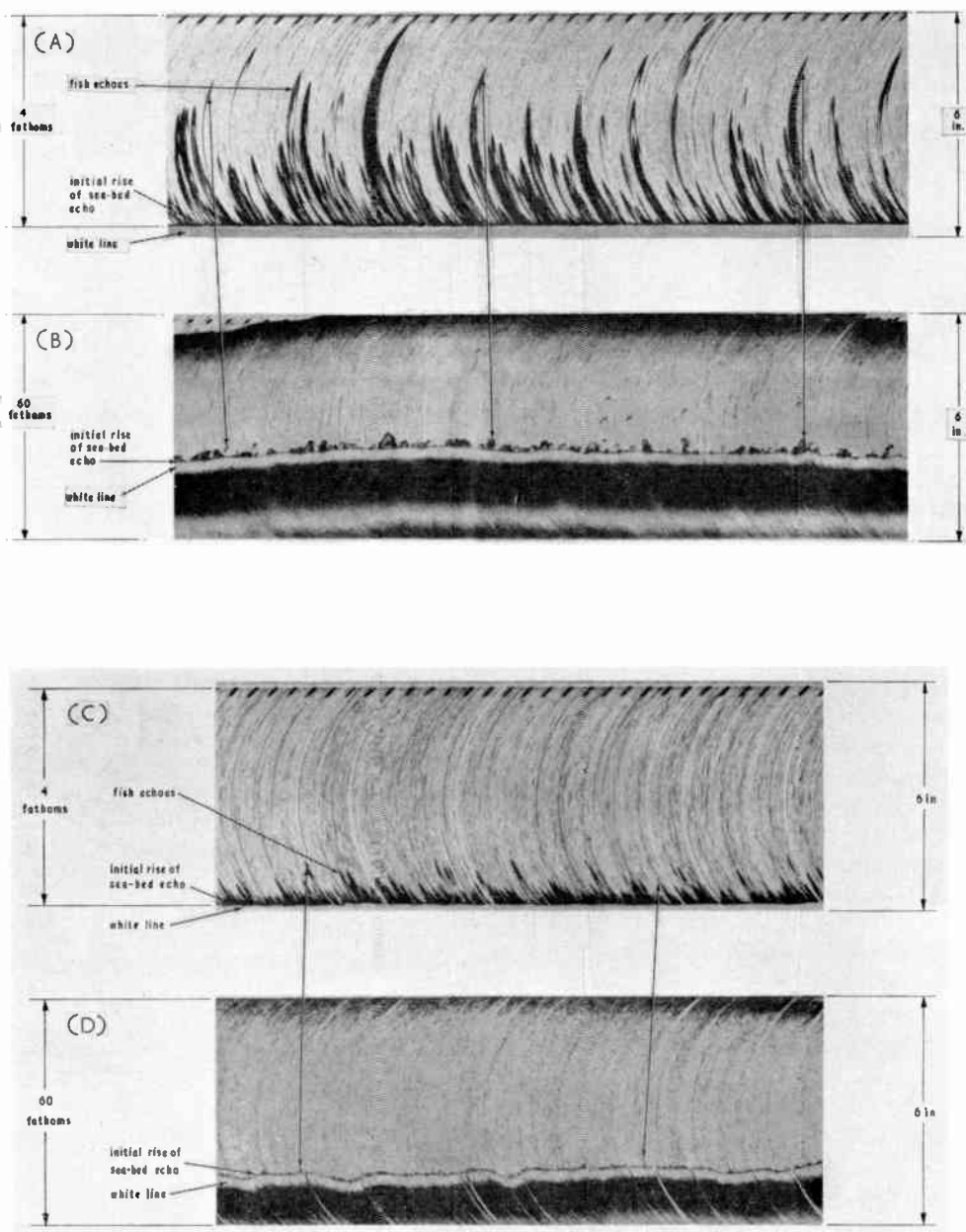


Fig. 11. Typical charts obtained with the final recording scale-expander, compared with corresponding normal echo-sounder charts taken at the same time.

Numerous fish-echoes are seen above the seabed echo; some points corresponding in time are indicated:

- (A) scale-expander chart } depth 100 fathoms; fish—cod and haddock; 30 baskets/hour.
- (B) normal chart }
- (C) scale-expander chart } depth 50 fathoms; fish—haddock; 70 baskets/hour.
- (D) normal chart }

(M.T. *Westella* trawling at 4 knots, July, 1960.)

30 kc/s (30×20 cm). Thirdly, the transducer was mounted on a retractable shaft so that it could be lowered 2 ft below the ship when using the equipment. In this way the effect of aeration which follows the skin of the vessel is reduced.

7. Results of Sea Trials

Extensive trials of the full equipment have been conducted on trawlers off Iceland, Norway and the north-west coast of the U.S.S.R.

Under the worst conditions of fishing, the largest angle of roll was less than 20 deg for 75 per cent of the time (and pitch was less than this) so that the use of the unstabilized narrow-beam transducer is justified.

When the acoustic beam passes over a single fish, its echo is indicated as a characteristic half-hyperbola (or "comet") on the recording scale-expander. Such echoes are observed quite clearly at great depths. Typical records (Fig. 11) show that the irregularities in the fish echoes are eliminated by the seabed lock, so that the trace-to-trace correlation on fish echoes is greatly enhanced.

These records also illustrate the difficulties of estimating the catch, for, at first sight, it appears that more fish would be caught in respect of (A) than in (C), whereas the reverse is the case. Experience over the last two years has shown that larger catches occur when all the fish-echoes are within one fathom of the seabed echoes. This condition is almost invisible on the normal echo-sounder chart but may be observed readily on the recording scale-expander. The particular places and depth contours at which fish occur can be determined in relation to the time marks and the normal echo-sounder chart, so that the ship can return to the same place.

The new equipment gives a much more reliable indication than hitherto, either, when searching, of

the quantity of fish occurring at a catchable depth or when actually trawling, of the time for which the net should be left down.

8. Acknowledgments

In addition to Mr. J. E. L. Sothcott (who conducted the experimental work on the recorder) and others mentioned in the references, several other engineers, designers and instrument makers took part in this development, in particular Mr. P. Hall (mechanical design) and Mr. G. H. Ellis (sea trials).

The extensive facilities made available by J. Marr and Sons Ltd., Hull, on the M.T. *Westella*, S.T. *Lancellia* and other vessels, are gratefully acknowledged.

The author is indebted to the Directors, Kelvin Hughes Division of S. Smith & Sons (England) Ltd., for permission to publish this paper and would like to thank Messrs. Halliday, Sothcott, Hopkin and Ellis for valuable discussions whilst it was being prepared.

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DISCUSSION

Mr. D. Davies†: This equipment could if suitably modified have uses in deep sea work. It is often necessary to lower apparatus to within a fathom of the bottom for oceanographic research purposes and it is usual to attach a "pinger" to the cable just above the apparatus. The resultant signals are the direct and reflected pulses, and they converge on lowering. The pulses are received and displayed on a precision depth recorder receiver and display

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unit. If it were possible to lock into the seabed echo and display on a larger scale, the control of cable raising and lowering would be a much easier procedure and the display could be read directly by the winch-driver rather than the present procedure, in which the record has to be "interpreted" before instructions can be given.

Author's reply: I am grateful to Mr. Davies for drawing attention to a further use to which the triggered recorder may be put.

Artificial Cardiac Pacemakers for the Long-term Treatment of Heart Block

By

J. G. DAVIES†

Presented at the Symposium on "Practical Electronic Aids for the Handicapped" in London on 29th March 1962.

Summary: Electronic cardiac stimulators using blocking oscillators are described. The units operate (1) through wires from an external unit, (2) internally with external power supply and (3) as a completely self-contained implanted unit which can be controlled from outside the body by inductive pick-up.

1. Introduction

Over the past ten years many attempts have been made to control electrically the action of the heart in man suffering from a conduction defect of the Bundle of His. Impulses from the physiological pacemaker are blocked by disease and the heart beats too slowly and unreliably, often with periods of standstill causing attacks of loss of consciousness. The methods of artificial stimulation that have been used fall into two groups and are termed external and internal. In the former technique electrodes are applied to the outside of the chest and a comparatively large pulse of up to 200 volts is used.¹⁻³ Although these electrodes are readily applied, the main disadvantage is discomfort to the patient. For long periods of stimulation it is preferable to have at least one electrode sited on or in the heart as this enables the voltage to be reduced to the range of 1-5 volts. The other electrode may be placed anywhere on the surface of the body or buried in the tissues.

Internal stimulation falls into the following groups:

- (a) By means of wires attached to the heart and brought out to an external pacemaker.^{5, 7, 8, 9}
- (b) By an electrode catheter passed along a vein and terminating in the right ventricle.⁶
- (c) An implanted unit which receives its power from an external source.^{4, 11, 12, 13, 14, 18}
- (d) A completely self-contained implanted unit.^{10, 15, 16, 17}

It is desirable to have control of rate and voltage and this is readily achieved in groups (a), (b) and (c). Also it is occasionally an advantage to be able to increase the pulse duration and this presents no fundamental problem with these units. However, in groups (a) and (b) there is always the possible hazard of infection entering where the wires pass through the skin into the body. This objection is removed in group (c) as

transmission takes place across the intact skin but the transmitting coil must be accurately positioned over the receiving coil. The advantage of group (d) is that the patient is unhampered by external attachments but control over rate and voltage is more difficult.

2. External Pacemaker

Initially the problem that was presented was for a stimulator to be used externally but attached to wires on the myocardium. For economy of battery power a blocking oscillator circuit was chosen and this was required to give a pulse of 2 milliseconds duration at a repetition rate of between 30 and 120 per minute. The pulse had to have a maximum voltage of 12 V with a low output impedance. It had also to be stable over a temperature range of 0°C-50°C. Figure 1 shows the circuit of the unit. Silicon transistors are used throughout.

Whilst pulse duration is governed by the transformer, capacitor C1 and, to some extent, by the base voltage of the first transistor, increased loading on

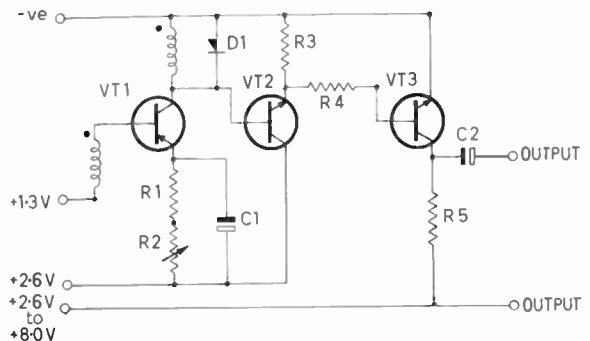


Fig. 1. Circuit diagram of a simple pacemaker.

R1 100 kΩ	C1 47 μF	D1 OA47
R2 1 kΩ	C2 22 μF	Transformer:
R3 10 kΩ	VT1 2S322 or OC201	Core FX 1011 pri. 950 turns
R4 1 kΩ	VT2 2S701	sec. 250 turns
R5 2.7 kΩ	VT3 2S711	48 s.w.g.

† St. George's Hospital, London, S.W.1.

the transformer tends to decrease the pulse length. It is for this reason that the first stage is coupled to an emitter-follower before the signal is fed to the output transistor VT3. Pulse repetition rate is controlled by resistance R2. The blocking oscillator using silicon transistors tends to increase its repetition rate with increasing temperature and this may be counteracted by using a germanium diode for D1. Output impedance is dependent upon the transistor VT3 which should be of the medium-power type with a low saturation resistance. Control of output voltage is achieved simply by driving the final transistor from a separate battery and selecting a suitable tapping point. A number of these units have been constructed all of which have functioned satisfactorily. The size of the unit is 65 mm x 40 mm x 20 mm and the batteries (RM1 Mallory cells) are renewed every six months. For portable equipment which patients carry in a belt, Mallory batteries are used but where space is not so critical cells whose state of discharge can be checked are employed.

3. Implanted Unit—Externally Powered

An implantable unit which receives its power from an external source (group (c)) has the advantage of allowing complete control over pulse rate and voltage. This normally takes the form of two coils, one internal and the other external which are placed in close proximity with each other. The stimulus is then transmitted to the heart along wires. Recently in Italy Cammilli *et al.*¹⁸ have reported placing the receiving coil on the heart. In order to allow greater latitude in the placing of the external coil a receiver has been designed to be implanted within the rectus sheath, whilst the transmitting coil is positioned around the patient's waist. This arrangement enables the belt to be moved ±4 inches with little effect upon the stimulation received. Figures 2 and 3 show the circuit diagram of the receiver and the transmitter. The latter is a squegging oscillator, the repetition rate of which is controlled by the variable resistor. Squegging duration is determined by capacitor C1 and to an

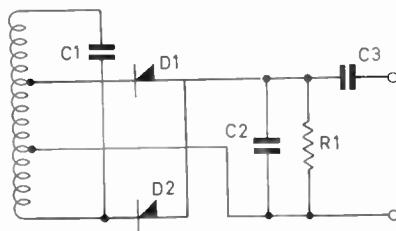


Fig. 2. Circuit diagram of receiver pacemaker.

- | | | |
|------------|-----------|---|
| C1 0.03 μF | D1 OA47 | Receiving coil: |
| C2 0.03 μF | D2 OA47 | 2-in Ferrox rod 3/8" diameter. |
| C3 22 μF | R1 4.7 kΩ | 120 turns with tappings at 30 and 60 turns. |

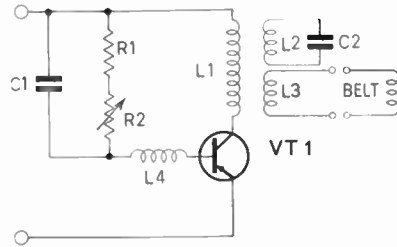


Fig. 3. Transmitter for implanted pacemaker.

- | | | | |
|------------|----------|--------------|-----------|
| C1 100 μF | R1 10 kΩ | L1 10 turns | } LA 2403 |
| C20-001 μF | R2 50 kΩ | L2 105 turns | |
| VT1 OC35 | | L3 3 turns | |
| | | L4 2 turns | |

extent upon the loading of the belt. However, the duration is readily fixed at between 2-4 ms. The operating frequency was chosen to be 20 kc/s as this was well within the range of transistors that were readily available at the time. The transmitting coil is an endless loop 42 inches in circumference and this is placed around the patient's waist and the excess folded back. A standard length is desirable in order to eliminate plugs and sockets and to have a coil which will fit most people.

The receiver coil is wound on a ferrite rod and full-wave rectification of the transmitted signal is achieved by the two diodes. It is, of course, important that the ferrite aerial is at right angles to the plane of the transmitting coil. However, this is not a problem as the receiver normally lies in the correct plane in the rectus sheath. Power requirements for the transmitter are considerably larger than for the units used directly as in groups (a), (b) and (d). The transmitters that have been constructed measure 4 in x 2½ in x 1½ in and have a useful life of 14 days.

4. Implanted Self-contained Unit

By combining the units shown in Figs. 1 and 2 into one, a more versatile apparatus may be constructed. This belongs to group (d) and has the safety factor that it may be used as a receiver (group (c)) should the components fail or the batteries become exhausted. Some control over pulse rate is desirable and this feature is incorporated. Figure 4 shows the circuit of the complete apparatus. The stimulator consists of a blocking oscillator, emitter follower and output stage as has been described for groups (a) and (b). The load resistor is made up of R4 and R5. Each pulse is then made to charge capacitor C2 through the diode D2 and this potential is applied via the feedback winding to the base of the first transistor. Also connected to C2 is the transistor VT4. All the transistors and diodes with the exception of D1 are silicon types.

Under normal conditions the stimulator is free running at a rate governed by R1, C1, R4 and R5.

However, if a signal is applied to L1 the transistor VT4 conducts and discharges C2 with the result that the blocking oscillator stops and in this state the circuit is stable. To start the stimulator again it is necessary to charge C2 and this is achieved by pulsing into L2. When C2 has charged to a predetermined level through R4 and R5 the blocking oscillator starts again and will continue in this stable condition. If the rate is required to be increased a continuous signal into L2 will raise the voltage across C2 and this in turn will cause the repetition rate of the blocking oscillator to be increased. Conversely a continuous signal lowering the potential on C2 but not of sufficient amplitude to discharge it completely will make the blocking oscillator run more slowly than normal. However, once the influence on L1 or L2 is removed the stimulator will return to its predetermined rate. In practice L2 is tuned to 20 kc/s and L1 to 150 kc/s. The output is taken from points B and D (Circuit diagram Fig. 4). Should the batteries become exhausted it is possible to pulse into L2 by means of a belt connected to a transmitter of the type described for group (c) and so deliver enough power to B and D to maintain an adequate stimulus to the patient.

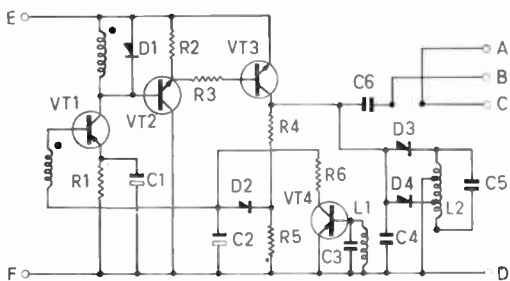


Fig. 4. Circuit diagram of a controllable pacemaker.

R1 500 kΩ	C1 47 μF	D1 OA47	VT1 2S322
R2 10 kΩ	C2 47 μF	D2 OA200	VT2 2S701
R3 1 kΩ	C3 0.002 μF	D3 OA47	VT3 2S711
R4 470 kΩ	C4 0.03 μF	D4 OA47	VT4 2S322
R5 1.2 kΩ	C5 0.03 μF	Transformer:	
R6 8.2 kΩ	C6 22 μF	Core FX 1011 pri. 950 turns	
		sec. 250 turns	
		48 s.w.g.	

An indication that the batteries are beginning to become exhausted is shown by a slowing of the heart rate. The life of the unit is calculated to be several years but is dependent to a large extent upon loading. This is not an immense problem as the change will take place over two or three months. In practice a unit of this type has been in use for six months with no change of rate except under the influence of an external signal and has been turned off for 10 days during that period. The pacemaker is powered by four Mallory batteries type RMIW which are estimated to have a shelf life of five years. The batteries are separate from the electronic circuit in order to

facilitate battery changing when this becomes necessary.

5. Problems of Connecting the Pacemaker

The wires connecting the pacemaker to the heart and the electrodes present a number of problems. In the early cases a wire consisting of a stainless steel spiralled ribbon attached to a platinum electrode was used. Later this electrode assembly was replaced by a braided stainless steel wire surrounded by polythene, the electrode being the uninsulated end. By this method any possible fault caused by the junction of electrode to the wire was eliminated. A silver tinsel wire insulated by polythene was also used but this proved unsatisfactory, and fractures occurred in the other systems.

The present method is a loose triple spiral of stainless steel wire 0.005 inches in diameter. This is insulated with silicon rubber tubing. The wire after going from the pacemaker to the heart (where the insulation is removed), is returned to a second connection in the pacemaker. By this means the wires leading to the electrodes are duplicated. In 29 cases this system has been used without evidence of fatigue or electrolysis.

The wires are sealed into the pacemaker with nylon screws and silicon grease is also inserted into the connectors. In the process of attaching the wires to the pacemaker the silicon grease is forced into the lumen of the rubber tubing and this helps to prevent the entry of body fluids into the pacemaker.

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- Manuscript received by the Institution on 8th May 1962*
(Contribution No. 56/MBE8).

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The Programme and Papers Committee invites the submission of short contributions (less than 2000 words) for consideration for publication. Comment on papers is also welcomed for publication as written discussion, regardless of whether the paper has been read at a meeting.

REPORT OF THE ANNUAL GENERAL MEETING

The Institution's first Annual General Meeting since incorporation by Royal Charter was held at the London School of Hygiene and Tropical Medicine on Wednesday, 24th October 1962.

The Chair was taken by Mr. W. E. Miller, M.A. (an Executive Vice-President), and when the meeting opened at 6.5 p.m. 64 Corporate Members had signed the Minute Book.

The Secretary, Mr. Graham D. Clifford, read the notice convening the meeting as circulated to members and published in the August 1962 *Journal*. The meeting then proceeded to the Agenda as follows:

1. To note the Minutes of the 36th Annual General Meeting of the Incorporated Institution held on 24th January 1962.

The Secretary stated that the Minutes of the last Annual General Meeting held by the Incorporated Institution were published on pages 94–96 in the February 1962 *Journal*. The Chairman's proposal that those Minutes be signed as a correct record of the proceedings was approved unanimously.

2. To receive the Minutes of the first Special General Meeting.

The Minutes of the first Special General Meeting of Corporate Members of the Chartered Institution which were published on page 116 of the August 1962 *Journal*, were signed as a correct record of the proceedings.

3. To receive the Annual Report of the Council for the year ended 31st March 1962.

The Chairman first reminded members that although reference was made to the event at the last Annual General Meeting, the actual grant of a Royal Charter to the Institution was, in fact, only made in the year under review. Mr. Miller continued:

"This is an honour for which many of us have worked for over 30 years, but the grant of the Charter itself has not by any means finished the work of the Council and its Committees. We have had to submit to Her Majesty's Privy Council draft Bye-laws; this has involved us in a great deal of internal work and consultation with our Solicitors and Counsel. It therefore gives me particular pleasure to tell you that this work has resulted in the Privy Council accepting the draft Bye-laws with the minor amendments which were agreed subsequent to the Special General Meeting, and that we are now, from this very month, working in accordance with our Charter and the established Bye-laws.

"We have been grateful to the Privy Council for permitting us to enact separately conditions relating

to the formation, operation and administration of Local Sections in Great Britain and Divisions overseas. Members will be aware from the Annual Report† of the work that is going on in Canada as a result of the visit which our Secretary made to Canada and the United States of America at the end of last year.

"In addition, we still have tremendous work to do and decisions to make in regard to the future of the Institution in India and other parts of the Commonwealth. Your Council is also concerned with what steps should be taken to create Divisions of the Institution outside the British Commonwealth. You will not expect me to comment on this development other than to advise you that the Executive Committee in particular is studying the effect on the Institution's activities of British participation in the Common Market.

"All this is a background to the future development of our Institution. Immediately we have also been concerned to obtain here in Great Britain a measure of agreement with our sister Engineering Institutions. It gives me particular pleasure to refer to the press notice which has only been issued in the last few days on the formal setting up of an Engineering Institutions Joint Council,‡ of which our Institution is a founder member. Older members may recall that in Presidential Addresses and at Annual General Meetings we have very frequently discussed the desirability of the Engineering Institutions getting together in order to present united opinion on matters which affect the status and profession of qualified engineers. We believe that the Engineering Institutions Joint Council is the best possible medium by which to achieve this collaboration and I am sure that the news of the formal formation of this Council and of our participation as a founder member will be welcomed by every member of the Institution.

"This development also enables us to have closer relationship with other Engineering Institutions who have at some time or another an interest in the development of radio and electronics. I refer to the development of control systems which have resulted in the setting up of the British Conference for Automation and Computation. Details of the achievements of the Joint Committee of the British Institution of Radio Engineers and the Institution of Electrical Engineers are also dealt with in the Annual Report.

† *J. Brit. I.R.E.*, 24, No. 3, pp. 181–94, September 1962.

‡ *J. Brit. I.R.E.*, 24, No. 5, pp. 341, 342, November 1962.

"It is always difficult for a Chairman of an Annual General Meeting to decide how much of a circulated Annual Report needs further comment. I am certain, however, that you will want me to express on your behalf thanks to all the Standing Committees whose work has been mentioned in the Annual Report, and whose joint efforts are so responsible for the status which the Institution enjoys.

"Without therefore going into detail of all the work of our Standing Committees, I would like to confine my last remarks to the tremendous step which we have taken in these last few months in acquiring the adjacent building to No. 9 Bedford Square—No. 8. Those members who have visited No. 9 Bedford Square in recent months will know that the extension of Institution activities is reflected in the overcrowded state of the entire floor space of No. 9 Bedford Square. The acquisition of No. 8 will, the Council is confident, enable us to extend further the services which we wish to give to members and provide facilities, in due time, for the provision of at least some of our members' meetings to be held in our own buildings."

Mr. Miller said that he felt that the Annual Report now before the meeting was the most outstanding ever produced by the Institution. His proposal that it should be adopted, which was seconded by Mr. H. H. Harper (Associate Member), Chairman of the South-Western Section of the Institution, was approved unanimously.

4. To receive the Auditors' Report, Accounts and Balance Sheet for the year ended 31st March 1962.

The Chairman called upon the Honorary Treasurer, Mr. G. A. Taylor (Member), to deal with this item.

Mr. Taylor explained that the Income and Expenditure Account and Balance Sheet for the year ended 31st March 1962 were in a slightly different form to those which have previously been presented, and stated that:

"The Finance Committee had to anticipate that the costs involved for the year under review might result in a heavy deficiency of income as against expenditure. In fact, the loss has been very much less than anticipated, and this is not only due to some economies imposed by the Finance Committee but of course to the increase in revenue as a result of the growth in the Institution membership.

"I would be failing in my duty if I did not remind members that the year under review does not take into full account a year of increased charges in such items as printing, and a yet further increase is expected at the end of this year on such items as postage and telephones.

"We must also note that during the next year we can—and I say this almost thankfully—expect a rise

in expenditure because we now have the space to accommodate the staff which is necessary to cope, not only with increased work because of increased membership but because of the increasing services that we want to give to members of the Institution.

"In the presentation of the Balance Sheet we are now able to show members the exact amount of monies being expended on building. The Finance Committee are also proposing that we shall in future show some measure of the extent of Institution expenditure in connection with its division activities overseas. This I think would be a very important point in showing how the Institution is growing outside the U.K. and also what the Council of the Institution is doing to permit that growth in terms of expenditure.

"I have had tremendous pleasure, in the last twelve months of being associated with our Finance Committee in their endeavours to meet our expenditure, heavy though it has been, out of revenue. The measure of our success is reflected in the fact that we have not found it necessary to sell any of our investments, but that we have in fact been able to add to them.

"We are now approaching a reserve figure which will partly cushion us against increasing costs.

"In all the eight years that I have now been the Treasurer of the Institution, I have never been able to present such satisfactory accounts as are now before the membership and I have great pleasure in formally moving the adoption of the Accounts for the year ended 31st March 1962."

The proposal was seconded by Mr. F. G. Diver, M.B.E. (Member) and approved unanimously.

5. To confirm the election of the Council for 1962-3.

The Chairman said: "I am sure that all members of the Institution are pleased and honoured that Admiral of the Fleet the Earl Mountbatten of Burma has agreed to continue as President of the Institution for a further year. The President had hoped to preside over this meeting tonight but you will all appreciate that because of recent international tension his other more important duties prevent his attendance. He has, however, asked me to convey his thanks to all members for electing him as President for another year and to say that he hopes to meet as many members as possible when he attends the 1963 Convention which is taking place at the University of Southampton."

Mr. Miller said that members would be aware from the Agenda for the meeting that it had not been necessary to hold a ballot to elect the Officers and Council for the coming year and he had pleasure in formally confirming that the members named in the Agenda will comprise the 1962-3 Council. He then extended a particular welcome to those members who

would be serving on the Council for the first time and he added special congratulations on behalf of those present to Mr. A. A. Dyson on his election as a Vice-President.

6. To appoint Auditors and to determine their remuneration.

In recommending the appointment of Gladstone, Jenkins & Co. as Auditors to the Chartered Institution Mr. Miller said that the Council would like to take this opportunity to record its appreciation of the help given by Mr. R. H. Jenkins in the preparation of the new form of Accounts and Balance Sheet. He was sure that the recommendation that Gladstone, Jenkins & Co. should be appointed Auditors to the Institution would be unanimously approved. The Chairman added that it was customary for members to endorse also that the Auditors' remuneration should be at the discretion of the Council. The motion was approved unanimously.

7. To appoint Solicitors.

Dealing with this item, the Chairman said: "The Annual Report of the Council has paid tribute to the work done by Mr. C. Gray Hill in assisting with all the legal documentation in connection with the Institution's Charter. Since then he has given tremendous help in connection with the drafting of our Bye-laws and the fact that these have been approved by the Privy Council speaks highly of his efforts. This is in

addition to other legal matters which have required consultation and advice. I am sure that you will all wish to record a vote of thanks to Mr. Hill and to show your approval of Council's recommendation that Messrs. Braund & Hill be appointed Solicitors to the Institution."

The proposal was approved unanimously.

8. Awards to Premium and Prize Winners.

Mr. Miller said that it gave him particular pleasure to move to the next item on the Agenda, that of the presentation of Premiums and Awards to authors of outstanding papers published in the Institution's *Journal* during the year 1961 and of Prizes to outstanding candidates in the Graduateship Examination held during the year.

He then presented the Premiums, Awards and Prizes in accordance with the lists published in the September issue of the *Journal* (pages 186-7 and 190).

9. Any other business.

The Secretary stated that notice of any other business had not been received and the Chairman then declared the first Annual General Meeting of the Chartered Institution to be at an end.

The Annual General Meeting of the Institution was followed by the Annual General Meeting of Subscribers to the Institution's Benevolent Fund, a report of which will appear in the January issue of the *Proceedings of the Brit.I.R.E.*

APPLICANTS FOR ELECTION AND TRANSFER

The Membership Committee at its meeting on 29th November 1962 recommended to the Council the election and transfer of 32 candidates to Corporate Membership of the Institution and the election and transfer of 117 candidates to Graduateship and Associateship. In accordance with Bye-Law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

CORPORATE MEMBERS

Direct Election to Associate Member

BENNETT, John Bevan. *Wokingham, Berkshire.*
BILLINGS, Kenneth. *Mansfield, Nottinghamshire.*
BROOKS, Michael. *Gravesend, Kent.*
DAVIES, David Evan, Ph.D., M.Sc., B.Sc. *Birmingham.*
DELL, Frederick Reginald, B.Sc.(Eng.). *Harrow, Middlesex.*
DICK, Alexander Morrison, B.Sc.(Eng.). *Ty-Croes, Anglesey.*
GORDON, Frederick Begg. *Watford, Hertfordshire.*
HANNAFORD, Douglas, B.Sc. *Upminster, Essex.*
HOLT, Roy Kenyon. *Uddington, Lanarkshire.*
JONES, Wilfred James. *Alsager, Cheshire.*
MATTHEW, James Donald. *Barnet, Hertfordshire.*
RATTLE, Peter Henry. *Waltham St. Lawrence, Berkshire.*
TOBEY, Richard, B.A.(Eng.). *Redhill, Surrey.*
WHIPP, Edwin, B.Sc. *Birmingham.*

Transfer from Associate to Associate Member

COOPER, Captain Bertram Norman, R.E.M.E. *Salisbury, Wiltshire.*
FARRELL, Roy Colin. *Bishop's Stortford, Hertfordshire.*

MALSTER, John. *Welwyn Garden City, Hertfordshire.*
MITCHELL, Victor Leslie. *Stoke-on-Trent, Staffordshire.*
STRIDE, Regent. *Morden, Surrey.*
WHITHAM, Eric. *Bracknell, Berkshire.*

Transfer from Graduate to Associate Member

ASHEN, David John. *London, E.4.*
ATKINSON, Peter, B.Sc. *Letchworth, Hertfordshire.*
AVISS, Hugh Wilson. *Tunbridge Wells, Kent.*
BROWN, John Millard. *Leigh-on-Sea, Essex.*
CHRISTMAS, Kenneth James. *Colchester, Essex.*
JENNINGS, David Michael., B.Sc.(Eng.). *St. Annes-on-Sea, Lancashire.*
McCAW, John. *East Kilbride, Lanarkshire.*
PALMER, Reginald Charles. *Cardiff.*
REID, Cynthia Margaret, M.A. *London, S.W.6.*
SAUNDERS, Michael William. *Sutton, Surrey.*
SOLMAN, Stanley Jack. *West Croydon, Surrey.*
TODD, Robert John. *Feltham, Middlesex.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

ABBOTT, Edwin James Michael. *London, N.W.2.*
ADAMS, Raymond. *Leicester.*
ARORA, Chandra Kishore, B.Sc. *London, S.W.4.*
AWOSIKA, Festus Ayotunde. *London, S.W.4.*
BALDWIN, David Michael. *Bourne End, Buckinghamshire.*
BARRATT, Lionel Stanley. *Southend-on-Sea, Essex.*
BATT, Eric Leon. *Bishop's Stortford, Hertfordshire.*
BEARDSWORTH, William. *Chesham, Buckinghamshire.*
BEVAN, David William. *Hayes, Middlesex.*
BLYTHE, David Colin. *London, N.W.4.*
BONTOFT, Frederick John. *Maldon, Surrey.*
BRANCH, Julian Philip George. *Ruislip, Middlesex.*
BREINGAN, William John Malcolm. *Harrow, Middlesex.*
BRIGGS, Leonard Albert. *Rayleigh, Essex.*
BUCKROYD, Allen. *Thornton Heath, Surrey.*
BURTON, Peter Leonard Arthur. *Purley, Surrey.*
CABLE, Valerie Ann. *Chelmsford, Essex.*
CADWALLADER, Michael Vaughan. *Hatfield, Hertfordshire.*
CARTER, David George. *Maidenhead, Berkshire.*
CARTER, Eric Vance. *Old Coulsdon, Surrey.*
CHADWICK, William Ernest. *Stoke-on-Trent, Staffordshire.*
COOK, David George. *Portsmouth, Hampshire.*
DOWNTON, Ralph. *Woodlands, Hampshire.*
DURLING, Robert Ross. *Stansted, Essex.*
EADY, Evelyn Patricia. *London, E.7.*
EARNEY, Kenneth John. *Danbury, Essex.*
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The names of 37 students registered at this meeting will be published in the January issue of *The Proceedings of the Brit.I.R.E.*

The Fundamental Importance of the Heaviside Operational Calculus

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Summary: It is an essential part of Heaviside's operational calculus that the symbol p is an operator equivalent to d/dt and that p and p^{-1} , the inverse or integrating operator, are commutative. This is ensured if the operation of integration is not confused with the operation of selection, i.e. if the lower limit of integration is taken as minus infinity, and is only raised to a finite value when we are sure that this change has no effect. We have first set forth in as explicit a manner as possible what we believe to be the basis of Heaviside's own work, with particular attention to the properties of the unit function $H(t)$ and the way in which differentiation and integration can be extended to include functions containing $H(t)$ or its derivatives as factors. A continuous function approximating to $H(t)$ is considered in an Appendix. The kinds of function (of time) that can occur in nature are carefully considered; the case of both passive and active networks is discussed. Heaviside's contemporaries were not prepared to accept his premises and methods even if they were forced to accept his results. The relation between Heaviside's calculus and Fourier analysis, symbolic calculus and Laplace transforms is therefore considered; the advocates of symbolic calculus and particularly of Laplace transforms have introduced difficulties and even errors which need not have occurred if they had followed Heaviside more faithfully.

The full power and universality of Heaviside's approach (in which the mathematics was always subordinate to the physics) is made clear in Section 4, where the relation between input and output is considered for any system, not necessarily electrical; this relation is expressed by a single operational equation, but there are several possible ways of handling that equation and it is important not to choose too early which of these ways should be used.‡

1. Introduction

It has been known for a long time that a function $f(t)$ of time (which must be finite for all t and have a finite integral if it is to be capable of representing a physically realizable state of affairs) can be specified not only in terms of its value for every relevant value of t , but also in various other ways. $f(t)$ can, for instance, be completely specified in terms of its moments, or in terms of its "spectrum" or Fourier components, that is, in terms of sines and cosines. Expressions in terms of other orthogonal functions (Legendre, Laguerre, etc.) are also known. $f(t)$ can also be expressed as the result of a set of differential or integral operations on an "operand" $\phi(t)$; this form of expression is more general in scope than those already mentioned, and is known as "operational calculus". Heaviside's pioneer work in this field was done with his "unit function" $H(t)$ as the universal operand, which we called $\phi(t)$ above ($H(t)$ is zero for negative t and unity for positive t). Heaviside's

calculus consists essentially in specification of $f(t)$ by means of a differential equation. There are, however, important and far-reaching generalizations and simplifications in Heaviside's calculus which were not available in the relatively crude operational calculus known before Heaviside's time. The manipulation by algebraic methods of expressions involving the symbol

$$D = d/dx \quad \dots\dots(1)$$

has been used and accepted for many years by orthodox mathematicians, and this constituted a very tentative operational calculus. It was in effect only a convenient notation; results obtained by this notation were only accepted if also obtainable otherwise. Thus the expression

$$e^{aD}f(x) = f(x+a) \quad \dots\dots(2)$$

was a symbolic shorthand statement of Taylor's theorem and no more.

‡ In the case of receivers, this may throw emphasis on what the receiver does rather than on the nature of the input. In effect a receiver merely takes a kind of running average over a few cycles of the input, weighted according to the receiver bandwidth, etc.

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There was no particular attempt to establish a valid and provable calculus. In fact, one of the essential requirements for such a calculus, namely that

$$DD^{-1}f(x) \equiv D^{-1}Df(x) \equiv f(x) \quad \dots\dots(3)$$

was not secured by adequate and precise specification of D and D^{-1} .

Perhaps the greatest advance in extending the usefulness of the concept of "operator" was made by Heaviside¹ when he introduced his "unit-step" and "unit-impulse", and adjusted the meaning of "differentiation" (represented by p) and "integration" (represented by p^{-1}) so that they applied to discontinuities of this kind.† Unfortunately, however, this brilliant advance met with disapproval from most contemporary mathematicians. Heaviside manipulated his operator p in such a way that for any function $h(t)$ of time which he regarded as useful

$$p \cdot p^{-1}h(t) = p^{-1} \cdot ph(t) = h(t) \quad \dots\dots(4)$$

After Heaviside's death, the idea that p was an operator became unpopular and went out of fashion—van der Pol² and others developed "symbolic calculus" in which $f(t)$ "corresponded" to an expression $\phi(p)$ formally the same as Heaviside's alternative specification, but p was now regarded as a pure algebraic number; $f(t)$ and $\phi(p)$ were in different "worlds",‡ and there were certain rules to be followed for journeys between the two "worlds" in either direction.§ Nevertheless at this period Jeffreys^{12,13} was pursuing and extending the treatment of operational methods, and examining the commutativity of the operators of differentiation and of definite integration. The authors believe that deviation from Heaviside's idea, that his calculus is *operational* ($p \equiv d/dt$), in the direction of symbolic calculus or the equivalent Laplace transforms, leads to ambiguity and error. They hope to show that there are no real difficulties in operational calculus, and that the difficulties usually met are due to faulty fundamental initial concepts and misleading and ill-understood notation. The discussion is on an elementary yet fundamental level commensurate with the errors associated with symbolic calculus and Laplace transforms—errors made by those who sought in vain to improve upon Heaviside.

In the main text of this paper, we have tended to take a viewpoint analogous to that of Heaviside, and mathematical and other difficulties have been circumvented mainly by means of physical reasoning.

† We here adopt the symbol p , the "time differentiator" used by Heaviside in his earlier work.

‡ Namely, K. W. Wagner's "Unter- und Oberbereich".³

§ Laplace transforms are in effect another form of "symbolic calculus" in which $f(t)$ is represented by $\phi(s)/s$ in the notation used above, and again s is regarded as a pure number.

This treatment will hardly be adequate for mathematicians, but in order not to confuse students and engineers who are not mathematicians, we have relegated the detailed discussion of points of difficulty to footnotes and appendices.

2. Heaviside's Calculus

In this section, the authors wish to be considered as counsel for Heaviside, endeavouring to formulate explicitly the assumptions and techniques he used. We believe that his principles were always correct, though not by any means fully explained by him. In the following, we shall where desirable make full use of terms which have been defined and formulated since the death of Heaviside. The efficacy of Heaviside's results has always been admitted. We shall try to show the fundamental soundness of his concepts also.

The fundamental concept is that the symbol p is an operator equivalent to d/dt ; this operator operates upon an operand $h(t)$ which is a function of time. Thus the equation

$$\phi(p)h(t) = h^*(t) \quad \dots\dots(5)$$

means that the result of performing the series of differentiations, integrations and manipulations represented by $\phi(p)$ upon $h(t)$ is to arrive at the function $h^*(t)$. As already stated, p is equivalent to d/dt ; p^r is likewise equivalent to d^r/dt^r . We have to define p^{-1} , the operator inverse to p , with care; if we can formulate the definition in such a way that equation (4) is true, then the operations specified by $\phi(p)$ will yield the same result whatever the order in which differentiations and integrations are performed. Any algebraic manipulation of $\phi(p)$, such as splitting it up into partial fractions, is therefore permissible. It is taken for granted that we do not assume

|| One of the most confusing difficulties associated with the formulation of an operational calculus is the loss of a constant after differentiation. In the main text, we have resolved this difficulty by saying that a constant having zero derivative for all time is a non-entity. The following slightly different but ultimately equivalent approach is due to G. G. Gouriet. If a current is passed through a "pure inductor" (whose action is represented by the operator Lp) the d.c. component of the current is immaterial. In the world, however, the action of an actual inductor is better represented by the operator $L(p+\epsilon)$ (ϵ small and positive) which does not lose the d.c. component. We shall see later that the reciprocal operator $1/L(p+\epsilon)$ is perfectly clear and understandable. Normally, we should regard p as

$$\lim_{\epsilon \rightarrow 0} (p+\epsilon);$$

in certain very special cases p should be regarded as

$$\lim_{\epsilon \rightarrow 0} (p-\epsilon).$$

If this is always done, no anomalies occur. For simplicity we normally omit, but should never forget, the ϵ . Mathematical errors can easily be made here and elsewhere by prematurely proceeding to the limit when ϵ tends to zero.

$h_1(t)g(p)h_2(t)$ to be the same as $h_2(t)g(p)h_1(t)$.† We are in fact dealing with a non-commutative algebra. Whereas the “orthodox” mathematicians already mentioned were prepared to use only continuous and differentiable functions as operands, Heaviside allowed his operand to have a finite number of “step”, “impulse” or “higher-order impulse” discontinuities, and he extended the meaning of differentiation and integration to include such discontinuities. All such discontinuities can be expressed in terms of Heaviside’s “unit function” or “unit step”, $H(t)$, and its derivatives. $H(t)$ is defined by the equations

$$\begin{aligned} H(t) &= 0 & t < 0 \\ H(t) &= 1 & t > 0 \end{aligned} \quad \dots\dots(6)$$

Heaviside found the definition (6) sufficient, and never allowed himself to be involved in a situation where the value of $H(0)$ mattered. The orthodox mathematicians were unwilling to accept these concepts. They did not fully appreciate the ultimate significance of the method of limits although they themselves used this method in (for example) the differential calculus by writing dy/dx as the limit of $\delta y/\delta x$ when δx tends to zero. There were, and still are, apparent anomalies in orthodox mathematics due to misuse of limits. But Heaviside knew intuitively that the difficulties were introduced by the mathematicians, not inherent in the subject.

As explained later, in all cases where the limit is closed (so that, in the differential calculus, for example, we refer to dy/dx and not $\delta y/\delta x$) the result is only a convenient “nominal” value. In fact, dy/dx means $\delta y/\delta x$ with δx as small as we please but never zero. $H(t)$ and $\delta(t)$ are likewise “nominal” or “conceptual” expressions, used for convenience of notation, to express limiting forms which approach the “ideals” as nearly as we wish, however near, but never actually reach the “ideal” or “conceptual” form specified. Dirac in his specification of $\delta(x)$ pointed out that the nominal or ideal form can nearly always be used without error to obtain a given result with less labour than is needed if such a peculiar function is forbidden. But in all cases of difficulty or doubt (which usually occur when some sort of critical comparison is involved) the nominal or conceptual must be replaced by the actual. Metrical or actual entities which approach arbitrarily closely the nominal ideas of unit step, unit impulse, pure differentiation and pure integration are discussed in Appendix 1.

It is clear from equations (6) that if we are to extend the meaning of differentiation to include the

† This apparent limitation is actually an improved notation. In the English language (but not in some other languages) order of words is significant; this is a great advantage. Thus “dog bites man” is not the same as “man bites dog”. The operator p , like an English verb, operates on what is written after it.

function $H(t)$ (which has an isolated discontinuity when $t = 0$), the derivative must be zero for all values of t other than zero. Further, the derivative must be infinite when $t = 0$, for otherwise the integral of this derivative, that is, the difference between $H(t)$ and $H(t')$, would necessarily be zero for all values of t and t' . In fact, from equations (6), this difference is zero if t and t' have the same sign, and unity if t and t' have opposite signs ($t < t'$); since $H(0)$ is not defined, neither t nor t' may actually be zero. Thus we arrive at the fact that the derivative $\delta(t)$ of $H(t)$ is zero for all non-zero values of t , infinite when $t = 0$, and satisfies the equation

$$\int_{-\infty}^{\infty} \delta(t) dt = 1 \quad \dots\dots(7)$$

in which the upper limit may be replaced by any positive number ϵ_1 however small, and the lower limit by a negative number $-\epsilon_2$ however small. Thus if we define $\delta(t)$ as the derivative of $H(t)$ we in fact find that it is the same as Dirac’s delta function, and we have therefore used Dirac’s notation for it.‡ $\delta(t)$ is usually regarded as an even function of t ; for the purpose of this section, however, this need not be assumed; it only matters if we attempt to make one of the limits of the integral (7) actually equal to zero.

The function $\delta(t)$ is usually referred to as “unit impulse”; $\delta(t)$, like $H(t)$, can be differentiated if we are prepared to extend the meaning of differentiation to include a function $\delta'(t)$ having very violent behaviour when $t = 0$ but zero for all other values of t .§ In fact, we can form any number of successive derivatives of $\delta(t)$, which will all be zero for non-zero values of t , and will all behave violently when $t = 0$. We are not interested in the precise nature of this violent behaviour; we carry out differentiations of the form $f(t)H(t)$ by using the rule for differentiating a product (or its extension, Leibnitz’ theorem) just as for continuous functions. Inserting the derivatives of $H(t)$ in the places where they naturally occur warns us that, when $t = 0$ and at no other time, violent behaviour of these derivatives is to be expected. This is hardly surprising from the physical point of view when a process is suddenly started at $t = 0$; such a start is the basis of Heaviside’s thinking.

One other property of the impulse function which we shall require later is its “spotting” property, namely that

$$\int_{-\infty}^{\infty} h(\lambda)\delta(t-\lambda) d\lambda = h(t) \quad \dots\dots(8)$$

‡ We do this because Heaviside’s notation “ u ” does not contain the argument t . Further $\delta(t)$ is extensively used in the literature. The δ of course has different meanings in the expressions $\delta(t)$ and $\delta y/\delta x$.

§ A continuous and differentiable function approaching $H(t)$ in the limit is discussed in Appendix 1.

Here the limits can be made from $t - \epsilon_2$ to $t + \epsilon_1$ just as the limits in equation (7) could be replaced by $-\epsilon_2$ to ϵ_1 . The reason for equation (8) is that $h(\lambda)$ can be replaced by $h(t)$ without altering the integral, since $\delta(t - \lambda)$ is zero for all λ except t . When $h(\lambda)$ is replaced by $h(t)$, it can be taken outside the integral, and equation (7) applied.

We now have to consider the inverse operator p^{-1} . As this operator is inverse to differentiation, it is clearly an integration of some kind, and in defining it more precisely we have to bear in mind the importance of defining it so that equation (4) applies to all functions $h(t)$ regarded as useful and relevant. Before the definition can be made intelligible, we must consider the general nature of the functions associated with physical reality. It is also necessary to distinguish the process of "integration" from that of "selection".

We regard all functions that occur in nature as continuous and differentiable any number of times except at a finite number of isolated points where they may include terms consisting of the product of a continuous function differentiable any number of times, and a step function $H(t - t_0)$ or one of its derivatives. We also regard all such functions as integrable any number of times; the impulse or step-function type of discontinuity presents no difficulty from the point of view of integration. If $h(t)$ is a function occurring in nature, then the integral

$$\int_{-\infty}^{\infty} |h(t)| dt \quad \dots\dots(9)$$

will be finite. We shall only use the term "integration" for integrating from $-\infty$ to $+\infty$, that is to say, integrating all of the function $h(t)$ that there is.

In some cases, the lower limit $-\infty$ can be replaced by a finite lower limit without affecting the value of the integral, and it may also be possible to reduce the upper limit to a finite amount similarly. Such a reduction of limits is permissible at our risk and peril; if it has any effect on the value of the integral, the operation we have performed is not merely an integration, but an integration combined with a selection. There is one minor exception to this restriction; if we require the function $f(t)$ such that

$$pf(t) = h(t) \quad \dots\dots(10)$$

it is
$$f(t) = p^{-1}h(t) = \int_{-\infty}^t h(\tau) d\tau \quad \dots\dots(11)$$

that is to say, a finite *variable* upper limit is permissible when we are thinking of integration as the process inverse to differentiation, rather than as a summation or the evaluation of an area. Selection is then indicated also, but it is the selection of everything up to a stated variable time t . It is still possible to separate the operation of selection from that of

integration; thus the expression normally written as

$$\int_a^b h(t) dt \quad \dots\dots(12)$$

can be rewritten in the form

$$\int_{-\infty}^{\infty} h(t)[H(t-a) - H(t-b)] dt \quad \dots\dots(13)$$

in which the limits are appropriate for the process we have described as integration, while the process of selection is explicitly written into the integrand.

Any function which occurs in nature will have finite moments, that is to say

$$I_n(t) = \int_{-A}^B t^n h(t) dt \quad \text{is finite} \quad \dots\dots(14)$$

for all finite values of n , as A and B tend independently to infinity. Thus $I_n(t)$ must not be finite merely by virtue of a cancellation. If $n = 0$, the expression (14) is related to the expression (9), and in order that the integrals (14) may be finite, the integrands must tend to zero as t tends to infinity with sufficient rapidity; it must be possible to find numbers k_1 and k_2 such that

$$\int_{-k_1}^{k_2} t^n h(t) dt \quad \dots\dots(15)$$

differs from the expression (14) by as little as we please. As replacing the integral (14) by the integral (15) makes no effective difference, we escape the condemnation we should otherwise deserve for having performed a selection as well as an integration in expression (15), which is written only to explain the fact that the function $h(t)$ is of effectively finite duration from $-k_1$ to $+k_2$.

Since $I_1(t)$ and $I_0(t)$ are both finite for the functions we regard as relevant to physics, so also is their quotient $I_1(t)/I_0(t)$. If the variable t were replaced by a space-coordinate x , and we were given a graph of $h(x)$ (measured in the y -direction) as a function of x , $I_0(x)$ would be the total area under the curve, and $I_1(x)/I_0(x)$ would be the x -coordinate of the centroid of that area; expression (15) would now indicate that $h(x)$ had an effectively finite spread from $-k_1$ to $+k_2$. In this paper, we shall call the ratio $I_1(t)/I_0(t)$ the "epoch" of the time-function $h(t)$; it is in effect the mean time at which $h(t)$ occurs. If $h(t) = \delta(t - \lambda)$, the epoch is λ , the time at which the impulse-function is different from zero. All functions of time which here concern us have finite epoch and duration; in order to specify epoch and duration we require both a time-origin and a measure of time before or after that origin.

We are now in a position to see that defining the operator p^{-1} by means of equation (11) gives us a completely satisfactory operational calculus, in which

equation (4) is true for all useful functions $h(t)$ which have finite duration, definite epoch and finite moments as discussed above. For

$$p \cdot [p^{-1}h(t)] = \frac{d}{dt} \int_{-\infty}^t h(\tau) d\tau = h(t) \dots\dots(16)$$

$$p^{-1}[p h(t)] = \int_{-\infty}^t \frac{dh(\tau)}{d\tau} d\tau = h(t) - h(-\infty) = h(t) \dots\dots(17)$$

$h(-\infty)$ being zero because $h(t)$ is, as we have seen, of finite duration. Equation (4) means that we can have complete confidence that all processes involving differentiation and integration and summation of the results of repeated differentiations and integrations are always reversible and may therefore be performed in any convenient order. Provided that only operands $h(t)$ satisfying equation (14) are used, and that we always regard the operator p^{-1} as defined by equation (11), and we only alter the lower limit when we are absolutely certain that doing so will have no effect, there is complete certainty that any solution of equation (5) obtained by any manipulation of $\phi(p)$, or by regarding $h(t)$ as the result of performing some other operation $\psi(p)$ on a different function of time such as $H(t)$ and manipulating the product $\phi(p)\psi(p)$, is valid.

One consequence of equation (14), however, is that many simple functions we have in the past taken for granted become inadmissible, that is to say meaningless, such as

$$a, e^{at}, e^{-at}, \sin \omega t, aH(t) \dots\dots(18)$$

a, α and ω being real, positive and constant.

On the other hand

$$e^{-at}H(t), e^{at}H(-t), e^{-at}H(t) \sin \omega t, a[H(t) - H(t-t_0)] \dots\dots(19)$$

are admissible; in the case of the last expression, t_0 may be as large as we please provided that it is finite.

If now $h(t)$ and $f(p)$ are such that

$$h(t) = f(p)H(t) \dots\dots(20)$$

then $f(p)H(t)$ can be regarded as an alternative means of specifying the function $h(t)$. The simplest case of this is

$$\delta(t) = \frac{d}{dt} H(t) = pH(t) \dots\dots(21)$$

so that $pH(t)$ is the alternative means of specifying $\delta(t)$.†

† Strictly speaking $aH(t)$ is not an admissible function in equation (21) with $a = 1$, as was foreshadowed in the list (18). But the eternal constant a is merely another nominal or conceptual idea (this time of a theological nature) so that $aH(t)$ must be replaced by a metrical expression like $a \exp(-\epsilon t)H(t)$ when difficulty or doubt arises.

Equation (20) is in Heaviside's operational language; it means that $h(t)$ is the result of applying the operator $f(p)$ to the operand $H(t)$, and the sign of equality is necessary, correct and sufficient.

In the notation of symbolic calculus or Laplace transforms, however, it is necessary to find some way of expressing directly the relation between $f(p)$ or $f(p)/p$ on the one hand and $h(t)$ on the other. This is done by saying that $f(p)$ is the "image" or "operational equivalent" of the "original" $h(t)$, or that $f(p)/p$ is the "Laplace transform" of $h(t)$, and by writing

$$f(p) \doteq h(t), f(p) = h(t), f(p)/p = L[h(t)], \text{ etc.}$$

In operational calculus, however, the sign of equality in equation (20) is all we require, and the confusing terms image, operational equivalent, transform, etc., are irrelevant and unnecessary.

Now consider the operator $e^{p\tau}$. We know already that any algebraic manipulation of this operator is permissible, so that we can write

$$e^{p\tau} \phi(t) = [1 + p\tau + (p^2\tau^2/2!) + (p^3\tau^3/3!) + \dots] \phi(t) = \phi(t) + \tau\phi'(t) + (\tau^2/2!)\phi''(t) + \dots \dots\dots(22a)$$

$$= \phi(t + \tau) \dots\dots(22b)$$

Equation (22a) is valid provided that the relevant derivatives of $\phi(t)$ exist. If necessary, we can regard a function like $H(t)$, for which these derivatives are somewhat abnormal functions, as the limit of a continuous function (such as the expression discussed in Appendix 1) which does possess derivatives, and we therefore regard equation (22b), which is meaningful whatever the nature of $\phi(t)$, as universally applicable. If τ is negative, the operator may be regarded as having a possible physical representation in the form of a delay cable. If τ is positive, however, there is no such corresponding physical representation. In equation (22b) taken by itself, we have no means of distinguishing the positive time direction from the negative. That distinction arises from considerations of noise and entropy. In a noiseless world, equation (22b) is perfectly valid physically and mathematically whatever the sign of τ and whatever the nature of $\phi(t)$. If in equation (22b), $\phi(t)$ is replaced by $\delta(t)$, and we then substitute for $\delta(t)$ from equation (21), we have

$$e^{-p\tau} pH(t) = \delta(t - \tau) \dots\dots(23)$$

Now reconsider equation (8) (with λ replaced by τ) and substitute for $h(t)$ on the right-hand side from equation (20) and for $\delta(t - \tau)$ in the integrand on the left-hand side from equation (23).

We obtain

$$h(t) = f(p)H(t) = [p \int_{-\infty}^{\infty} e^{-p\tau} h(\tau) d\tau]H(t) \dots\dots(24)$$

so that
$$f(p) = p \int_{-\infty}^{\infty} e^{-p\tau} h(\tau) d\tau \quad \dots\dots(25)$$

The essential simplicity and directness of Heaviside's calculus is revealed in equation (24), and it is worth while to try and express the significance of this equation in words. Since τ is a mere variable of integration, the initial p on the right-hand side of equation (24) may be moved next the operand $H(t)$ and $pH(t)$ is $\delta(t)$, the unit impulse at time zero. $e^{-p\tau}\delta(t)$ is $\delta(t-\tau)$, unit impulse at time τ ; this is multiplied by $h(\tau)$, the appropriate ordinate at time τ , and integrated to build up $h(t)$ in accordance with equation (8). The same result would have been obtained if the operand had been $-H(-t)$, since $p[-H(-t)]$ is also $\delta(t)$. The difference between these two operands is merely conventional if we consider equation (24) by itself. The distinction between forward-looking time (which is appropriate to normal physical problems such as those associated with passive networks) and backward-looking time (which is appropriate to the consideration of active networks, unstable amplifiers, atom bombs, etc.) is only forced upon us by considerations of noise, entropy, the precedence of cause over effect, and so on.

If $h(t)$ is an "admissible" function, equation (25) is perfectly clear, and it is not necessary to make any illegal hypothesis that p is a purely imaginary number. This will be made clear by an example, in which

$$h_1(t) = e^{-at} H(t) \quad a > 0 \quad \dots\dots(26)$$

The alternative specification $f_1(p)$ is then given from equation (24) by

$$\begin{aligned} h_1(t) = f_1(p)H(t) &= [p \int_{-\infty}^{\infty} e^{-p\tau} \cdot e^{-a\tau} H(\tau) d\tau]H(t) \\ &= \left[\lim_{t_1 \rightarrow \infty} p \int_0^{t_1} e^{-(p+a)\tau} d\tau \right] H(t) \\ &= \lim_{t_1 \rightarrow \infty} \left[\frac{p}{p+a} - \frac{p}{p+a} e^{-at_1} \cdot e^{-pt_1} \right] H(t) \\ &= \frac{p}{p+a} H(t) - \lim_{t_1 \rightarrow \infty} \frac{p}{p+a} e^{-at_1} H(t-t_1) \dots\dots(27) \end{aligned}$$

Since a is positive, the last term in equation (27) is zero, so that

$$f_1(p) = \frac{p}{p+a} \quad \dots\dots(28)$$

The same kind of argument can be applied in the case of other admissible functions, but the mathematics may be more laborious. If $h(t)$ starts suddenly at $t = 0$, so that

$$h(t) = v(t)H(t) \quad \dots\dots(29)$$

the lower limit in equation (25) can if we please be

† Or more rigorously by $0-\epsilon$ ($\epsilon > 0$) or $(0-)$.

replaced by zero;† we thus obtain what is usually known as the Carson integral; the "operational equivalent" of $v(t)H(t)$ is $\phi(p)$ where

$$\phi(p) = p \int_0^{\infty} e^{-p\tau} v(\tau) d\tau \quad \dots\dots(30)$$

but it is important to note that the operand in the integral (30) is really $v(\tau)H(\tau)$ and not simply $v(\tau)$; in all cases of doubt, equation (25) should be used.

Equation (25) gives us $f(p)$ when we know $h(t)$, that is to say, equation (25) gives us the operator which must operate upon the operand $H(t)$ to give $h(t)$. We shall consider later how to derive $h(t)$ when $f(p)$ is given. Now Heaviside usually liked to express everything in terms of an operator $f(p)$ operating upon the particular operand $H(t)$. This is obviously convenient in dealing with any situation in which a step-voltage is applied to a network when $t = 0$. It is also convenient when any other voltage is suddenly applied, provided that we know $f(p)$, so that the voltage $v(t)H(t)$ can equally well be regarded as $f(p)H(t)$, the result of applying the operator $f(p)$ to $H(t)$. For then, if the network has admittance $y(p)$, the current $i(t)$ in it will be the result of applying the operator $y(p)f(p)$ to $H(t)$. Normally, it is simplest to obtain $i(t)$ direct from $y(p)f(p)$ as indicated below. This is directly analogous to the procedure of symbolic or Laplace-transform calculus, although it avoids errors discussed below. But operational calculus has the advantage that we need not regard $H(t)$ as the universal operand, if it is to our advantage not to do so. It is to our advantage not to do so, for example, in the case of a frequency-modulated input voltage discussed elsewhere,⁴ because then $f(p)$ cannot conveniently be explicitly formulated. It may also be more convenient to express the input voltage as a series (for example, a power series, or a series of Laguerre functions⁵), to each term of which the admittance operator $y(p)$ can conveniently be applied. It should also be noted that even if we do regard $H(t)$ as the universal operand, it does not follow that $f(p)H(t)$ is a function of time which contains the step-factor $H(t)$; the operator $f(p)$, for example, may include $\exp(\pm p\tau)$ as a factor, and this would shift the step. Sometimes $f(p)H(t)$ does not involve a step response.

Sometimes the real problem is not to determine $i(t)$ given the input voltage, say $v(t)H(t)$ as above, but rather to arrange that $i(t)$ shall have a prescribed form, as in the case of servo-mechanisms. There is no reason why additional networks should not be provided to achieve this end if necessary. This type of problem has been discussed by Gouriet⁶ and his method of attack requires essentially that p be regarded as an operator. Gouriet's main point is that if $y(p)$ is the reciprocal of a polynomial of degree n in p , say of

$(a_0 + a_1 p + a_2 p^2 + \dots + a_n p^n)$, then instead of having to determine the response $I(t)$ to the excitation $r(t)H(t)$, we provide additional networks so that the response is no longer $I(t)$ but

$$\left(a_0 + a_1 \frac{d}{dt} + a_2 \frac{d^2}{dt^2} + \dots + a_n \frac{d^n}{dt^n} \right) I(t)$$

and this is equal to the input voltage $r(t)H(t)$ which is under our control. In practice noise will limit the accuracy to which the additional networks will make the modified response equal to $r(t)H(t)$.

We now therefore have to consider how to find $I(t)$ knowing $y(p)f(p)$ which we shall abbreviate to $\eta(p)$. Heaviside had four general methods of doing this. The most obvious of these is to expand $\eta(p)$ in the form of a series of descending powers of p . From equation (11) with $h(t)$ replaced by $H(t)$, we have

$$\frac{1}{p} H(t) = \int_{-\infty}^t H(\tau) d\tau = t H(t) \quad \dots\dots(31)$$

$$\frac{1}{p^2} H(t) = \frac{1}{p} [t H(t)] = \int_{-\infty}^t \tau H(\tau) d\tau = \frac{t^2}{2!} H(t) \quad \dots(32)$$

and therefore, continuing this process

$$\frac{1}{p^n} H(t) = \frac{t^n}{n!} H(t) \quad \dots\dots(33)$$

It follows that if

$$\eta(p) = a_0 + \frac{a_1}{p} + \frac{a_2}{p^2} + \dots + \frac{a_n}{p^n} + \dots \dots(34a)$$

then

$$I(t) = [a_0 + a_1 t + (a_2 t^2/2!) + \dots + (a_n t^n/n!) + \dots] H(t) \quad \dots\dots(35a)$$

It is of course also in general perfectly possible to expand $\eta(p)$ in ascending powers of p instead of descending powers. In this way we might obtain

$$\eta(p) = b_0 + b_1 p + b_2 p^2 + \dots \quad \dots\dots(34b)$$

and thus

$$I(t) = \left[b_0 + b_1 \frac{d}{dt} + b_2 \frac{d^2}{dt^2} + \dots \right] H(t) \quad \dots\dots(35b)$$

If as in equation (35b) the operand is $H(t)$, equation (35b) merely tells us that a sufficiently long time after t is zero must elapse before $I(t)$ reaches its ultimate value b_0 . If the operand in (35b) is some other function $\phi(t)$, (35b) tells us the ultimate value of $I(t)$, a sufficiently long time after all discontinuous changes in $\phi(t)$ are over. To obtain this ultimate value, any term $\psi(t)H(t-\tau)$ contained in $\phi(t)$ should be differentiated neglecting any impulse terms which may arise. Equation (35a) on the other hand is primarily concerned with the behaviour of $I(t)$ at times in the immediate neighbourhood of zero. An

alternative expression of this fact is given by equations (51) and (52) below. Usually we are mainly concerned with the behaviour of a system immediately after a discontinuity, so that equation (35a) is a more useful form for $I(t)$ than equation (35b). The time required before equation (35b), possibly with the operand $H(t)$ replaced by $\phi(t)$, becomes relevant is of the order of several times the reciprocal of the decrement associated with the least-damped element of the system.

Now if the network with admittance $y(p)$ has only lumped elements, $y(p)$ will be a rational function of p , say $P(p)/Q(p)$, where P and Q are polynomials. If the input voltage $r(t)$ is of a reasonably simple kind, such as a step-voltage or the product of an exponential and a circular function of time or the sum of a number of such terms suddenly applied at time zero, then $f(p)$ will also be a rational function of p , so that $\eta(p)$ will also be such a function. The series for $\eta(p)$ can then be obtained (if there is no obvious and simpler way) by dividing the denominator of $\eta(p)$ into the numerator. Since the corresponding time-series (35a) has a general term $a_n t^n/n!$, there is in general little difficulty about its convergence. We shall not consider general conditions of convergence here, because series expansion was not Heaviside's only weapon for determining $I(t)$ given $\eta(p)$. It is sufficient to note that it has been said that if the series (34a) for $\eta(p)$ is convergent for any finite range of p , the corresponding series for $I(t)$ will be convergent for all t .

Two weapons frequently used by Heaviside to determine $I(t)$ given $\eta(p)$ are the so-called "shift theorems", namely that if

$$I(t) = [y(p)f(p)]H(t) = \eta(p)H(t) \quad \dots\dots(36)$$

then

$$I(t) = e^{at} \eta(p + \alpha) [e^{-at} H(t)] = e^{at} \eta(p + \alpha) \frac{p}{p + \alpha} H(t) \quad \dots\dots(37)$$

and

$$I(t - \lambda) = [e^{-p\lambda} \eta(p)]H(t) \quad \dots\dots(38)$$

Equations (37) and (38) can frequently be used to reduce $\eta(p)$ to a more manageable form, amenable to the series expansion (equation (34a)) already discussed, or to the partial-fraction expansion which is the last of Heaviside's own elementary weapons for extracting $I(t)$ from $\eta(p)$. Heaviside's expansion theorem amounts to saying that if $\eta(p)$ is in the form of a rational function of p whose numerator $N(p)$ is of lower degree than its denominator $D(p)$ and the zeros of the denominator are $\alpha_1, \alpha_2 \dots \alpha_n$ which may be real or complex but are distinct, then

$$\eta(p) = \sum_{r=1}^n \frac{A_r}{p + \alpha_r} \quad \dots\dots(39)$$

where $A_r = N(-\alpha_r)/D'(-\alpha_r)$ (40)

But $\frac{1}{p+\alpha_r} H(t) = \frac{1}{\alpha_r} (1 - e^{-\alpha_r t})H(t)$ (41)

if α_r has a positive real part as it would in the case under discussion where a voltage is applied to a passive network (the case of active networks is considered later). It follows that

$$I(t) = \sum_{r=1}^n \frac{A_r}{\alpha_r} (1 - e^{-\alpha_r t})H(t) \quad \dots\dots(42)$$

If k of the α_r coincide,† say $\alpha_1, \alpha_2 \dots \alpha_k$, then the terms

$$\frac{A_1}{p+\alpha} + \frac{A_2}{p+\alpha} + \dots + \frac{A_k}{p+\alpha} \quad \dots\dots(43)$$

in equation (39) can be replaced by

$$\frac{B_1}{p+\alpha_1} + \frac{B_2}{(p+\alpha_1)^2} + \dots + \frac{B_k}{(p+\alpha_1)^k} \quad \dots\dots(44)$$

and with the aid of equations (37) and (33) we find that

$$\frac{p}{(p+\alpha_1)^r} H(t) = e^{-\alpha_1 t} \frac{t^{r-1}}{(r-1)!} H(t) \quad \dots\dots(45)$$

The contribution to $I(t)$ from the terms (44) can be explicitly formulated by repeated application of equation (45), since

$$1 = \frac{1}{\alpha_1} \{(p+\alpha_1) - p\}$$

and thus the last of the terms (44) can be reduced to a term to which (45) applies directly, together with a term having a lower power of $(p+\alpha_1)$ in the denominator, and then the B_{k-1} term can be similarly reduced, and so on.

Alternatively, if $\eta(p) = f_1(p)f_2(p)$, $I(t)$ can also be obtained by means of the convolution integral, a simple derivation of which is given in Example 1 (Appendix 2).

Now it is possible (when $\eta(p)$ is a rational function with denominator of higher degree than numerator, as we have supposed) to determine the constants A_r in equation (40) and even the terms (44) by purely algebraic means. Alternatively, we can determine the A_r in equation (40) as the residues of the function $\eta(p)$ at its poles when p is regarded as a complex variable; this was the approach of Bromwich. This approach leads to the basic formula that if

$$h(t) = f(p)H(t)$$

† In practice the quantities α_r can always be treated as at worst metrically equal, that is to say, differing by arbitrarily small amounts different from zero. Equation (42) then applies, but will involve differences between large and nearly equal quantities, and it will usually be reasonably easy to see what happens as the differences between nearly equal roots all tend to zero.

where $f(p)$ is given $h(t)$ is required, then

$$h(t) = \frac{1}{2\pi j} \int_{-j\infty}^{j\infty} e^{zt} \frac{f(z) dz}{z} \quad \dots\dots(46)$$

This integral is not an integral round a closed contour, but it may be evaluated by contour-integral techniques. The contour is usually completed by a great semicircle to the left or to the right of the imaginary axis; the contribution of this great semicircle to the total is usually zero.‡

In order to appreciate fully the significance of equation (46), we must consider a special case, whose importance will be obvious from equation (45) with $r = 1$, namely

$$f(p) = f_1(p) = \frac{p}{p + \varepsilon + j\omega_0} \quad \dots\dots(47)$$

where ε is real and positive. If t is positive, the contour associated with the integral (46) must be completed by the infinite semicircle to the left of the imaginary axis, and the pole at $z = -\varepsilon - j\omega_0$ is therefore included; the residue is $\exp(-\varepsilon t - j\omega_0 t)$. On the other hand, if t is negative, the contour must be completed by the infinite semicircle to the right of the imaginary axis, which contains no pole. These facts can be combined into the single result

$$h(t) = h_1(t) = e^{-(\varepsilon + j\omega_0)t} H(t) \quad \dots\dots(48)$$

Similarly, it may be shown that the time function $h(t)$ such that

$$f(p) = f_2(p) = \frac{p}{p - \varepsilon - j\omega_0} \quad \dots\dots(49)$$

is $h(t) = h_2(t) = e^{(\varepsilon + j\omega_0)t} H(-t)$ (50)

In equation (46), the expression $f(z)/z$ will in general have poles which are all in the left half of the z -plane in the case of a passive circuit. There will not normally be a pole at $p = 0$ because it can be shown that§

$$\lim_{z \rightarrow 0} f(z) = \lim_{t \rightarrow \infty} h(t) = 0 \quad \dots\dots(51)$$

$$\lim_{z \rightarrow \infty} f(z) = \lim_{t \rightarrow 0+} h(t) \quad \dots\dots(52)$$

so that the apparent pole at the origin will have zero residue. Each simple pole will give rise to a term of the form (48) in $h(t)$. Multiple poles may give rise to combinations of terms of the form (45), the real part of α_1 being positive. If on the other hand there are

‡ The same result may be obtained by applying Jordan's lemma. In fact, however, the convergence of the integral (46) is much better than Jordan's lemma suggests.¹⁷

§ Equations (51) and (52) are the well-known Abel-Tauber relations; see van der Pol.¹¹ It is also worth more than passing notice that when $f(p)$ and $h(t)$ are connected by equations (25) and (46), $f(p)$ and $h(t)$ have the same dimensions, whereas in the Laplace-transform equations (60) and (61) an extra factor s is somewhat unnaturally required before this balance of dimensions can be restored.

any poles in the positive half-plane, they will give rise to terms of the form (50) in $h(t)$ (in the case of a multiple pole with the real part positive, the right-hand side of (50) will have correspondingly to be multiplied by a polynomial in t). These terms of the form (50) will only be effective for negative time, owing to the factor $H(-t)$. Now it has to be carefully noted that when $\varepsilon \rightarrow 0$, equations (48) and (50) do not give the same result. The significance of this can be seen with the help of equation (25). If in this equation $h(\lambda)$ is $\exp(k\lambda)$, then the integral will not converge at its lower limit if k has a negative real part unless there is a definite time at which the exponential starts, i.e., unless $h(\lambda)$ is really $\exp(k\lambda) \times H(\lambda + t_0)$. Likewise, if k has a positive real part, the integral will not converge at its upper limit unless there is a definite time at which the exponential finishes, so that $h(\lambda)$ is really $\exp(k\lambda)H(t_0 - \lambda)$. The physical interpretation of these facts (which were correctly represented by Heaviside) is in general terms as follows. A situation is represented by a mathematical statement of $f(p)$. $f(z)$ in equation (46) may have poles in either half of the complex plane. In respect of the poles in the left half-plane, the apparatus is in effect a passive network and is capable of continued existence in the world. In respect of poles in the right half-plane, the apparatus was in effect an unstable active network and could not remain in that state indefinitely. Its history can, however, be traced backwards in time from the moment at which specified values of the variables were reached. If the apparatus is brought into existence at a definite time, its future performance is indefinite because random noise, which is universally present in the world of experience, is sufficient to start an exponentially increasing explosion. The essential boundary between the poles in the positive half-plane (if any) which give information about the past behaviour of the apparatus as an active network and the poles in the negative half-plane which give information about its future behaviour as a passive network is the imaginary axis. This suggests that displacement of the path of integration in equation (46) from the natural boundary of the imaginary axis is a falsification of a fundamental nature.

By following Heaviside we have thus obtained an operational calculus in which the operator p is equivalent to d/dt , all operations are reversible, and the alternative specification $f(p)$ of a given function $h(t)$ can be determined, and vice versa. Cases such as that of frequency modulation, where the alternative specification of the frequency-modulated voltage cannot be satisfactorily formulated, can also be handled by this calculus. No difficulties about convergence and no ambiguities arise provided only that we extend the meaning of differentiation to apply to Heaviside's discontinuous unit step and impulse

functions. The authors believe that all is well as long as we follow Heaviside completely, accepting his premises and concepts and techniques as well as his results. We believe that the present fashion of regarding p as a number instead of as an operator limits the scope of operational calculus (because cases like that of frequency modulation cannot be satisfactorily handled) and introduces errors, particularly when the Laplace transform is used. These errors we consider next.

3. Fourier Analysis, Symbolic Calculus and Laplace Transforms: Difficulties and Errors associated with Departure from Heaviside

Various techniques now exist which appear to have the same object as Heaviside's operational calculus discussed in the last section. We shall now discuss these—the Fourier transform, symbolic calculus and Laplace transforms—in relation to Heaviside's calculus.

There is no doubt that the use of the Fourier integral has greatly increased the mathematical armamentarium available to operational calculus; indeed, Heaviside himself acknowledges his indebtedness to Fourier. But the all-important concepts of operands such as unit step and unit impulse, although latent, are often unobserved in the theory of Fourier transforms. In the Fourier approach there is an unreal implication that a "steady state" has been reached; in Heaviside's approach, there is a sudden start at a specified time, which is more in accordance with practice, and a complete solution involving both transient and "steady-state" terms is obtained, so it is not necessary to analyse the input voltage into a series or spectrum of sinusoidal components. Nevertheless, the Fourier identities can readily be derived from equation (8).

For

$$\delta(x) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \cos \omega x \, d\omega = \frac{1}{2\pi} \int_{-\infty}^{\infty} e^{j\omega x} \, d\omega \dots(53)$$

and substitution from equation (53) into equation (8) yields the Fourier identities

$$h(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(\lambda) \cos \omega(t - \lambda) \, d\omega \, d\lambda \dots(54)$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} e^{j\omega(t - \lambda)} h(\lambda) \, d\omega \, d\lambda \dots\dots(55)$$

This simple derivation of the Fourier identities was pointed out by Heaviside.† $\delta(x)$ can also be expanded

† Reference 1, Vol. II, p. 100, equations (54) and (55). A metrical version of the Fourier identity, which is free from the difficulties associated with equation (55), is discussed in reference 17.

in terms of infinite series of other orthogonal functions, and substitution into equation (8) leads to expansions of $h(t)$ in terms of such functions, e.g. Fourier series, Bessel series, Laguerre series, Legendre series, etc. Thus equation (8) is in a sense the prototype of such expansions.

Fourier analysis consists essentially in regarding any function of time as equivalent to a finite or infinite series of sinusoidal components. If the function of time is periodic with period $2\pi/\omega_0$ the frequencies of the sinusoidal components will all be multiples of ω_0 , but if the function is not periodic, the sinusoidal components form a continuous spectrum. Periodic or nearly periodic functions do frequently occur in practice, and well-known practical techniques exist for the harmonic analysis of a given waveform, but it nevertheless remains true that many functions of time arise in practical circuits which simply do not fit neatly into the harmonic mould. This is particularly true of those associated with single sudden starts or changes, and, as we have seen, Heaviside used his operator p and his step function and its derivatives to cope with just this kind of excitation. Heaviside, however, was primarily an engineer, rather than a mathematician; if with his calculus he obtained useful results from the practical point of view, he did not care if the mathematicians of his day disliked his methods for alleged lack of mathematical rigour. At about the time when the third volume of Heaviside's "Electromagnetic Theory" was being written, Heaviside was in touch with Bromwich, who had applied contour integrals to a number of problems solved by operational methods. It was evident to Heaviside and Bromwich that the contour integral and operational calculus were mutually adjutant and complementary. Thus began the era of mathematical justification and vindication of Heaviside. At first p was regarded as an operator, in accordance with Heaviside's practice; later the concept of "symbolic" calculus was brought into being, notably by van der Pol,² and the p -world was regarded as entirely distinct from the time world. The function $\phi(p)$ defined by equation (30) was regarded as the "image" or "operational transform" of the "original" time-function $v(t)$ while $v(t)$ could be obtained from $\phi(p)$ by means of the "Bromwich integral"

$$v(t) = \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} e^{pt} \phi(p) \frac{dp}{p} \quad \dots\dots(56)$$

which differs from equation (46) in certain important respects. c is a real positive constant, and is intended to be chosen so that the line $p = c$ in the complex plane shall be to the right of all poles of $\phi(p)/p$. Within the p -world, p was regarded as a number, and

sometimes $v(t)$ changes suddenly from one form to another when the real part of p crosses a particular boundary.

Thus although there is great similarity between symbolic calculus and Heaviside's operational calculus, there are differences between them. It is the authors' belief that these differences are significant and important, and that they are frequently the cause of serious errors committed unconsciously or only semi-consciously. All these can be avoided by returning to Heaviside's methods.

The first point is that in symbolic calculus equation (30) is often interpreted as meaning that $\phi(p)$ is the transform of $v(t)$, whereas it is in fact the transform of $v(t)H(t)$. This error is due to the bad notation of equation (30). A more subtle error is to read equation (30) as implying that $v(t)$ exists for $t < 0$, but the integral giving $\phi(p)$ is concerned only with the values of $v(t)$ for t positive. The Heaviside expression (equation (25)) makes it clear that the transform takes account of all the $v(t)$ there is. This is a point of fundamental significance and may be stated again thus—correct Heaviside integration never involves selection, and if integration over a range is omitted, it is because over this range there is zero integrand. It is always clearer to use complete limits as in equation (25) and to define the integrand properly where required by using $H(t)$.

Next, there is the question of the constant c which is present in equation (56) but not in equation (46). This is a conceptual error of great significance; here we merely point out a specific error to which it leads. If in equation (56), $\phi(p)/p$ has a pole in the positive half of the p -plane, with real part say α , the Bromwich integral, equation (56), directs us to make c greater than α , and the integral gives rise to a term

$$e^{(\alpha+j\beta)t} H(t) \quad \dots\dots(57)$$

which increases indefinitely and, as we have seen, could not have arisen in equation (25). The corresponding term that would be obtained by analogy with equation (50), however, raises no difficulty, and is the convergent result of equation (25).

Again, there is the question of the convergence of the integrals (25) and (30). If p is an operator and only admissible time-functions are involved, the integrals are automatically convergent. The factor $e^{-p\lambda}$ in the integrand is merely a shifting operator, and cannot have any effect on convergence.

In the absence of $e^{-p\lambda}$, the integrals must be convergent or the time functions would not be admissible. If, however, p is regarded as a number, another error of fundamental importance creeps in; it may then be postulated that p has a real part greater than α , so that the integral in equation (30) is convergent with $v(t)$ equal to the expression (57).

This as we have already seen is incorrect, but some very hard thinking is necessary to see that it is incorrect if we accept the premise of symbolic calculus that p is a number.

In symbolic calculus, p is a mere number, and there is therefore no difficulty about the order of operations. The products pp^{-1} and $p^{-1}p$ are both unity. But instead of regarding differentiation as multiplication by p in the p -world, we have the rule that, if $f(p)$ corresponds to (or is the operational transform or image of) $h(t)$, then

$$\frac{d^n h(t)}{dt^n} \text{ corresponds to } p^n f(p) - \sum_{s=0}^{n-1} p^{n-s} \left[\frac{d^s h(t)}{dt^s} \right]_{t=0} \dots\dots(58)$$

This is an error due to incorrect specification of the operand. The correct statement is that

$$\frac{d^n h(t)}{dt^n} \text{ corresponds to } p^n f(p) \dots\dots(59)$$

In equation (58) $h(t)$ should be replaced by $h(t)H(t)$; equation (58) is then a special case of equation (59).

In about 1937, the similarity of the classical Fourier transform to an integral transform, used in 1812 by Laplace, was noticed by a number of mathematicians, notably Doetsch.⁷ It became fashionable for many workers to base their analysis on the Laplace transform

$$\psi(s) = L[v(t)] = \int_0^\infty e^{-st} v(t) dt \dots\dots\dagger(60)$$

and its inverse

$$v(t) = \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} e^{st} \psi(s) ds \dots\dots\dagger(61)$$

The first of these transforms, equation (60), differs from equation (30) in that s is substituted for p (a trivial matter; it is preferable to have different variables for clarity) and in the omission of a factor s multiplying the right-hand side. The second correspondingly differs from equation (56) in the omission of a divisor s . The operational concepts of the original Heaviside calculus at this stage began to be relegated to oblivion, and indeed to be treated by some workers with contempt.

Equations (60) and (61) are open to all the objections already mentioned in connection with equation (56), and with equation (30) if the factor $H(\tau)$ is omitted, and there are also additional objections peculiar to these last equations. There is an implied assumption with the Laplace transform, as with the Fourier transform, that the exciting function (or force function) $v(t)$ is associated with unit impulse at time $t = 0$,[‡] whereas in Heaviside's operational

† See footnote to equations (51) and (52).

calculus or in symbolic calculus the assumption is that the exciting function is correspondingly associated with unit step. With the Fourier transform

$$G(\omega) = \int_{-\infty}^\infty e^{j\omega t} v(t) dt \dots\dots(62)$$

this causes no difficulty, but with equation (60) there is ambiguity, since the limits of integration are from 0 to ∞ , and the unit impulse at $t = 0$ is not clearly defined if negative values of t are excluded.

4. Operational Calculus in its Element: Fundamental Principles with special reference to Communication Networks and Modulation

Suppose we have a machine or instrument M which can be represented by a network N composed of ideal elements (capacitance, resistance and inductance) although M itself need not be electrical. Suppose that the input is $V(t)$ and the output $I(t)$. We can if we wish think as if $V(t)$ is a voltage applied to the network N and $I(t)$ is the resulting current, and the network N can be regarded as having an admittance $y(p)$.§ The behaviour of the network is completely represented by the operational equation

$$I(t) = y(p)V(t) \dots\dots(63)$$

The simplest case is that in which the input $V(t)$ contains no information or coherence. In a laboratory experiment $V(t)$ would then be $H(t)$ (or possibly $\delta(t)$) which contains no information except the epoch of switching on. In a field experiment, $V(t)$ would be noise, represented by $N(t)$; in a communication experiment $V(t)$ would be the unpredictable signal $S(t)$. In the general case, the input may have some degree of coherence or predictability; this implies the existence of constraints which may be represented by a network N' of admittance say $y'(p)$, so that for the cases cited above, namely laboratory experiments, field experiments and communication networks, we have respectively

$$V(t) = y'(p)H(t), \\ V(t) = y'(p)N(t) \text{ or } V(t) = y'(p)S(t) \dots(64)$$

The way in which equation (63) should be manipulated, after substitution from (64), depends on both the type of problem under consideration (i.e. whether the incoherent part of $V(t)$ is $H(t)$, $N(t)$ or $S(t)$) and the nature of the networks N and N'. In general, it is a problem in operational calculus which reduces to a problem in symbolic or Laplace-transform calculus only in special cases.

‡ This means that half of the contribution to $\delta(t)$ which lies in the region $-\infty < t < 0$ is lost, and we have $\int_0^\infty \delta(t) dt = \frac{1}{2}$

§ Recently the scope of the idea of impedance (or admittance) has been greatly extended, notably by Booker.¹⁴

The operational equation (63) answers implicitly the question "How is the receiver (or the network N representing it) affected by the input (or transmission) $V(t)$?" This question is perfectly sensible and reasonable. On the other hand the question "What components of frequency are contained in $V(t)$?" is not a correct question; it prejudices the issue, in regarding a frequency component as constant over all eternity. This is implied by the form of the Fourier integral, which would be upset by a drift of even 1 c/s per century. Gabor⁸ has pointed out that this does not correspond to our intuitive engineering concept of frequency, which is more akin to van der Pol's "instantaneous frequency"⁹ associated with his quasi-stationary theory of frequency modulation. But even the "instantaneous frequency" is not necessarily the "effective frequency" registered by a receiver. We shall see that a typical receiver takes a kind of running average of the signal over a finite number of cycles related to the receiver bandwidth. This agrees well with our ordinary engineering understanding, and introduces no mathematical difficulties concerning convergence.

Let us therefore next consider various ways in which the expression $y(p)V(t)$ can be manipulated. In some cases the explicit form of $I(t)$ is obvious, thus if $y(p) = p$, then $I(t)$ is simply $V'(t)$ and is thus completely determined. In general, however, it is necessary to express either $y(p)$ or $V(t)$ or both as a combination of terms (by means of power-series, series of orthogonal functions (Fourier, Laguerre, etc.), partial-fraction expansion, and so on) for each one of which the contribution to $I(t)$ is clear. Thus we might expand $y(p)$ in a Maclaurin series

$$y(p) = y(0) + py'(0) + \frac{p^2}{2!} y''(0) + \dots + \frac{p^r}{r!} y^{(r)}(0) + \dots \quad \dots(65)$$

from which it follows that

$$I(t) = y(p)V(t) = \sum_{r=0}^{\infty} \frac{1}{r!} y^{(r)}(0) \cdot \frac{d^r V(t)}{dt^r} \quad \dots(66)$$

This can also be written (in a form which leads easily to various solutions, particularly the "quasi-stationary" solution,⁹)

$$i(t) = V(t+q)y(p) \quad \dots(67)$$

where $q^r y(p)$ is an operational expression equivalent to

$$\left[\frac{d^r y(p)}{dp^r} \right]_{p=0}$$

Again, if
$$y(p) = \sum_{r=1}^m A_r e^{-\lambda_r p} \phi_r(p) \quad \dots(68)$$

that is to say, if $y(p)$ can be expressed in terms of delay cables and simpler admittances, then

$$I(t) = \sum_{r=1}^m A_r \phi_r(p) V(t - \lambda_r) \quad \dots(69)$$

This result is known as Heaviside's second shift theorem. Again, if the admittance is $y_1(p-p_0)$ instead of $y(p)$, we can write

$$I(t) = e^{p_0 t} y_1(p) [e^{-p_0 t} V(t)] \quad \dots(70)$$

This result is known as Heaviside's first shift theorem, and equation (37) is really a special case of it. In the general form just given, it is easily used; p_0 can often be chosen so that the product $\exp(-p_0 t) \cdot V(t)$ varies relatively slowly with time, and this may simplify the numerical determination of

$$y_1(p) [\exp(-p_0 t) \cdot V(t)]$$

should this be necessary.

Alternatively, if $y(p)$ is a rational function of p , as it often is, it can be expressed in partial fraction form

$$y(p) = a_k p^k + a_{k-1} p^{k-1} + \dots + a_1 p + a_0 + \sum_{r=1}^m \frac{A_r}{p-p_r} \quad \dots(71)$$

on the assumption that the denominator $D(p)$ of $y(p)$ is of degree m (having m distinct zeros $p_1, p_2 \dots p_m$, which may be real or complex, but will necessarily have negative real parts) and that the numerator $N(p)$ of $y(p)$ has greater degree $m+k$. The polynomial terms of $y(p)$ present no difficulty; the contribution of the term $a_r p^r$ to $I(t)$ is simply $a_r d^r V(t)/dt^r$. If the degree of the denominator of $y(p)$ exceeds that of the numerator, the polynomial terms are absent. In either case

$$A_r = \lim_{p \rightarrow p_r} [(p-p_r)y(p)] = N(p_r)/D'(p_r) \quad \dots(72)$$

so that the typical term we have to consider is

$$I_r(t) = \frac{A_r}{p-p_r} V(t) \quad \dots(73)$$

which yields, by the first shift theorem (equation (70))

$$I_r(t) = A_r e^{p_r t} \frac{1}{p} [e^{-p_r t} V(t)] \quad \dots(74)$$

In the exceptional case when the p_r are not all distinct (cf. expression (44)) we may have to handle a term of the form

$$J_r(t) = \frac{B_r}{(p-p_r)^k} V(t) = B_r e^{p_r t} \frac{1}{p^k} [e^{-p_r t} V(t)] \quad (75)$$

Normally, we shall be concerned with passive networks, so that p_r will have a negative real part. In this case there is no difficulty about equations (74) and (75): the operator $(1/p)$ is replaced by an integration from $-\infty$ to t in (74) and the operator $(1/p^k)$ by k successive integrations from $-\infty$ to t in (75). We have however already seen in connection with equations (48) and (50) that there is a clear-cut

dichotomy between the function $h_1(t)$ resulting from the operation of $p/(p + \varepsilon + j\omega_0)$ on $H(t)$ and the function $h_2(t)$ resulting from the operation of $p/(p - \varepsilon + j\omega_0)$ on $H(t)$. This dichotomy persists however small ε may be. A similar dichotomy is associated with equations (74) and (75) according to the sign of the real part of p_r . In equation (11) the operator p^{-1} was defined as an integration with lower limit $-\infty$; the vital equation (17) associated with commutativity would be unaffected if the lower limit had been $+\infty$ instead. Normally our ideas of the causal sequence force us to regard equation (11), which involves integrating all the integrand up to time t , as correct. These ideas are, however, ultimately based on the direction of running down or increasing entropy, in the case of irreversible processes. There are, nevertheless, certain exceptional situations in which the time direction of running down is abnormal, such as unstable amplifiers or atom bombs. In these exceptional situations, looking in the normal forward time direction will merely give warning of an impending explosion. If, however, the normal time sequence is reversed, we can obtain information of the way in which the contribution from a term $\{A_r/(p - p_r)\}V(t)$ reached its value at time t , p_r having a positive real part. In these exceptional situations, either the variable t should be replaced by $-t$ initially or the lower limit of each integration associated with p^{-1} or p^{-k} should be taken as $+\infty$ instead of $-\infty$.

In modulation problems two further simplifications occur, namely first

$$V(t) = e^{-j\omega_0 t} \psi(t) \quad \dots\dots(76)$$

where $\omega_0/2\pi$ is the carrier frequency and $\psi(t)$ is the modulation. Secondly, the network N representing the receiver will usually be a confluent band-pass structure (or approximately so). This means that every reactance in N is tuned at carrier frequency with the appropriate opposite reactance in parallel or in series (except where the effect of such tuning is negligible). In symbols we can therefore write

$$y(p) = y_0 \left(\frac{p}{\omega_0} + \frac{\omega_0}{p} \right) = y_0(q) \quad \dots\dots(77)$$

where $q = (p/\omega_0) + (\omega_0/p)$. Now $y_0(q)$ will usually be a rational function of q , and can be split up into partial fractions of which a typical one is

$$y_r(q) = \frac{C_r}{q - \alpha_r} \quad \dots\dots(78)$$

where C_r, α_r are constants which may be complex, and α_r has a negative real part and modulus small compared to unity. (There may also be terms involving integral powers of q (cf. equation (71) above) but these present no difficulty.) But $y_r(q)$ can itself be split up into partial fractions with respect to p , namely

$$y_r(q) = \frac{jC_r}{2\beta_r} \left[\frac{\gamma_{2r}}{p - \gamma_{2r}} - \frac{\gamma_{1r}}{p - \gamma_{1r}} \right] \quad \dots\dots(79)$$

where

$$\left. \begin{aligned} \beta_r &= [1 - (\alpha_r^2/4)]^{1/2} \\ \gamma_{1r} &= \omega_0[(\alpha_r/2) + j\beta_r] \\ \gamma_{2r} &= \omega_0[(\alpha_r/2) - j\beta_r] \end{aligned} \right\} \quad \dots\dots(80)$$

It follows that

$$\begin{aligned} I_r(t) &= y_r(q) e^{-j\omega_0 t} \psi(t) \\ &= \frac{jC_r}{2\beta_r} \left[\frac{\gamma_{2r}}{p - \gamma_{2r}} - \frac{\gamma_{1r}}{p - \gamma_{1r}} \right] e^{-j\omega_0 t} \psi(t) \quad \dots\dots(81) \end{aligned}$$

The second term, by using Heaviside's first shift theorem with $p_0 = -j\omega_0$, can be written

$$\text{constant} \times e^{-j\omega_0 t} \frac{\gamma_{1r}}{p - \gamma_{1r} - j\omega_0} \psi(t) \quad \dots\dots(82)$$

and is therefore negligible unless the modulation $\psi(t)$ contains frequencies near twice the carrier frequency (so that $p - \gamma_{1r} - j\omega_0$ has roughly the effect of a small denominator). This term gives the "image response"; it can be ignored unless $\psi(t)$ does contain the aforementioned high frequencies, and if it cannot be ignored, its effect can be removed by means of suitable additional networks. The remaining term can be written, by means of Heaviside's first shift theorem, with $p_0 = \gamma_{2r}$, in the form

$$I_r(t) = \frac{jC_r}{2\beta_r} \gamma_{2r} e^{\gamma_{2r} t} \int_{-\infty}^t e^{-(\gamma_{2r} + j\omega_0)\tau} \psi(\tau) d\tau \quad \dots\dots(83)$$

since the operator $(1/p)$, which remains after the shift theorem has been applied, means integration from $-\infty$ to t as noted earlier (equation (74)). A slight algebraic rearrangement of this last result gives us finally (neglecting the image response)

$$I_r(t) = \frac{1}{2} j e^{-j\omega_0 t} \cdot \frac{C_r \gamma_{2r}}{\beta_r} \int_{-\infty}^t e^{(\gamma_{2r} + j\omega_0)(t - \tau)} \psi(\tau) d\tau \quad \dots\dots(84)$$

For a practical network, α_r will always have a negative real part and $|\alpha_r|$ will be small. Hence β_r will be almost purely real and γ_{2r} will also have a negative real part; the imaginary part of $(\gamma_{2r} + j\omega_0)$ will be small. The modulation function $\psi(\tau)$ will therefore be multiplied in the integral by a factor which decreases exponentially as we proceed from the time t under consideration into the past. The receiver output $I(t)$ thus consists of a number of terms, of which $I_r(t)$ is typical; each represents an oscillation at carrier frequency, and has an amplitude and phase determined by a kind of weighted, running average over the immediate past.

As the real part of γ_{2r} is $\omega_0 \alpha_r/2$ (which is negative), at time $2\pi/\omega_0 |\text{Re } \alpha_r|$ sec before time t the exponential factor in the integrand will be $\exp(-2\pi)$ which is

about 0.000185. If we regard the fourth power of this quantity as negligible, we can say that this weighted average extends for the term $I_r(t)$ over $4/|\operatorname{Re} \alpha_r|$ carrier cycles, and therefore for the whole of $I(t)$ over $4/\alpha$ carrier cycles, where α is the least value of $|\operatorname{Re} \alpha_r|$ for varying r . Since α_r is small, perhaps of order 0.04 and mainly real, this suggests that the running average extends over 100 carrier cycles or so—a time long compared with one carrier cycle (at say 10.7 Mc/s) but short compared with one cycle of the highest frequency (say 10 kc/s) likely to be involved in the modulation $\psi(t)$ if this is associated with an audio-frequency programme.

It should be clearly appreciated that the formula we have obtained above for $I_r(t)$ has been derived without making any approximation whatever (other than discarding the “image” term) and without asking any questions about the modulation $\psi(t)$. It can therefore be applied to any kind of modulation, including for example impulsive interference, in which $\psi(t)$ may have discontinuities.

It will also be appreciated that this solution is a *complete* solution, and not merely the steady-state part thereof. The quantities effective in this solution are not the conceptual Fourier frequencies in the modulation $\psi(t)$, nor even the “instantaneous frequencies” of van der Pol, but the “effective frequencies” that would be observed by a receiver. This agrees very well not only with ordinary engineering practice, but also with what we believe to be the ideas of Heaviside, as well as with the ideas of Wittgenstein¹⁵ and Dingle.¹⁶

5. Conclusions

We have endeavoured in the foregoing to set out explicitly the operational calculus which Heaviside himself used with such consummate skill; Heaviside knew exactly what he could do with this calculus, but he did not explain it so that others could operate it with equal facility (particularly if they were initially prejudiced against him).

Owing to the climate of prejudice which existed in Heaviside's lifetime, other workers have tended to try to obtain Heaviside's results by other methods, or to improve upon his methods. We believe, as explained in Section 3, that these other workers have thus introduced errors and difficulties which were not present in Heaviside's original work, and which would not have occurred if they had been content to keep within the landmarks laid down by Heaviside. The fact that Heaviside's original calculus was operational in concept is in particular an advantage, rather than a stumbling-block.

6. Acknowledgments

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These additional references can be divided into three general categories, namely (A) references significantly associated with the historical development of the subject, (B) authoritative references, where proofs or fundamental points are discussed or in some cases originated, (C) elucidatory references, in many of which points shortly stated in the text have been discussed more fully.

This bibliography is intended only to be initially adequate, and no claim for its completeness is put forward. In a few cases the particular part of a reference which is specially relevant is briefly indicated and explained.

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9. Appendix 1

A Continuous and Differentiable Function Simulating a Switching Process in the Limit

We could approach the question of nominal or conceptual entities mathematically, but it is more illuminating not to disdain the help of physical intuition. If there is a voltage V across a conceptual pure unit capacitance and the current flowing in the capacitance is I , then

$$I = \frac{dV}{dt} = pV \quad \text{or} \quad V = \int_{-\infty}^t I(\lambda) d\lambda = \frac{1}{p} I \dots (85)$$

In the second of equations (85), the lower limit of integration does not necessarily mean a large negative time, but simply a time before the beginning of the current I so that the integration includes *all* the charge on the capacitance. If we know that $I = 0$ before time T_1 , the lower limit can be replaced by $T_1 - \epsilon$. This, however, is a special case. In general, we suppose that I had a beginning so that $I(-\infty)$ can be taken as zero. Otherwise, the study of I is a theological problem, in the province of an archbishop, not of a physicist.

Clearly what the capacitance does is neither differentiation nor integration, but a process which regarded as determining I for given V is differentiation, and regarded as determining V for given I is integration. Likewise a monetary transaction appears as a debit to the buyer and a credit to the seller: the transaction, however, is the same entity, whether the buyer's viewpoint or the seller's is taken by an observer. p and p^{-1} are thus "logical reciprocals" (operators each of which can express the same process as the other, but the other way round) and this is the real significance of the equation

$$pp^{-1}V = p^{-1}pV = V \quad \dots\dots (86)$$

on which we have laid such emphasis. In the actual world, however, no capacitance is conceptually pure, but has associated with it shunt conductance ϵ and series resistance ν so that equation (85) should be replaced by

$$\left. \begin{aligned} I &= \frac{p}{1 + \nu p} V \\ V &= \frac{1}{p + \varepsilon} I \end{aligned} \right\} \dots\dots(87)$$

where ε, ν are arbitrarily small but not zero; the product $\varepsilon\nu$ can be neglected, and so can numerator terms involving ε and ν .

Even if a pure capacitance could exist, the observation of it would constitute a non-zero dissipation, so that a pure capacitance is not even observable. The capacitance is a mere representation of a form of energy storage; any other form of energy storage would give similar results. Conceptual pure differentiation and integration are likewise not observable.

Solving equations (87), we have

$$I = \frac{p}{1 + \nu p} V; \quad I(t_0) = e^{-t_0/\nu} \int_{-\infty}^{t_0} e^{t/\nu} \frac{dV}{dt} dt \dots(88)$$

$$V = \frac{1}{p + \varepsilon} I; \quad V(t_0) = e^{-\varepsilon t_0} \int_{-\infty}^{t_0} e^{\varepsilon t} I dt \dots\dots(89)$$

If $V = H(t)$, I becomes the metrical representation $\delta_m(t)$ of $\delta(t)$ given by

$$\delta_m(t) = \frac{p}{1 + \nu p} H(t) = (1/\nu) e^{-t/\nu} H(t) \dots\dots(90)$$

and if $I = \delta_m(t)$, V becomes the metrical representation $H_m(t)$ of $H(t)$, namely

$$H_m(t) = \frac{1}{p + \varepsilon} \delta_m(t) = [e^{-\varepsilon t} - e^{-t/\nu}] H(t) \dots(91)$$

Equation (90) gives an observable representation of $\delta(t)$ and equation (91) gives an observable representation of $H(t)$ which is continuous at $t = 0$; both expressions can approach their ideal or conceptual counterparts arbitrarily closely if ν, ε are sufficiently small.

$\delta(t)$ has "nominal" derivatives $\delta'(t), \delta''(t)$ etc., for which metrical counterparts can be found similarly; these can be used in all cases of difficulty, but in most cases the "nominal" expressions will give the same result with less labour.

We regard all functions which correctly represent observations in nature as continuous and differentiable or integrable any number of times. The "nominal" or "conceptual" functions $H(t)$ and $\delta(t)$ may be associated with ordinary analytic functions; usually the differentiations and integrations can be done using "nominal" derivatives of $H(t)$ and $\delta(t)$ as required. In cases of difficulty these nominal derivatives can always be replaced by "metrical" derivatives obtained by substituting $\delta_m(t)$ (equation

(90)) for V in equation (88) to obtain $\delta'_m(t)$ (the first metrical derivative of $\delta(t)$) as the resulting I ; substituting $\delta'_m(t)$ for V in equation (88) to obtain $\delta''_m(t)$ (the second metrical derivative of $\delta(t)$) as I , and so on.

10. Appendix 2: Examples

1. The Convolution Integral: An Alternative Derivation to that of van der Pol¹¹

Suppose we have a piece of apparatus A which can be represented by a network N_1 (having ideal elements) for which the input voltage is V_{11} and the output voltage is V_{12} , and that

$$V_{12} = f_1(p)V_{11}$$

Then if $h_1(t) = f_1(p)H(t)$

we have a further picture of the same network N_1 , that its response to unit-step input is $h_1(t)$. $f_1(p)$ is thus a "transfer-voltage picture" and $h_1(t)$ a "unit-step-response picture" of the same network N_1 .

Now suppose that we have a second similar network N_2 whose input voltage is V_{21} and output voltage V_{22} , and that

$$V_{22} = f_2(p)V_{21} \quad \text{and} \quad h_2(t) = f_2(p)H(t)$$

so that $f_2(p)$ and $h_2(t)$ are respectively the "transfer-voltage picture" and the "unit-step response picture" of this second network N_2 . Suppose further that the networks N_1 and N_2 are connected in cascade, so that $V_{12} = V_{21}$.

Then $V_{22} = f_2(p)f_1(p)V_{11}$

so that the "transfer-voltage picture" of the combination is obtained by a mere multiplication of the separate "transfer voltage pictures". The question is how the "unit-step response picture" $h_3(t)$ of the combination is related to those of the separate networks. By operational calculus this can be derived immediately. For

$$\begin{aligned} h_3(t) &= f_2(p)f_1(p)H(t) \\ &= f_2(p) \left[p \int_{-\infty}^{\infty} e^{-p\lambda} h_1(\lambda) d\lambda \right] H(t) \end{aligned} \quad \text{from equation (24)}$$

$$\begin{aligned} &= \int_{-\infty}^{\infty} h_1(\lambda) \cdot [p e^{-p\lambda} f_2(p)H(t)] d\lambda \\ &= p \int_{-\infty}^{\infty} h_1(\lambda) [e^{-p\lambda} f_2(p)H(t)] d\lambda \end{aligned}$$

But $f_2(p)H(t) = h_2(t)$ by definition, and $e^{-p\lambda}h_2(t) = h_2(t - \lambda)$ (equation (22b)), and p is d/dt by definition, so that

$$h_3(t) = \frac{d}{dt} \int_{-\infty}^{\infty} h_1(\lambda)h_2(t - \lambda) d\lambda$$

In the special case when $h_1(t)$ and $h_2(t)$ both start at $t = 0$ so that

$$h_1(t) = \phi_1(t)H(t); \quad h_2(t) = \phi_2(t)H(t)$$

this reduces to

$$h_3(t) = \int_{-\infty}^{\infty} h_1(\lambda)h_2'(t-\lambda) d\lambda \\ = \left[\int_0^t \phi_1(\lambda)\phi_2'(t-\lambda) d\lambda + \phi_1(t)\phi_2(0) \right] H(t)$$

2. Impulsive Input to a Discriminator

For the Foster-Seeley discriminator arrangement discussed in reference 10, the relations between the primary voltage V_1 , secondary voltage V_2 and input current I_1 were found to be

$$\frac{V_1}{I_1} = \frac{q + \alpha_1}{(q + \alpha_1)(q + \alpha_2) - (\lambda^2/p^2)}$$

$$\frac{V_2}{I_1} = \frac{\lambda}{p\{(q + \alpha_1)(q + \alpha_2) - (\lambda^2/p^2)\}}$$

where α_1 and α_2 are the primary and secondary loss ratios, λ is the coupling coefficient between primary and secondary, q means the operator $p + 1/p$ and the equations are normalized so that the primary and secondary inductances are taken as equal and tuned to a frequency $1/2\pi$. The unit of time is thus $1/\omega_0$ seconds where $\omega_0/2\pi$ is the carrier frequency or i.f.

For simplicity we shall take $\alpha_1 = \alpha_2 = \alpha$; α and λ are conditioned by the relative bandwidth of the discriminator and will be small (in the case discussed in reference 10 they were of order 0.04, for ω_0 corresponding to 10.7 Mc/s).

By adding and subtracting the above equations for V_1 and V_2 (with $\alpha_1 = \alpha_2 = \alpha$) we obtain, after substituting $(p + 1/p)$ for q ,

$$V_1 + V_2 = \frac{p}{p^2 + \alpha p + (1 - \lambda)} I_1 \\ = \frac{1}{\eta_1 - \eta_2} \left[\frac{\eta_1}{p - \eta_1} - \frac{\eta_2}{p - \eta_2} \right] I_1$$

$$V_1 - V_2 = \frac{p}{p^2 + \alpha p + (1 + \lambda)} I_1 \\ = \frac{1}{\eta_3 - \eta_4} \left[\frac{\eta_3}{p - \eta_3} - \frac{\eta_4}{p - \eta_4} \right] I_1$$

where η_1 and η_2 are the roots of $p^2 + \alpha p + (1 - \lambda) = 0$ so that

$$\eta_1 = -\frac{\alpha}{2} + j\zeta_1 \quad \eta_2 = -\frac{\alpha}{2} - j\zeta_1 \quad \zeta_1 = \left(1 - \lambda - \frac{\alpha^2}{4} \right)^{\frac{1}{2}}$$

and η_3, η_4 are the roots of

$$p^2 + \alpha p + (1 + \lambda) = 0$$

so that

$$\eta_3 = -\frac{\alpha}{2} + j\zeta_2 \quad \eta_4 = -\frac{\alpha}{2} - j\zeta_2 \quad \zeta_2 = \left(1 + \lambda - \frac{\alpha^2}{4} \right)^{\frac{1}{2}}$$

If $I_1 = e^{-jt} \psi(t)$, that is to say, we have a carrier of frequency $1/2\pi$ and modulation $\psi(t)$ which may be of any kind, even having discontinuities in the event of impulsive interference, we can apply the "first shift-theorem" repeatedly (equation (70)); we thus obtain

$$V_1 + V_2 = \frac{je^{-jt}}{2\zeta_1} \left[\eta_2 \int_{-\infty}^t e^{(\eta_2 + j)(t-\tau)} \psi(\tau) d\tau - \eta_1 \int_{-\infty}^t e^{(\eta_1 + j)(t-\tau)} \psi(\tau) d\tau \right]$$

$$V_1 - V_2 = \frac{je^{-jt}}{2\zeta_2} \left[\eta_4 \int_{-\infty}^t e^{(\eta_4 + j)(t-\tau)} \psi(\tau) d\tau - \eta_3 \int_{-\infty}^t e^{(\eta_3 + j)(t-\tau)} \psi(\tau) d\tau \right]$$

and so far there has been no approximation at all, except the simplification of making $\alpha_1 = \alpha_2 = \alpha$. $(\eta_2 + j)$ and $(\eta_4 + j)$ are both small with negative real parts, whereas $(\eta_1 + j)$ and $(\eta_3 + j)$ are both complex numbers with small negative real parts and imaginary parts nearly equal to $2j$. Hence the terms involving η_1 and η_3 are the terms we called "image response" terms, and neglected in Section 4 above. If $\psi(t)$ has discontinuities, the integrals involved in the above expressions for $(V_1 + V_2)$ can still be evaluated, and as in Section 4, the discriminator-receiver is essentially taking a weighted running average of the values of the modulation $\psi(t)$ in the present and the immediate past, say the last 100 carrier-cycles. If $\psi(t)$ contains no high frequencies, as in the frequency modulated audio signal case of Section 4, the exponentials in the "image" η_1 -term and η_3 -term are oscillating so rapidly that these integrals are inevitably small. If however $\psi(t)$ has a discontinuity, the balance of the cycle of carrier-frequency centred about the point of discontinuity may be grossly disturbed, so that this cycle makes a far greater contribution to the "image-term" integrals than any other such cycle. It may therefore be necessary to retain the image-terms, or to provide special additional networks to nullify their effect. In the terms of $(V_1 \pm V_2)$ involving η_2 and η_4 , discontinuities in $\psi(t)$ will also mean that an important contribution to the total value of the integral will come from the immediate neighbourhood of the discontinuity, but the importance of this will decrease as the discontinuity fades into the past and will become negligible after say 100 carrier-frequency cycles.

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SPACE RESEARCH EXPERIMENTS

International Time Comparison by Telstar

An experiment was performed on 25th August, 1962 making use of the TELSTAR communications satellite to relate the time standards of the United States and the United Kingdom more accurately than had previously been possible. The experiment was conducted by the U.S. Naval Observatory via the American Telephone and Telegraph Company's ground station at Andover (Maine) and by the National Physical Laboratory via the British Post Office Satellite Ground Station at Goonhilly (Cornwall), and determined with high precision the time difference between the reference clocks of the U.S. Naval Observatory and the Royal Greenwich Observatory. Previously, the indications of these clocks had been compared to 1 or 2 milliseconds, which is the limiting accuracy possible using conventional radio time signals transmitted over long distances by routes depending on reflection within the variable ionosphere. The accuracy achieved during the 424th orbit of TELSTAR was 10–20 microseconds and it is now known that the two clocks differ in their readings by 1980 microseconds. The residual uncertainty is probably in terminal equipments and links, hence further improvement in accuracy should be possible.

The TELSTAR experiment constitutes a notable advance in the accuracy of co-ordination between the U.S. and U.K. time services, first established in 1959. This co-ordination is a basic link in the world-wide system of time and frequency reference. The accuracy of time comparisons has previously been limited by variations and uncertainties in the time of propagation of signals transmitted by long and short radio waves. The result will be of use in verifying theoretical predictions of the characteristics of electromagnetic wave propagation, in improving the accuracy of radio-navigational systems and in maintaining a uniform time standard in the two countries. It also provides the basis for better co-ordination between satellite tracking stations.

The experiment involved the simultaneous transmission of precise time signals in both directions over the TELSTAR circuit, the satellite being directly excited through the ionosphere by the microwave transmitters at Andover and Goonhilly. These signals were obtained from the quartz clocks already installed at each station as part of the equipment of the TELSTAR communication system. The times of reception of the received signals were measured with reference to those generated locally. Because the transmission path from East-to-West and West-to-East is symmetrical it is possible to eliminate from the comparisons at each station the common delay due to propagation of the signals along the path of length 10 000–15 000 miles. The difference between the reception times recorded at Andover and Goonhilly was a direct and independent

measure of the difference in time existing between the respective station clocks. To complete the experiment it remained to relate these time differences to the national time standards at the U.S. Naval Observatory, Washington, and the Royal Greenwich Observatory, Herstmonceux. This was done using the long-wave transmissions of radio time signals available in each country. Over short ranges, utilizing ground-wave propagation, these have stable and reproducible characteristics. On this side simultaneous measurements were made at Herstmonceux and Goonhilly of the reception time of seconds pulses radiated by MSF on 60 kc/s from the Post Office Radio Station at Rugby. The MSF transmitter is operated by the Post Office on behalf of the National Physical Laboratory and its signals have been synchronized as closely as possible by existing techniques with those of the U.S. and other countries. It is estimated that the clocks at the two British locations were related with an uncertainty of less than 20 microseconds. This type of experiment makes it possible to determine directly the signal delay across the Atlantic to better than 0.1 millisecond.

The Canadian Topside Sounder Satellite

On 30th September, the Canadian space satellite, *Alouette*, was successfully launched from the Vandenberg Air Force Base, California. The satellite, which weighs 320 lb, was placed into a near polar orbit with an apogee of 650 miles and perigee of 600 miles, circling the earth every 105 minutes.

The *Alouette* is designed to carry out ionospheric investigations by means of the top side sounder technique and a full description of the satellite and its design was given in a paper read at the Institution's 1961 Convention.† The power supply of the satellite is obtained from solar cells and it is expected to remain in orbit for about a year. Data are transmitted to ground control stations in Ottawa and the North West Territories and to a number of other stations throughout the World. Ionospheric measurements on the F2 layer are made for the band 1.6 Mc/s to 11.5 Mc/s. The provision of long dipole aerials for this waveband is a feature of particular interest since these long thin metal strips which were extended to lengths of 150 ft and 75 ft after achieving orbit are believed to be the largest aerials to have been used on space vehicles to date. Equipment is also installed in the satellite for measuring ionospheric noise.

The satellite itself was built in Canada and the launching by the National Aeronautics and Space Administration was carried out as part of NASA's programme of co-operative space research.

† R. C. Langille and J. C. W. Scott, "The Canadian Defence Research Board topside sounder satellite" *J. Brit.I.R.E.*, 23, pp. 61–8, January 1962.

Measuring Circuits used in an Automatic Transistor Tester

By

J. J. SPARKES †

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd-4th April 1962.

Summary: A machine which is capable of testing, sorting and measuring all the significant parameters of germanium r.f. alloy transistors at a rate of 1200 transistors per hour is described. Since the machine uses essentially only one test head, each measuring circuit, which is mounted on a printed circuit card, is connected in turn to this test head by means of dry reed relays. The paper describes how 10 d.c. parameters and 7 a.c. measurements and tests are carried out on each transistor and explains some of the circuits in detail. In addition the relative merits of single test station and multi-test station machines are discussed.

1. Introduction

The design of measuring circuits for use in an automatic testing machine primarily involves adapting established measuring techniques to the requirements of automatic operation and of their inclusion in a large machine. This usually means that some form of digital or quantized measurement must be used and that special arrangements have to be made to eliminate measurement errors.

Since the overall design of the machine greatly influences the form the circuits must take, a description of the machine will first be given. This will be followed by some examples of the circuits themselves and finally some remarks will be made regarding the accuracy achieved and the precautions taken to maintain accuracy.

2. The Grading Machine

Broadly speaking there are five classes of measurement that have to be made in the course of transistor manufacture. These are:

- (1) Go/no-go tests for sorting and batching transistors.
- (2) Measurements of parameter values for quality control.
- (3) Measurements of parameter values in the course of life testing.
- (4) All the above three items on each type of transistor produced.
- (5) All the above four items at various ambient temperatures.

A machine capable of achieving all these functions would be rather cumbersome, and in any case would be

somewhat inefficient since not all its capabilities could be used simultaneously. In practice therefore it is both sensible and expedient to attempt only a limited number of the functions by one machine. Other machines can be made to fill in the missing items.

In the production line for which the present machine was designed, the number of one type of transistor to be manufactured was more than adequate to keep one machine fully occupied, so that item (4) above was deleted from the list, and the machine was designed to deal only with germanium r.f. alloy transistors with cut-off frequencies in the range 2 to 25 Mc/s.

In general, high (or low) temperature measurements have only to be made on particularly temperature sensitive parameters. Since these form only a small proportion of the whole range of measurements to be made, this item also was omitted from the list, although it was realized that 100% testing at elevated temperature is normally necessary and always desirable.

In order to satisfy the first three requirements listed above the following features have to be incorporated:

For go/no-go testing it is necessary for the operator to be able to set up limit values for a variety of parameters (e.g. breakdown voltage, cut-off current, current gain, cut-off frequency, etc.) and to sort the transistors according to various combinations of pass and fail on each test. In general, failure of any one test does not mean the transistor is a reject, it means it falls into a lower category of transistor (e.g. lower voltage, lower frequency, etc.). Of course, for every parameter there is a value so low that the transistor is useless on that account alone, so that in practice it is usually necessary to set up two (or more) limit values. If the transistor fails the lower one it is a reject, but if it fails the higher one it falls into a different type number.

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In order to carry out this sorting process, in which passes and failures on a variety of parameters have to be combined in a variety of ways so that the type number of each device can be decided, it is necessary to store the results until all the measurements are completed and then reach a decision. This suggests strongly the technique of doing all the measurements on one transistor, sorting it, and then starting on the next. The alternative technique of having a test head for each measurement, so that transistors are measured in parallel, involves much greater storage facilities, although it greatly simplifies the measurement technique, as will be shown later, as well as easing the interchangeability of test circuits, and speeding up the testing rate.

For the *measurement* of each parameter, either for quality control or for life testing, it is necessary to produce some permanent record of each parameter value as it is determined. This suggests a digital method of measurement, the simplest form of which consists of a series of go/no-go tests around limits whose values depend on the previous measurement. For example, to measure a cut-off frequency value of 11.5 Mc/s, limit values of 8, 4, 2 and 1 Mc/s, or any additive combination of them can be made available. The first go/no-go test sets up 8 Mc/s. The transistor will pass this test. The limit value of $8 + 4$ is then tried and the transistor fails, so $8 + 2$ is tried. The transistor passes this so $8 + 2 + 1$ is tried and is again passed. It can therefore be concluded that the transistor's cut-off frequency is between 11 and 12 Mc/s. This is essentially the technique used in all the measuring circuits in the machine under consideration, and is clearly compatible with the go/no-go tests required for the sorting process. It also lends itself to punching out results on paper tape. The range of values covered by the circuit can in all cases be altered to give greater or less sensitivity. In the example just quoted, fifteen sub-ranges between 0 and 15 Mc/s are available, but this can be changed to 15 1-Mc/s sub-ranges between 16 and 32 Mc/s, or to 15 2-Mc/s sub-ranges between 0 and 32 Mc/s. The controls to implement these alterations of sensitivity and range are called "gain and shift" controls.

Results punched out on paper tape are a convenient form of presentation of life test results, since results on a set of transistors can be punched out before and after life test, and by means of a tape comparator any changes in parameter values can immediately be found.

Such results are inconvenient for quality control, however, since for this purpose an immediate picture of the spread of parameter values is required. To achieve this in the present machine, counters are operated so that they record the number of transistors falling in each sub-range of the parameter involved. From these numbers, histograms can immediately be

drawn. In its present form the machine enables histograms to be "displayed" for any seven parameters.

As indicated earlier, the machine is essentially a single-test-station machine, and this greatly simplifies the data handling problems. However, it brings with it three attendant disadvantages. First the transistor must remain at the test station long enough for all the necessary measurements to be made. Up to 18 parameters can be tested and measured on this machine, and 100 milliseconds is allotted to each parameter, so that the transistor dwells for 2 seconds at the test station. Taking into account about 1 second for transistor transport this limits the speed of the machine to about 1200 transistors per hour. With parallel measurement on a multi-station unit the limiting factor is the mechanical transport of the devices, and readily allows about a five-fold increase in testing speed. The first disadvantage in the single-station tester is therefore its speed of operation.

The second is the difficulty of designing accurate measuring circuits all of which are connected to one test station. In this machine each test circuit is connected in turn to the test head by means of dry reed relays. This presents no problem with the d.c. tests, but with the a.c. tests the relays and the lengthy leads between the test head and the measuring circuits present special problems and these will be discussed.

The third disadvantage is that interchangeability of test circuits is considerably diminished, since some of the a.c. circuits can only be made to yield accurate results by carefully balancing stray capacitances, etc. A portion of these strays are contributed by neighbouring circuits and any change in the circuits changes the strays and upsets the accuracy of the critical circuits. This objection does not normally apply to d.c. circuits.

Despite these disadvantages, which were well appreciated at the initial design stage, it was thought that none of them was likely to cause serious trouble, whereas the necessity of handling about 800 bits of information in shift registers, if a multi-station tester were built, and the expense this would involve, led to the choice of a single-test-station machine.

3. The Measuring Circuits

The machine in its final form measures 17 parameters, 10 d.c. and 7 a.c. as follows:

- (1) Collector-base breakdown voltage $V_{(BR)CBO}$
- (2) Collector-base breakdown voltage $V_{(BR)CBX}$
- (3) Emitter-base breakdown voltage $V_{(BR)EBX}$
- (4) Collector-base cut-off current I_{CBO}
- (5), (6), (7) Direct current gain at a specified value of $V_{CE(sat)}$ for three different values of collector current.

- (8) Emitter-base saturation voltage $V_{BE(sat)}$ at preset values of I_B and I_C .
- (9) Collector emitter breakdown voltage $V_{(BR)CEO}$.
- (10) Emitter-base voltage V_{BE} at preset values of I_B and V_{CE} .
- (11) Small-signal, low frequency current gain h_{fe} .
- (12) Small signal, high frequency current gain $|h_{fe}|$. This measurement yields a value of f_T .
- (13) Common-base output capacitance c_{ob} .
- (14) Extrinsic base resistance ($r_{bb'}$) by a feedback method.
- (15) Turn-off charge (Q_{OFF}) at specified values of I_C , I_B and $V_{CE(OFF)}$.
- (16) Turn-on charge (Q_{ON}) at specified values of I_C and $V_{CE(OFF)}$.
- (17) Noise factor at 1000 c/s.

A description of all these circuits would be unnecessarily lengthy. Accordingly only one d.c. test circuit will be described since these measurements present no particular problems, and only the more troublesome a.c. measurements will be described.

3.1. General Comments

(1) Since it is impossible to measure 1000 c/s noise factor in 100 milliseconds, a separate test head has been provided for noise measurement, which is located immediately prior to the main test station. Thus while 16 parameters are being measured on one transistor, noise is being measured on the next one. The result of this noise measurement is then regarded as the first result to be obtained on this next transistor, and is printed out accordingly. The remaining 16 results are obtained when the transistor has moved to the next test station. Thus parallel measurements are made but no storage of information is needed, and logically the machine is still a single-test-station machine.

(2) Since all tests are carried out sequentially it is possible to use one detector for all measurements, but this means that all measurement circuits must yield a d.c. output. For stability purposes it was thought desirable to have as few amplifiers in the measuring circuits as possible, and in fact only two have been used, one converts an r.f. signal level to an analogue d.c. level for the measurement of collector capacitance, base resistance and f_T . The other is used for measuring noise. All other a.c. circuits contain only rectifying and smoothing circuits.

The detector itself is a d.c. null detector and is sensitive to changes of less than $0.1 \mu\text{A}$ or 1 mV .

(3) During the 100 ms period for each test (except noise) two 10 ms periods are used for the two go/no-go tests, and four 10 ms periods are used for the para-

meter measurements as described earlier. The remaining time is used for allowing the circuits to reach steady state. The outputs of the circuits are sampled by a 1 ms pulse 7 ms after the commencement of each 10 ms test period. At connection of the circuit when d.c. transients occur 17 ms of settling-down time are allowed, but only 7 ms is allowed for the circuit to reach a steady state when the drive to the transistor is changed.

(4) There are two ways in which a null at the detector can be obtained. First, the transistor can be given a constant drive so that the circuit output is an analogue of the parameter to be measured (e.g. with constant base current, the collector current is an analogue of current gain). To determine the parameter therefore it is only necessary to "step" the reference voltage applied to the detector until a balance is achieved, then the parameter (e.g. current gain) is proportional to the reference voltage, which in turn is given by the setting of the circuit which "steps" the reference voltage.

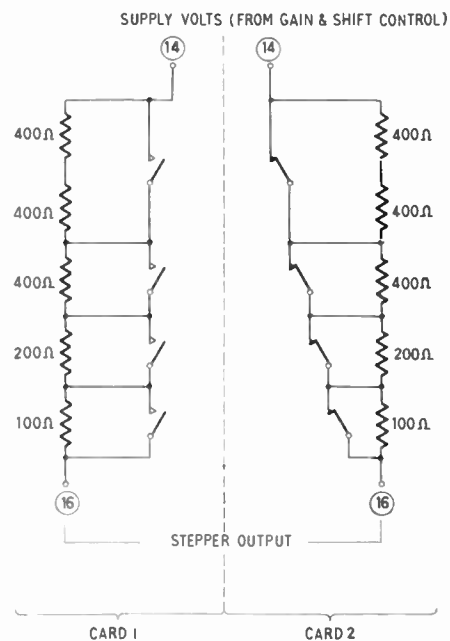
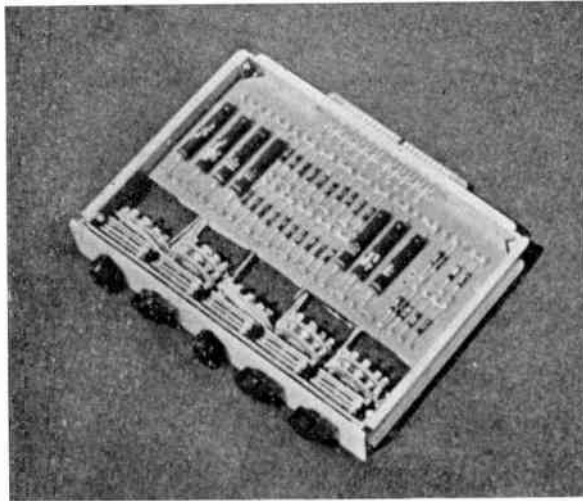


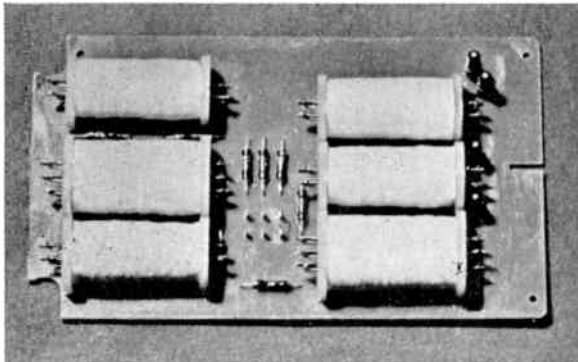
Fig. 1. Circuit diagram of the d.c. stepper.

Secondly, the reference voltage can be set to a particular value (e.g. the required collector emitter saturation voltage) and the drive to the transistor varied in steps until balance is achieved.

The circuits for achieving these stepping processes are called steppers. For all d.c. measurements and for circuits using the first described balancing process a common "d.c. stepper" or "A stepper" is used. Its circuit diagram is shown in Fig. 1. It is essentially a



(a)



(b)

Fig. 2. (a) A control unit. (b) A measuring circuit mounted on a printed circuit card showing in particular the reed relay bobbins.

quantized potentiometer operating between terminals 14. The output appears at terminals 16. The dry reed relays operate in pairs; when one is closed the other is opened. "Gain and shift" is achieved simply by altering the voltages applied to terminals 14. For other measurements, other types of steppers are used as described later.

In the go/no-go tests the steppers are not normally used. Instead an output is obtained from a control unit which replaces the stepper output. These outputs are called "batch boundaries" (since they are used in the transistor sorting process) and can be set up to any value, in the range available, by the operator. Each test circuit has an associated control unit on which are located the "batch boundaries" controls and the "gain and shift" controls. An illustration of a control unit is shown in Fig. 2(a).

(5) All the test circuits are mounted, together with their reed relays for connecting them, on printed circuit cards, an example of which is shown in Fig. 2(b). The steppers are on further adjacent cards. All the cards are arranged on two shelves; the a.c. circuits are located as near to the test heads as possible, but longer wiring runs are allowed for the d.c. circuits.

3.2. Examples of Test Circuits

3.2.1. Common emitter current gain h_{FE} at preset $V_{CE(sat)}$

There are two ways in which direct current gain can be measured at a preset value of collector-emitter voltage. If a constant base voltage is used and the base drive resistance R_B is varied, then, when a predetermined value of I_C is reached, it is easy to show h_{FE} is proportional to R_B . Hence by stepping R_B , h_{FE} can be measured.

Alternatively with a fixed R_B but variable generator voltage V_G , $I_B = V_G/R_B$

and
$$1/h_{FE} \propto V_G$$

In practice it is often of more interest to determine the spread of the base current to achieve a particular value of V_{CE} than the spread of h_{FE} . For this reason, therefore, the second method, which has the effect of giving smaller and smaller sub-ranges of h_{FE} as h_{FE} decreases, was chosen. This yields greater accuracy at lower values of h_{FE} , around which batch boundaries are normally set.

The circuit is shown in Fig. 3. V_{CE} is set using R_2 and R_3 , R_4 or R_5 and the d.c. stepper driving a fixed base resistor is used to provide the stepped base current and at balance $h_{FE} \propto 1/I_B$.

The machine contains three circuits similar to Fig. 3 each operating at a different collector current. The base current ranges for the 15 mA circuit are:

- (1) 50 μA steps from 50 to 800 μA

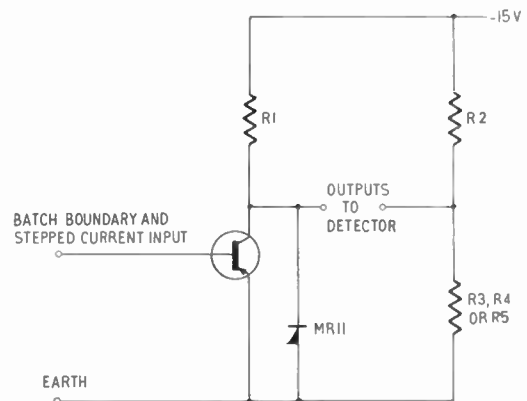


Fig. 3. The basic circuit for the measurement of h_{FE} at preset value of V_{CE} and I_C .

- (2) 25 μ A steps from 25 to 400
- (3) 25 μ A steps from 425 to 800.

The silicon diode MR11 is to prevent excessive dissipation in the transistor if it comes out of saturation.

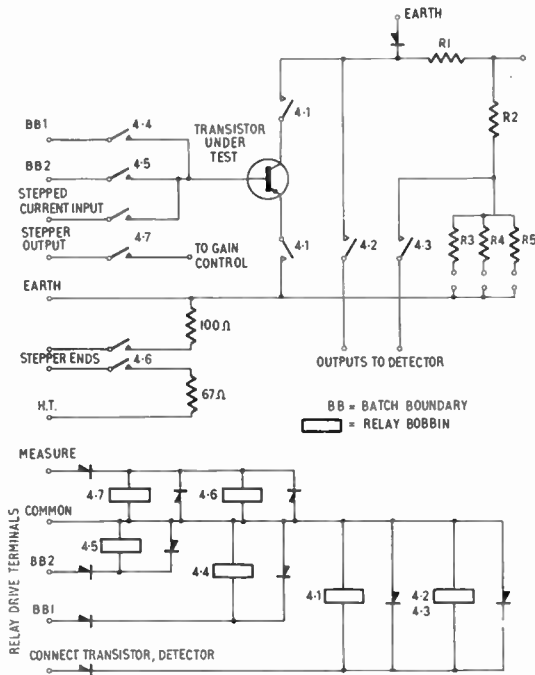


Fig. 4. The complete circuit for the measurement of h_{FE} .

The actual test circuit diagram including all the reed relays is shown in Fig. 4. Ten relays are involved and about the same number is required for each test circuit. Figure 5 shows the circuit diagram of the control unit.

SWA is the gain and shift control, and SWB, C and D, E are the two batch boundary controls.

3.2.2. Collector capacitance (c_{ob})

For all other circuits only the basic circuit diagram will be shown.

In determining this quantity particular care has to be taken to avoid the effects of stray capacitance and inductance. These strays are in general distributed so that a lumped circuit analysis does not lead to a complete solution; the residual errors are, however, sufficiently small. The principal strays are shown in Fig. 6 which also indicates the principle of measurement. If a constant 1 Mc/s signal voltage is applied to the collector of the transistor then, in the absence of strays, the base current is proportional to the collector capacitance. In the measurement, however, the input voltage is varied by means of the C stepper until the base current reaches a preset value, when c_{ob} is inversely proportional to this input voltage.

The stray capacitance C_1 from collector to ground is unimportant since the generator impedance is low. The collector-base capacitance C_2 produces an error current in the base lead which has to be allowed for. Since the circuit is a high impedance one, L_1 , the wiring inductance, can readily be kept insignificant. The inductance L_2 , which is the inductance of a strip of copper on a printed circuit, has to present an impedance small compared with $R_2 = 50 \Omega$. At 1 Mc/s this presents no difficulty. Similarly the effect of C_4 is negligible, its impedance must be large compared with the input impedance of the transistor.

The effect of C_3 is not so readily seen but it can be shown that

$$c_{ob} \text{ (measured)} \simeq c_{ob} \left(1 + \frac{(1 - \alpha)C_3}{c_{ob}} \right)$$

where α is the complex current gain of the transistor at 1 Mc/s.

For the lowest frequency transistor this gives about 20% error if $C_3 = c_{ob}$. In practice C_3 was kept to about 1 pF so that the theoretical error is not large, and it can in any case be reduced further in calibration.

The measurement is carried out by operating the switches in stepper C (see Fig. 6). The resistors labelled 16, 8, 4, 2, 1 comprise the measuring resistors. Gain and shift control is obtained in this circuit by selecting four of these resistors as follows. If 16, 8, 4, 2 are selected c_{ob} is measured in 2 pF steps. If 8, 4, 2, 1 are selected c_{ob} is measured in 1 pF steps. 16 can be permanently connected or disconnected to shift the range covered. Since the impedance of C_{IN} is large

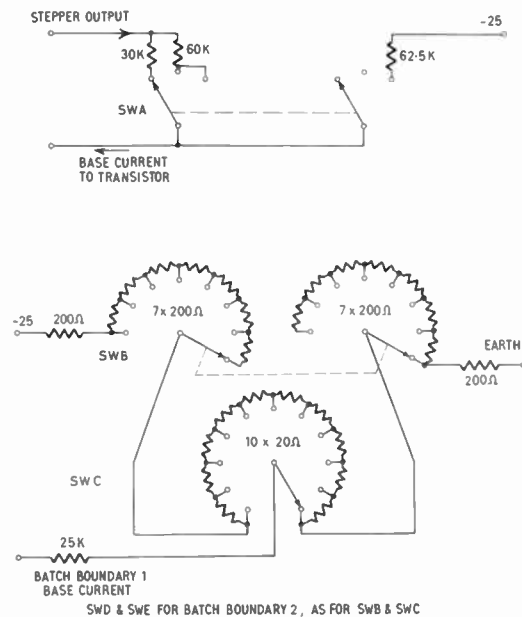


Fig. 5. The circuit diagram of the control unit for the measurement of h_{FE} .

compared with these resistors, the output voltage from the stepper is proportional to the net resistance connected by the switches. This means that c_{ob} is proportional to the sum of these conductances, so that at balance the setting of the switches associated with these resistors yields the value of c_{ob} . R_7 is about 40 Ω and sets the minimum value of c_{ob} measurable (5 pF); it also allows for any additive error in the measurement and in fact absorbs some error due to C_3 .

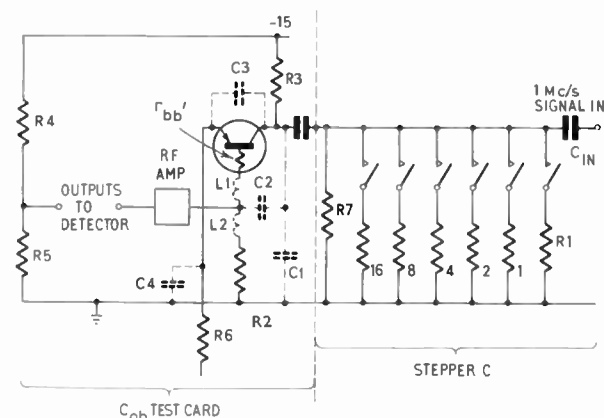


Fig. 6. The basic circuit for the measurement of collector capacitance. Stray components are shown dashed.

R_1 corrects for C_2 . The conductance of R_1 is set to correspond to the value of C_2 at 5 millimho per picofarad. The reason for it being a switched resistor is described in the next section.

The ranges of measurement provided are:

- (1) 5 pF to 20 pF in 1 pF steps
- (2) 20 pF to 35 pF in 1 pF steps
- (3) 5 pF to 30 pF in 2 pF steps.

In practice it is the earth currents which cause the greatest difficulty in the operation of this circuit. They arise from the large current circulating in the C stepper.

3.2.3. Base resistance ($r_{bb'}$)

Much the same circuit is used to measure base resistance as collector capacitance. The difference is that R_2 in Fig. 6 is shorted out and the r.f. amplifier is connected to the emitter instead of the base lead.

In the absence of strays, and with the same voltage applied to the collector as was used to measure c_{ob} , the emitter voltage is proportional to $r_{bb'}$, thus

$$\frac{V_e}{V_o} = \frac{r_{bb'}}{R_2}$$

where V_o is the output voltage obtained from R_2 when measuring collector capacitance.

R_5 and R_4 in Fig. 6 are replaced by stepper A, which then measures $r_{bb'}$ when a null is obtained on the detector. The important feature of the measurement is that the setting of stepper C, corresponding to c_{ob} , is retained, for a particular transistor, for use in measuring $r_{bb'}$.

The strays assume different degrees of importance now however. The wiring inductance L_1 is now in series only with $r_{bb'}$ (i.e. in a low impedance circuit), as viewed from the emitter, and must therefore be low. In order to achieve a sufficiently low value the base-resistance test-card is located immediately behind the transistor test head. (Fortunately this is the only circuit in which series base inductance is especially critical.) The inductance of the reed relay grounding the transistor contributes about 0.5 Ω to $r_{bb'}$.

In measuring c_{ob} the stray capacitance C_2 passed an error current in R_2 , but not in $r_{bb'}$. But it is essential for this measurement of $r_{bb'}$ that the currents in R_2 and $r_{bb'}$ should be the same. For this reason the resistor R_1 in the C stepper which is connected for the measurement of c_{ob} is disconnected for the measurement of $r_{bb'}$. This ensures the equality of the two currents in the two measurements.

The capacitor C_3 is the principal source of error in the measurement of base resistance. It can be shown that for the highest frequency transistors, a resistance is added to $r_{bb'}$ of approximately 25. C_3/C_{ob} , when the transistor is passing 1 mA of d.c. current. Thus if C_3 is 1 pF an error of typically 2.5 Ω results. This error can however be reduced in calibration by adjusting the voltage supplies to the A stepper for the purposes of measurement.

The ranges provided in the machine are:

- (1) 25 Ω to 400 Ω in 25 Ω steps
- (2) 25 Ω to 212.5 Ω in 12.5 Ω steps
- (3) 212.5 Ω to 400 Ω in 12.5 Ω steps.

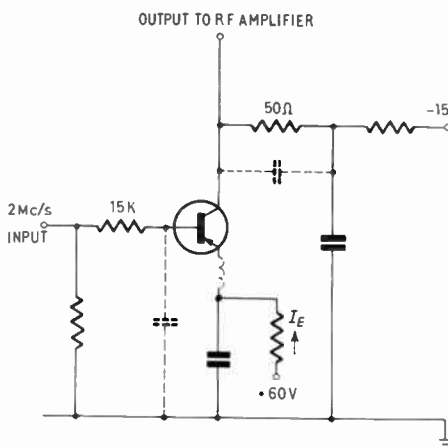


Fig. 7. The basic circuit for the measurement of h_{fe} at 2 Mc/s (or of f_T).

3.2.4. Gain bandwidth product f_T

The circuit for this measurement is shown in Fig. 7 together with the significant strays. The 2 Mc/s output from the circuit is approximately proportional to f_T so that measuring the output of the r.f. amplifier by means of the d.c. stepper enables f_T to be determined.

Unfortunately the output is not exactly proportional to f_T owing to the presence of the strays in the circuit as a result, primarily, of the long leads and the relays, and to the presence of collector capacitance. The principal sources of error in the external circuitry are the base to ground capacitance (about 40 pF) and inductance in the emitter, amounting to about 1 Ω impedance at 2 Mc/s. Figure 8 shows the measured f_T plotted against the true f_T for the two extreme values of collector capacitance encountered in the transistor

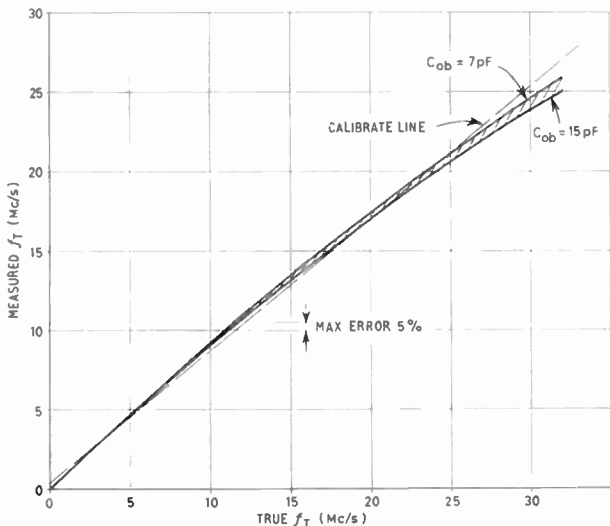


Fig. 8. The measured value of f_T compared with the true value. The dashed line shows the position of the calibrate line used to obtain minimum measuring error.

type to be measured by the machine—namely 7 pF to 15 pF. At 25 Mc/s this error amounts to about 20% low. However, by suitably adjusting the gain of the amplifier (or the levels of the d.c. stepper) the indicated values can be kept within $\pm 5\%$ of the true value. This is illustrated by the calibrate line in Fig. 8.

It is possible to improve matters even further by adjusting the circuit so that it measures correctly at two values of f_T which can be chosen to coincide with the two batch boundary values. Errors up to $\pm 5\%$ may be encountered at other values but these will not significantly alter the histogram patterns.

3.2.5. Turn-off charge Q_{OFF}

This measurement determines the charge required to turn-off abruptly a transistor after it has been turned

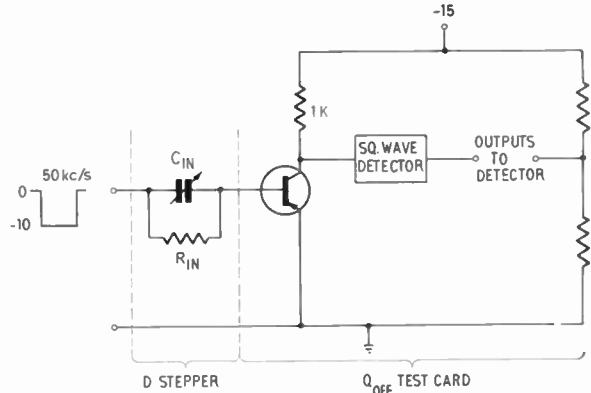


Fig. 9. Basic circuit for the measurement of turn-off charge, Q_{OFF} .

on to some specified steady state value. The measurement essentially involves changing the value of the variable capacitance in Fig. 9 until a square wave output is obtained. For this to be achieved in an automatic machine some form of square wave indicator is required. The method adopted here is to gate out the part of the square wave corresponding to the ON state, so that a pure square wave input to the gate will emerge as a constant d.c. level at the OFF voltage. Any departure from squareness at the turn-off edge is revealed as a change in the mean d.c. level.

This technique gives a very sensitive indication for values of charge less than the required amount, but it gives no indication if too much charge is being extracted at turn-off. To resolve this difficulty the d.c. level is shifted slightly to a point corresponding to "not-quite-off", so that a transistor driven fully off gives a change of sign of the d.c. output of the circuit. Of course this means that the circuit always reads low, but in practice the sensitivity of the circuit is such that the degree of off-set can be made a small fraction of the smallest measurement sub-range, so that the measurement error

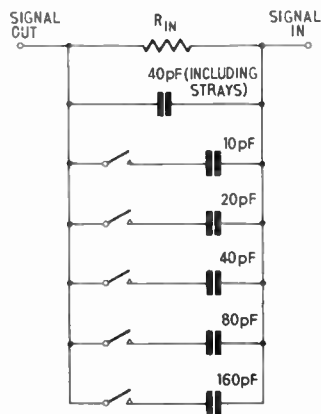


Fig. 10. The basic circuit diagram of the D stepper used in the measurement of Q_{OFF} .

is normally negligible. Poor transistors such, for example, as those with low current gain or high cut-off current can give rise to erroneous values of Q_{OFF} , but again the limits are such as normally lead to the rejection of the transistor in any case.

For this measurement the stepper used comprises parallel capacitors of values proportional to 8, 4, 2, 1 connected by switched reed relays as shown in Fig. 10. It takes the place of the variable capacitance in Fig. 9. R_{IN} sets the ON base current.

3.2.6. Low frequency current gain h_{fe}

All the circuits described so far reach a steady state well within the allotted 7 ms period. Meeting this requirement was the principal difficulty in the measurement of small signal current gain at a low frequency; at 1000 c/s only seven cycles have taken place in the available time. For this reason a frequency of 5000 c/s was used.

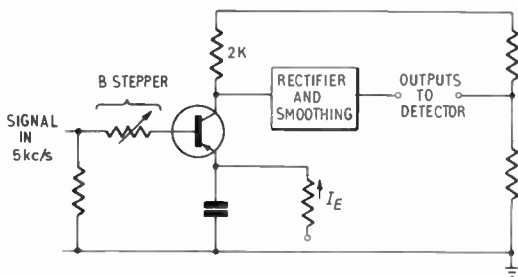


Fig. 11. The basic circuit for the measurement of h_{fe} at 5 kc/s.

The measurement circuit was as shown in Fig. 11; the B stepper is shown in Fig. 12. In measurement the B stepper is switched until a preset value of output signal level is reached; at which time h_{fe} is proportional to the value of resistance in the stepper.

The component values in the circuit represent a compromise between on the one hand obtaining a value of h_{fe} with negligible percentage error and on the other hand giving adequate sensitivity, adequate smoothing and small enough time constants to reach the steady state. In practice this means that the emitter decoupling capacitor and the load resistance present impedances greater than desirable, but since the errors arising from both decrease as h_{fe} decreases the percentage error can be kept well within 5%. The principal difficulty arises from the smoothing circuit which must reduce the ripple to negligible proportions and yet reach a steady value. At 5 kc/s this was found to be just possible but, of course, 5 kc/s is too high a frequency for measuring low frequency h_{fe} for those transistors which have high h_{fe} and low values of f_T . Percentage wise, however, the accuracy remains within 5% over the measurable range of h_{fe} for all transistors whose $f_T > 3$ Mc/s.

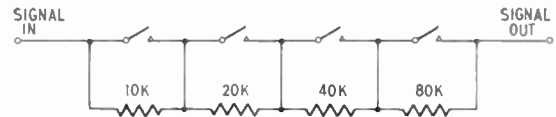


Fig. 12. The B stepper used in the measurement of h_{fe} at 5 kc/s.

3.2.7. Noise figure

In essence this measurement consists of driving the transistor to be measured (which is connected in the common base configuration) by a wire wound 500 ohm resistor, and integrating the noise output voltage at 1 kc/s over a period of 0.9 seconds. The noise output is amplified in several stages, rectified and used to charge of 1 μ F capacitor. The potential across this capacitor after 0.9 seconds is measured (after it has been suitably buffered), by the A stepper, whose setting at balance therefore yields an indication of noise figure. The amplifier incorporates limiting diodes which (a) give the input/output characteristic a logarithmic shape so that the output can be calibrated directly in decibels, and (b) substantially reduce the effects of voltage transients.

The principal source of error in this measurement arises from these voltage transients. They arise from the fact that the noise test head is close to the other test head on which all the other measurements are being made, and during the 0.9 seconds of the noise measurement, voltages are being switched on and off at this other test head. Unavoidable capacitive coupling, of the order of 1 pF, between the two test stations transmits these transients to the noise amplifier and causes some degree of ringing at 1 kc/s. By various techniques, including that of incorporating limiting diodes, the effect of these transients has been reduced to a level such that they contribute an error

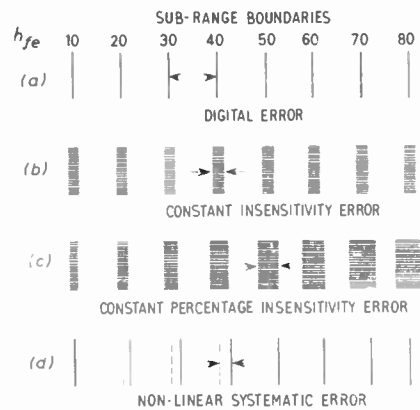


Fig. 13. Sources of error. (a) The subranges of measurement. (b) Constant insensitivity error. (c) Constant percentage insensitivity error. (d) Systematic error.

noise of 2 dB, and this is allowed for in the calibration of the measurement.

An accuracy of ± 2 dB is obtained with this circuit over the range 0 to 30 dB.

4. Accuracy and Sensitivity

The error arising in any of the measurement circuits falls into three categories, and these are illustrated in Fig. 13.

(1) *Digital error.* The "digital error" describes the source of uncertainty in the determination of a parameter arising solely from the fact that the measurement results are quantized. The magnitude of this uncertainty can readily be found from the size of the sub-range and the magnitude of the parameter in the sub-range involved. For example at point X in Fig. 13(a) it is $\pm 500/45\%$. Expressed as percentage of full scale it is $\pm 500/160\%$.

This is not a source of error in the go/no-go tests, the only source of error in these tests arises from a lack of precision in locating the batch boundaries. There are two sources of error here which, of course, also apply to the measurement process as well, since a sub-range boundary is much the same from the measurement point of view as a batch boundary, and will now be discussed.

(2) *Insensitivity error.* Any lack of sensitivity of the detector causes a blurring of the batch or sub-range boundaries and may be called the "insensitivity error". The sensitivity of the detector, as has been stated, is $0.1 \mu\text{A}$ or 1 mV , but the error this can cause in the determination of a particular transistor parameter varies from circuit to circuit. In general it is a small fraction of the smallest digital error; for example, in determining f_T , the r.f. amplifier yields a d.c. output of 1 volt for an f_T value of about 20 Mc/s, so that the uncertainty in f_T with a detector sensitive to 1 mV is 20 kc/s, or 1% of an f_T value of 2 Mc/s.

The two different measurement techniques yield different kinds of insensitivity error. Those circuits in which the drive to the transistor is stepped yield a constant percentage error, as illustrated in Fig. 13(c). Those in which the reference is stepped yield a constant error, as indicated in Fig. 13(b), so that the percentage error increases as the parameter value decreases. The measurement of f_T is an example of this kind of error. The measurement of h_{fe} at low frequency is an example of the first kind.

In practice the circuits for the measurement of h_{fe} and noise are the worst from the sensitivity point of view. For h_{fe} the constant percentage error is of the order of $\pm 4\%$. This means that in measuring low values of h_{fe} , the insensitivity error is a small fraction of the digital error, but at high values of h_{fe} the insensitivity error becomes as large as a measurement

sub-range and a transistor whose h_{fe} value is, say, 135 will nearly always appear in the sub-range 130 to 140, but will occasionally appear in the sub-ranges 120–130 or 140 to 150.

All the other circuits have markedly better performance than this one, and only those parameter values which are close to a sub-range boundary suffer from some measurement uncertainty. For example, a collector capacitance of 10 pF will appear either in the sub-range 9–10 pF or 10–11 pF. One of 10.1 pF will occasionally appear in the 9–10 pF sub-range, but one of 10.2 pF will always appear in the 10–11 pF sub-range.

(3) *Systematic error.* All the a.c. test circuits possess some systematic error as a result of circuit strays, etc. These have been dealt with in detail in the previous sections, and lead to the kind of error illustrated in Fig. 13(d). But in the circuits where this error arises it is possible, by adjusting the stepped reference levels, or the drive circuits to the transistor, to find the best linear fit of the measured value to the true value. The worst two circuits in this respect are f_T and low frequency h_{fe} , since the true value is not a constant factor times the measured value.

Under the heading of systematic error also come any inaccuracies in the reference levels, or in the other supply levels. To keep these errors to a minimum all resistors are of high stability and 1% tolerance. All d.c. and a.c. supplies are maintained within an accuracy of $\pm 1\%$ of their nominal values. The r.f. and noise amplifiers maintain comparable gain stability. In this way the accuracy of the circuits is maintained as constant as possible.

As a further check, however, when the machine is running, a "standard" transistor is measured every 25 transistors and any departure from the expected result stops the machine. And further, a magazine containing simulated transistors consisting of passive elements can be used to check most of the circuits.

In practice it has been found that the machine is very rarely put out of commission as a result of inaccurate measurement.

5. Conclusion

It has been found possible with an automatic machine to measure all the significant parameters of alloy transistors (except low frequency h_{fe} and noise) with an inherent accuracy of better than $\pm 3\%$ over most of the measurement range. The principal uncertainty in any measurement has arisen in practice from the quantized nature of the measurement and at low values of parameters the size of the quanta become a fairly large fraction of the measured quantity. But this could always be improved by using smaller steps in the digitizer.

It is felt, however, that any attempt to adapt the single test station philosophy to the measurement of higher frequency transistors is unlikely to succeed owing to the magnitude of the circuit stray capacitances which inevitably arise when many measurement circuits are connected in turn to one test station.

6. Acknowledgments

The author would like in particular to thank Mr. M. R. A. Paesler for his assistance in the engineering of all the circuits described in this paper and for designing and engineering the most difficult of them

all, the circuit for measuring noise. In addition thanks are due to Dr. L. J. Herbst for designing the r.f. amplifier, and to Mr. J. Lawton, Director of Research of British Telecommunications Research Ltd., and Dr. P. Tipple of Associated Transistors Ltd. (for whom the Automatic Transistor Tester was built) for permission to publish this paper.

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POINTS FROM THE DISCUSSION

Mr. M. Stoughton: More and more transistors nowadays have I_{cbo} figures substantially less than one nano amp. h_{FE} is also often measured at 10 μ A collector current. Can the existing machine measure such low currents and to what accuracy?

In life testing there is evidence that mechanical handling of the transistor gives rise to an appreciable number of faults; it seems possible that the introduction of this machine with its capabilities of comparison of present readings with previous readings will show much worse life figures than a machine comparing figures whilst leaving the transistor in a life-test rack without interference.

The author (in reply): The current measuring capability of the transistor tester is limited by the sensitivity of the detector, which is about 0.1 μ A. This means that I_{CBO} of silicon transistors could not normally be measured although h_{FE} at $I_C = 10 \mu$ A could be measured to an accuracy approaching 1%.

Using the automatic tester for life testing should not interfere with the transistors any more than many other test equipments. For life-test purposes the transistors are not removed from the magazines, indeed they are soldered into them, and the only handling the magazines receive is removing them from the ovens from time to time, as with the majority of life-test programmes. The tester cannot however be used to test transistors in their life-testing ovens.

Mr. H. C. Kearton (Associate Member): In view of the adverse comments on noise in reed relay circuits, has the

author experienced difficulty in the measurement of transistor noise (such as thermal noise, contact resistance, etc.) with these devices?

The author (in reply): The noise generated by the relays themselves was much less than other sources. The principal source of spurious noise was the presence close to the noise test station of the other test station at which voltages were being switched on and off quite rapidly. The main difficulty in designing the noise measuring equipment was the elimination of this spurious noise.

Mr. L. Nelson-Jones (Associate Member): Could the author say in the light of the comments made on reliability required of testing equipment, what are the reasons for the apparent large use made of cracked carbon high stability resistors which, at any rate in my experience, can be anything but highly stable.

The author (in reply): If the automatic tester were to be built again today, two modifications would be made to improve the stability of the testing equipment. Firstly, many fewer resistors would be used; the control units in which most of the resistors are located would be designed to contain only diodes for the matrix selection of common resistors (this is already done in some of the a.c. circuits which were designed later than the d.c. ones), and secondly, metal film resistors would be used. The excuse for using cracked carbon resistors is that at the time the tester was built metal film resistors were not readily obtainable, nor were we as aware of the instability of cracked carbon types as we are now.

Data Collection Systems, their Application and Design

By

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Presented at the Symposium on "Data Transmission" in London on 3rd January 1962.

Summary: A distinction is drawn between two types of data gathering systems. Possible applications for one of these, namely data collection systems, are discussed. The requirements of such systems are then outlined and two types of data collection system are described to indicate possible solutions to these requirements. The final section deals with the total systems approach and the effect it will have on future data collection systems.

1. Introduction

There has been considerable discussion on the need for the automation of production processes but progress in this area has been relatively slow. A similar need for automation exists in handling and processing data and it is probable that progress in this area in the next decade will be more spectacular than in the production area. In an overall system two main areas, namely data communications and data processing, can be defined. A great deal of progress has been made in data processing but it is only recently that data communication has aroused interest. This paper is not concerned with the high-speed point-to-point transmission of data but rather with a particular class of multiple-input low-speed data gathering systems. A combination of these two types of system would frequently be required to satisfy an organization's data communication requirements.

The description of two data collection systems is preceded by a discussion of possible applications and of the salient features that data collection systems should have in order to satisfy the requirements of the various applications.

2. Data Gathering Systems

A data gathering system is required whenever data which exist at one point are needed for processing at another point. The basic requirements of such a system are:

- (a) Data must be intelligible to the receiver.
- (b) Data must arrive when the receiver needs them.
- (c) Data must be transmitted accurately.

Data gathering systems may be divided into two classes. In Class 1 the flow of data is controlled by the data source or input station and in Class 2 it is controlled by the data destination or receiving station.

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This implies that in Class 1 systems the input of data can be on a random basis whereas in Class 2 systems the input data must be continuously available. In this paper systems of the first class are referred to as data collection systems and those of the second class as data acquisition systems. Data acquisition systems are normally installed where the input data are mainly derived from analogue instruments such as thermocouples which give a continuous indication so that data can be output at any time.

In data acquisition systems data are usually transmitted under the control of a central scanner in analogue form. The input points are scanned sequentially or in random order under computer control. The central controller is able to convert the analogue data to coded digital pulses for output in any required form and it can also detect alarm conditions.

The definition of two classes of data gathering system is made largely because there are at present two classes of equipment available each with a different type of control system.

3. Data Collection Systems

If a central organization is to control a number of inter-related and inter-dependent events or activities it must collect data from the points where these events or activities occur. Modern progressive management is interested in obtaining a much higher degree of control over all aspects of its organization. The digital computer offers management a most effective tool for decision making because of its ability to accept and process large volumes of data at high speed. The enormous increase in processing speed that the computer gives over manual methods means that the arguments used in the decision-making phases of any operation can be far more sophisticated and the frequency of the decision making may be increased. If a computer is to be used effectively as a means of providing better management control it is probable

that it will require more data more often and more quickly than the manual system which it replaces.

The fact that the input data requirements are more stringent can in itself make an organization more efficient by introducing a stricter discipline.

3.1. Conventional Methods of Data Collection

The inadequacies of the conventional methods are most obvious in the areas of data conversion and data transmission.

Conventional collection systems are normally based on written messages and several stages of transcription may occur before the final message is produced. At each stage, time is wasted and errors may be introduced. If the data are required for computer processing the written messages must be converted to an acceptable computer medium, and this again is a time consuming and expensive operation although the probability of error introduction at this stage is low.

3.2. Application Areas

Consider Fig. 1 which outlines the data transmission requirements of a typical manufacturing and selling organization with a central unit controlling several geographically separate production units, warehouses and sales offices. This paper is not concerned with the high-speed point-to-point data transmission systems that would be required between the central controller and the production plants, warehouses and sales offices. Data collection systems will be required in each production unit to permit precise control of the complete production process from receipt of raw materials to the storing and shipping of the finished articles. They will also be required to enable the sales force to inform the sales offices of orders received.

These two applications areas lead to the definition of two types of data collection systems, namely in-plant systems and out-plant systems. Precise definitions

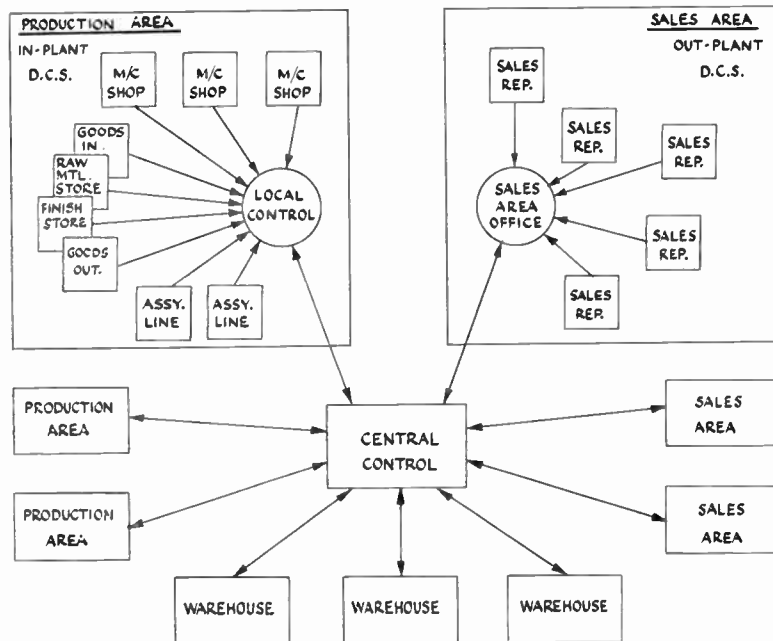


Fig. 1. Typical data transmission network.

The human and mechanical carriers used in the conventional methods of data collection are often inadequate for computer based data processing systems because of the inherent time delay between the event and the reporting of that event to the processing centre. Ideal control can only be exercised if the central computer is informed immediately an event occurs. However, for most applications data transmission speeds, from a collecting point to the controller, in the range 10 to 100 characters per second are adequate providing data input is random.

are neither possible nor useful but the division of the type of application can be related broadly to the data collection systems currently available.

3.2.1. In-plant systems

This type of system is entirely located on property owned by the organization installing it. Figure 1 shows a typical in-plant system for installation in a production plant. Collected data will be transmitted via a point-to-point data transmission system to the central controller to enable it to plan and schedule

work, re-schedule quickly and compute efficiencies and incentives.

3.2.2. Out-plant systems

An out-plant data collection system may be defined as a system for collecting data which involves the co-operation of a third party for the transmission of data. (In this country the third party will normally be the G.P.O.) Figure 1 indicates a possible application area for an out-plant data collection system where salesmen must return details of orders taken to their area office at the end of each day to achieve a quick turn-round between receipt of order and delivery of the goods. Such a system is required when the customer cannot or will not keep stocks and where there is a highly competitive sales situation between suppliers so that failure to deliver promptly could involve loss of business to a more efficient competitor.

4. System Requirements

Before discussing in detail particular data collection systems it is useful to consider the various types of input and output devices that such systems require and the method of transmission that can usefully be employed.

4.1. Input Devices

Input devices are normally based on punched card or punched paper tape readers and keyboards for the input of static and variable data respectively. All input devices should fulfil two basic requirements. They should be inexpensive, because in tightly controlled organizations they will be required in large numbers and they should require the minimum of maintenance, as they are likely to be placed at remote points and required to operate for 24 hours each day. The design of the devices is largely based on the type of application for which they will be used, i.e. in-plant or out-plant.

4.1.1. In-plant devices

These must be rugged and very simple to operate as they are to be used by unskilled personnel on shop floors, in stores, in goods receiving areas, etc. Indicators should be provided to guide the operator and to inform him of erroneous operation. Whenever possible the input of variable data such as quantity made, weight, length, etc., should be automatic to avoid errors. If this is not possible the operator should be given a visual indication of the variable data he intends to transmit so that he may correct it if necessary.

4.1.2. Out-plant devices

The requirements for reliability is even more essential with out-plant devices as they are likely to

be dispersed over a wide area that would make frequent maintenance calls impracticable. These devices need not be so rugged as the in-plant ones but portability would be a useful property in many applications. With this in mind the need for mains supply to the input device should be avoided.

4.2. Output Devices

Data collection systems will be either on-line or off-line to the central processor. On-line systems will not normally require any output and the data collection systems will be connected to the central processor via a multiplexer, a message concentrator or a message exchange. Off-line systems should produce output data in a form suitable for computer input at a later stage.

4.3. Transmission

For in-plant systems many methods of transmission may be used and the method chosen will normally be based on an economic compromise between cabling costs and terminal costs. Out-plant systems must use the transmission facilities offered by the third party carrier and these will normally be limited to telegraph or telephone lines.

The checking employed in an in-plant system need not be elaborate as the transmission paths are short and not subject to much interference. These conditions do not apply to out-plant systems and some form of error detection should be employed. The actual method will depend on the third party facility to be used and the transmission speed but will normally check the validity of each character received. Message length checking is essential in either type of system, in-plant or out-plant, to ensure that the correct number of variable characters has been input.

4.4. Systems Control

In systems employing multiple input stations some form of control must be included to allow only one station to transmit at any one time. Where direct input to a central processor is provided this facility may not be required.

It is necessary to provide circuitry to allow input stations to contact the output station or central processor when they have data to transmit. Circuitry must also be provided to request re-transmission when an error condition is sensed.

5. An In-Plant Data Collection System

In this section the IBM 357 Data Collection System is described to indicate one solution to the requirements of an in-plant data collection system. The system permits transmission of data from a number of input stations to centrally located card punches and it is therefore an off-line system.

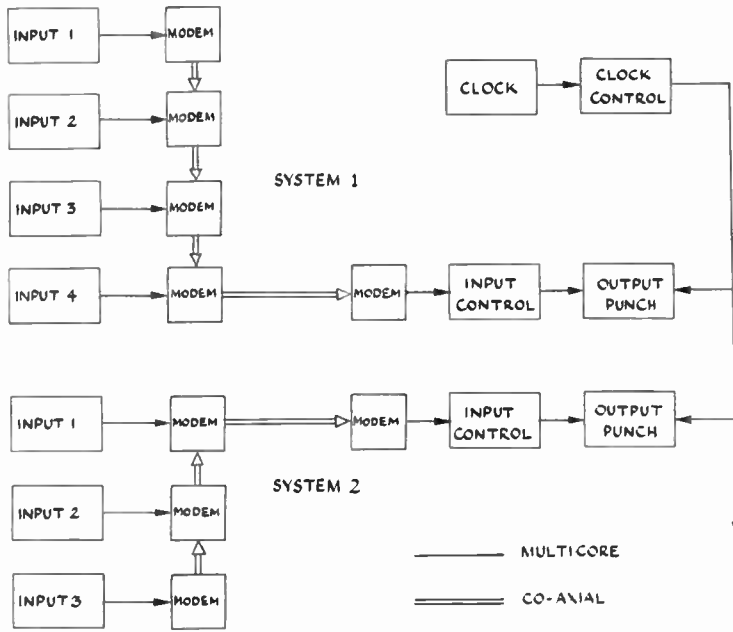


Fig. 2. In-plant data collection system (IBM 357).

5.1. System Layout

Two systems are shown in Fig. 2, one with four input stations and the other with three input stations. A maximum of 20 input stations may be connected to any one output punch via an input control unit. The clock and its associated control may be connected to up to 35 output punches. Since this is an in-plant system there are no restrictions on the type of transmission that may be used. Figure 2 indicated that two methods of transmission are employed, for short links, d.c. signalling on a multi-wire system, and for the longer links, a voice-frequency system using 12 frequencies in the range 640–8490 cycles per second is

employed over standard 75 ohms coaxial cable which is customer installed. The multi-frequency system requires modulator/demodulator units at each end but the coaxial cable is much cheaper than the multi-core, used on the short links, and far easier to install. The distance between the farthest input station and the input control unit must not exceed 5000 metres (3.108 miles).

5.2. Information Flow

The flow of information in the 357 system is outlined in Fig. 3. All input stations are connected to a common bus which terminates on the input control

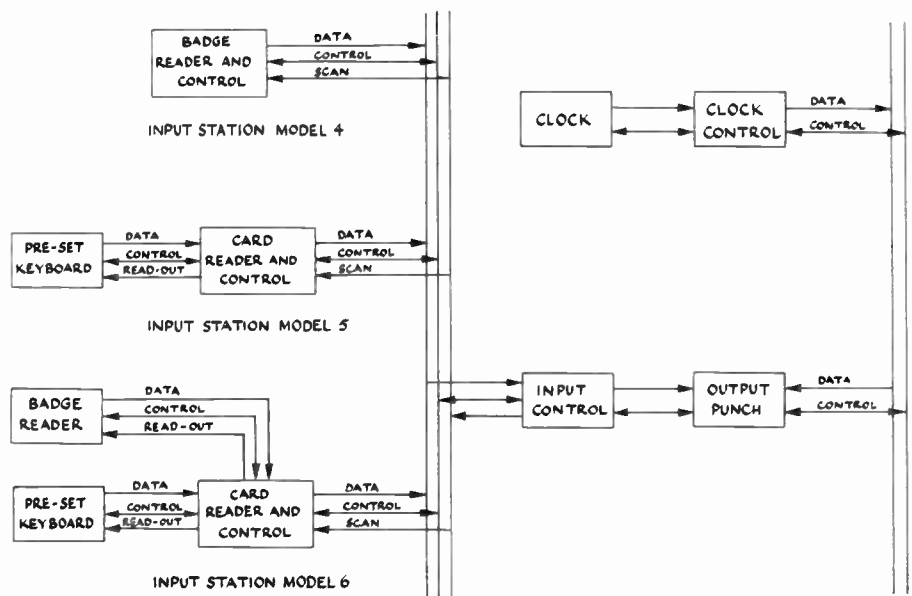


Fig. 3. Information flow in system shown in Fig. 2.

unit connected to the output punch. Data flow is simplex while the control flow is half-duplex.

5.2.1. Input station

The input station accepts data from punched cards (80 alpha-numeric characters), punched plastic badges (10 digits) and from a keyboard (12 digits). The three models are shown in Fig. 3. The keyboards are optional for models 5 and 6.

Each input station consists of the appropriate reading device, brush-type for cards, pin sensing for badges, a power switch, control lights and an auxiliary power supply. Card reading is not a continuous motion. The card, once started, is stopped by the sensing of a punched hole. The punched information is read and transmitted and the card movement is re-started. The card reading rate is 20 columns/second and the skipping rate is 60 columns/second. When the badge is inserted it is locked in position, read out in any desired column order and then released for removal.

The keyboard consists of 12 sliding columns which may be pre-set, and hence corrected when necessary, before being read-out. Keyboard read-out is controlled by code punches in the input card and it may be done in any sequence.

Each input station is assigned a three-digit station number. This number can be read-out on demand and the units position must be assigned on a patch panel in the input station.

Control lights on the input station indicate the status of the station, and instruct the operator to insert or remove cards or badges and to set up data on the keyboard.

5.2.2. Input control unit

This unit controls the transmission of data from input stations to the output punch. It is composed of a transmit control circuit, a scan-and-select circuit and an auxiliary power supply.

Instruction codes, punched in input cards or emitted from a program device on the badge reader (model 4) are analysed by the transmit control unit and the corresponding functions are performed. There is a patch panel to provide flexibility in the assignment of instruction codes to operational functions. There are four selector levels available on the patch panel and these are controlled by single digit transaction codes which must be transmitted at the beginning of each message. The transmit control unit also controls the input station indicator lights and upon completion of a transaction it controls the disconnection of the circuit to the input station and renewal of the scan-and-select function.

A ready light on the input station indicates that a card or a badge may be inserted for processing. When this is done the scan-and-select circuit is activated. The input stations are then scanned sequentially until one with data to transmit is sensed. This station is selected and the read operation initiated. The scanner takes 35 ms to move from one station to the next.

The tens and hundreds positions of the station numbers must be assigned on the station number read-out panel in the input control unit.

5.2.3. Output punch

The card punch is modified by a receive control feature, which may be used for manual operation when not required by the 357 system. The output punch is controlled from a punched program card fixed to a drum which rotates at the same rate as each data card being punched. There are 12 contacts down the edge of the drum and these may be assigned to any of the 10 output punch functions by means of a patch panel in the receive control feature. This panel has four levels of selection and the particular level is selected at the same time as that in the input control unit. The lengths of messages for different transactions will vary considerably and it is possible to punch several messages in one card or to use several cards for one message.

5.2.4. Output clock and control unit

A synchronous clock emits time in hours, tenths and hundredths of an hour within each 24-hour period for punching time in output cards. The clock has a control unit which contains a clock selector, circuits to prevent clock read-out during clock advance and an auxiliary power supply.

The clock selector is used to control output punch functions during specific clocking periods. A patch panel in the clock control unit permits the specification of any four periods in a 24-hour period. The starts and stops of these periods must be designated in multiples of 6 minute increments.

Clock read-out for any particular transaction is controlled by punched codes in the output punch program card.

5.2.5. Checking facilities

The main check on transmitted data is on the pre-determined message length. If an end of transaction code from the input station does not coincide with a column check programmed at the output punch an error light, indicating too few or too many characters in the message, is turned on at the input station. If this check is performed satisfactorily, the output card is punched in column 81.

Detection of a blank column during read-out of the keyboard stops transmission and causes the

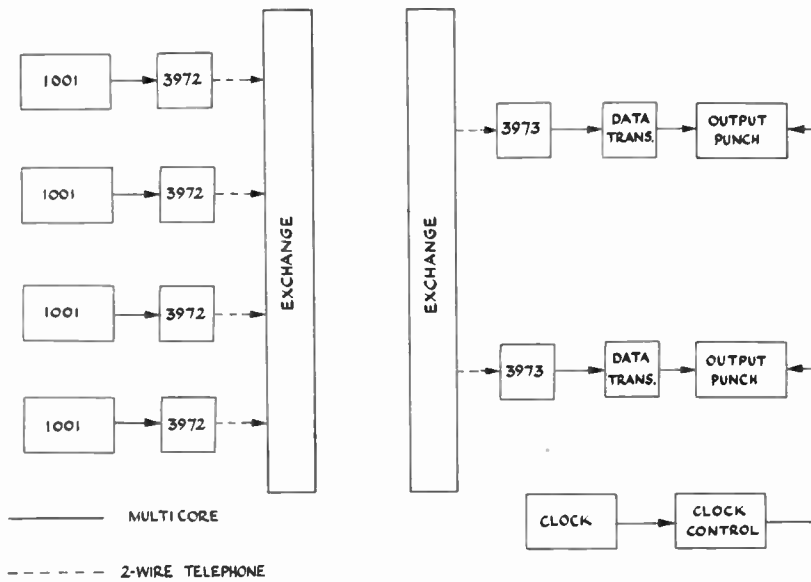


Fig. 4. Out-plant data collection system (IBM 1001).

manual-entry light to glow. The operator has 15 seconds in which to enter the missing digit and press start to resume transmission.

A light and buzzer are turned on at the punch if a malfunction of this unit occurs. If for any reason the individual operations at the input station are separated by more than 15 seconds, an error light is turned on at the input station and the transaction must be repeated.

6. An Out-Plant Data Collection System

The IBM 1001 system has been developed specifically for out-plant operation. It is based on the Bell Data-Phone transmission system and is composed of one or more input stations from which data may be transmitted over telephone lines to a central output punch. Speech transmission in either direction is also provided.

6.1. Systems Layout

Figure 4 indicates the type of system that can be achieved; any number of input stations can contact any number of output punches using a telephone network. In practice, systems usually consist of 6 inputs feeding one output punch. An output clock may be connected for punching time. The Bell Data-Phone 400 system has been designed for transmitting data over the switched network at low speed (i.e. up to 20 characters/second). It will operate over normal speech lines as long as the overall attenuation does not exceed 30 dB.

Connection between an input station and the central punch is established by dialling the number of the punch on the telephone at the input station. If the punch is being used the engaged tone is heard.

6.2. Information Flow

Figure 5 shows the flow of information in the system. Data flow is simplex while control flow is half-duplex, as is the case with the in-plant system already described.

6.2.1. Input station

The input station consists of two units, the input unit and the Data-Phone modulator.

The input unit contains a card reader (first 22 columns, numeric only), a ten-digit keyboard, functional keys and a monitor speaker. It weighs approximately 12 lb and is connected to the modulator by a multi-core cable. The function of the input unit is best considered after a brief description of the modulator.

The modulator unit is a subset installed in place of an ordinary telephone subset. In addition to the normal telephone service it provides a means of transmitting numeric data and of receiving control signals acknowledging receipt of this data. Contact closures in an external unit are used to select two audio frequency signals from eight frequencies generated in the modulator by two transistor oscillators. Each oscillator provides a rest frequency and four higher signal frequencies, one oscillator operates in the range 600 to 941 c/s and the other in the range 1100 to 1633 c/s. Closure of external contacts must select one frequency from each oscillator thus allowing 16 frequency combinations. The audio tones generated are transmitted to the line in parallel and to the external unit for monitoring purposes. When the external contacts are opened the oscillators transmit their rest frequencies for a period of 50 to 100 ms to reduce the effects of echo signals at the receiving terminal.

The modulator unit also contains a two-stage audio amplifier to drive a receiver in the external equipment. Control signals sent from the output station on the receipt of a message can thus be amplified and fed to the external unit for interpretation by the transmitting operator.

Power for the oscillators and amplifier in the modulator unit is derived from the central office battery via the subscriber's line. A bridge rectifier provides output whose polarity is independent of line connection and a surge protection diode is included. From the foregoing description it is obvious that the input unit must contain a switching device to select the signal frequencies generated in the modulator unit and a speaker for monitoring these signals and receiving control signals generated at the output punch. Two switching devices are provided, one is a 15-position keyboard, giving digits 0 to 9 and 5 functional keys. The functional keys control the start and finish transmission, the rejection of known error cards and the release of cards at the output punch. The other switching device is a card reader which consists of a spring-loaded carriage for the card and ten sensing pins which make the appropriate contact closures as the card passes under them. The carriage is manually loaded and when released is read at approximately 12 columns per second.

The monitor speaker used in the input unit is a standard telephone receiver. Transmitted tones can be heard and a continuous tone at the end of a message indicates correct transmission and an intermittent tone an incorrect transmission.

6.2.2. Output station

The output station consists of two units, a modified card punch and a Data-Phone 401 B demodulator. The latter is a subset installed in place of an ordinary telephone subset. In addition to the normal telephone service it provides a means of receiving data signals from the transmitter and converting these signals to an output suitable for operation of a receiving business machine. The demodulator can deal with 3-out-of-12 coding, which allows 64 combinations, and will therefore accept alpha-numeric data if required. It will operate in either the attended or unattended mode and it also generates control signals for transmission to the input station. In the demodulator data signals are fed to an a.g.c. amplifier and then to a cut-apart filter. This filter divides the signal into three frequency bands, only two of which are used in this system. Individual frequencies in each group are detected by anti-resonant circuits which control relays, the contacts of which are available for the operation of the external business machine.

The demodulator also contains an answer-back oscillator for generating a control tone. The power supply is derived from the mains.

The output punch is connected to the demodulator by a data translator unit which converts the 2-out-of-8 code to IBM card code and performs control and checking functions. Data output is controlled by a program card contained in the output punch.

6.2.3. Checking facilities

Two forms of checking are provided in the out-plant

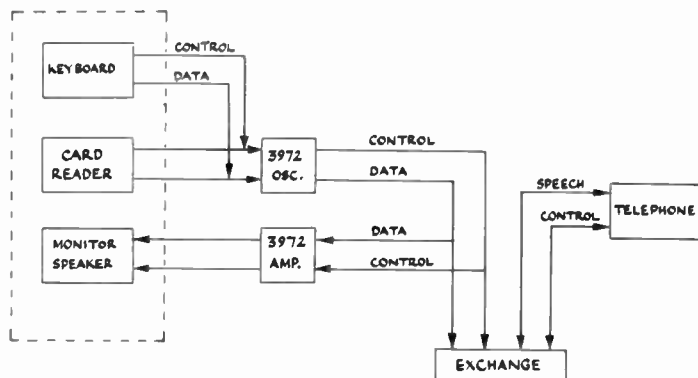
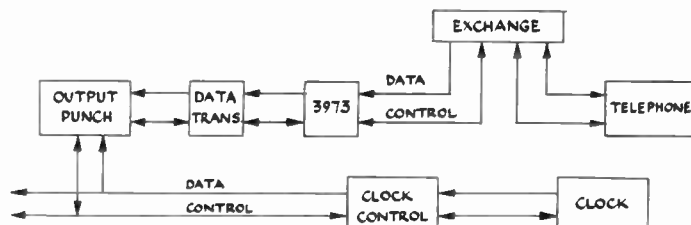


Fig. 5. Information flow in system shown in Fig. 4.



system. Firstly the data translator checks that only two valid tones are received in parallel and secondly the length of the received message is compared with a pre-determined length indicated on the program card. If either of these checks show error the sending operator is informed by an intermittent answer-back tone and the output card is not punched in column 81 so that it may be outsourced. If at the end of a message the checks are satisfied the sending operator hears a continuous answer-back tone.

An optional feature to check the validity of certain data input via the keyboard may be added to the output punch. This is called a self checking number feature and it performs a check sum on the meaningful data and compares the result with an extra digit which must be keyed with the data.

7. Total Systems

The main reason for the lack of widespread acceptance of data collection systems of the types described in Sections 5 and 6 is a financial one. Potential users of these systems consider them in isolation and attempt to set possible savings in labour against the equipment and installation cost, with the result that economic justifications can rarely be made.

Data collection systems must be considered as essential elements of much larger and comprehensive systems for which economic justification is quite feasible.

7.1. Tele-Processing Systems

The large systems referred to are known as total or integrated systems and their development requires an entirely new approach to an organization's activity and is only achieved by an exhaustive study of all aspects of that organization. Integration in this context refers to the total integration of functions and not to the limited integration of jobs within particular functions that some companies have achieved.

The term Tele-Processing is used to describe studies, systems and equipment associated with the total systems approach to an organization's operations. Tele-Processing is used to indicate that the proper solution of such an approach requires communication-based data processors. Most of the computing systems currently available are unsuitable as solutions to a total systems approach. The same is true for the collection systems which feed these computing systems and the requirements of future data collection systems are discussed in this context in the next section.

7.2. Future Data Collection Systems

Most data collection systems work off-line. This means that cards are punched with collected data and

have then to be placed in a card reader to input the data to a computing system. It is not practicable to feed each card to the computer as it is punched so a form of batch processing has to be employed. When the philosophy of the total systems approach has been accepted it will be necessary to engineer a direct connection between the data collection system and the computing system. This will allow on-line data collection and flow-processing of the data.

An expensive central processor must not waste time collecting wrong data; therefore the input of variable data, which in present systems is largely performed by a human operator, should be automated wherever possible. It is fortunate that this requirement is fully backed by the present economic necessity for a much higher level of productivity through automation.

The flow-processing of collected data means that in certain instances the data will be dealt with immediately and the results may indicate to the central processor the need for a certain course of action at particular points, one of which may be the data source. In this situation a data collection system might grow to become a data collection and dissemination system, if economies in equipment cost could be derived from combining terminals and transmission facilities.

In conclusion it can be said that successful on-line data collection systems will only be produced by careful co-operation between the designers of the data collection systems, the designers of the central processor and the third party carrier if an out-plant system is required.

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Data Acquisition Systems for Control Applications

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Summary: The features of data acquisition systems are discussed. The advantages of modularity are stressed with particular reference to the interface concept. Basic system elements are described. Operations on data which can be carried out within the acquisition system are mentioned briefly.

1. Introduction

The increased reliability and greater systems power of the second generation of computers has provided the process operator with a powerful tool for the on-line optimization of his plant.

Unfortunately the types of signals produced by the instruments currently employed in process plants and those used in the control computer are radically different. Thus among those which occur most commonly are low level electrical analogue signals in the millivolt range from thermocouples, and pneumatic analogue signals in the range 3–15 lb/in² which are very common in local controllers and remote indication systems. Furthermore the number of such data sources within the one plant can run into several hundreds, and yet all need to be scanned periodically according to the control program. On the other hand digital computers can accept data only in binary electrical form and on one, or at most two, input channels.

Thus to enable both process and computer to work together data acquisition systems are needed which are capable of scanning a large number of data sources, converting each to a standard digital form and feeding the digital signals to a computer either directly, or via a temporary storage medium such as punched cards or magnetic tape.

This paper will consider some of the basic requirements of such systems and also some of the engineering techniques employed in their major elements.

2. Data Gathering Systems

The gathering of data for subsequent processing is a basic requirement for all data processing systems regardless of the particular application or nature of the data source. Thus as well as the acquisition of data from instruments on a continuous chemical process, data must be collected from the machine shop floor for job reporting in production control applications. But in spite of the varied applications

and data sources most of these data gathering systems can be classified under two main headings:

- (a) Data collection systems
- (b) Data acquisition systems

and before proceeding to a detailed consideration of the second class, it is necessary to distinguish them from the first. Needless to say the need for this division into separate classes arises from the weaknesses of current technology, and the division should disappear as more sophisticated and advanced methods are developed.

*Data collection systems*¹ are normally associated with production control applications and are characterized by the following features:

- (1) The data sources are predominantly digital, set by human operators, and have few condition status indicators such as valve position indicators.
- (2) The local stations provided near the data sources are usually complex and require a local display of indicative data.
- (3) The control of the main system is exercised primarily by the local stations, and the data receiving centre acts as a slave, its only positive control action being to assign priority among the local stations should several seek its services simultaneously.

Several commercial equipments of this type are available; what is probably the most advanced example uses only a 2-wire connecting cable between all stations and the central terminal, the signalling technique using frequency division methods.

Data acquisition systems are normally associated with process control applications and have the following characteristics:

- (1) The data sources are varied in nature, but are predominantly analogue. Both electrical and pneumatic instrumentation and control systems are to be found in existing plants; furthermore a large number of condition status signals must be scanned.

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- (2) The local stations are invariably simple, and unmanned.
- (3) The main system control is exercised by the central equipment which is generally complex and far more sophisticated in concept and operation than its counterpart in Data Collection.

It is this type of data gathering system with which this paper is primarily concerned.

3. Basic Data Acquisition Systems

To illustrate the major elements of these systems a simple data acquisition system is shown in Fig. 1. This is a typical configuration used for low-level electrical analogue sources such as thermocouples.

3.1. Analogue-to-Digital Conversion

The major component of the system is the analogue-to-digital converter (a.-d.c.), which compares the analogue input signal by a process of repeated weighing and approximation with an equal analogue voltage from a series of digital switches and quantizers within the unit. These switches represent the digital equivalent of the analogue signal in a buffer register for output to either the digital control computer, or a recording device such as a typewriter, paper tape or card punch or magnetic tape unit.

3.2. Amplification

The amplifier is included to raise the level of the thermocouple signals from the millivolt region to several volts, which is the input level usually required by the a.-d.c. Such amplifiers are designed to have exceptional gain stability and linearity and, for thermocouple applications, provide rejection of common-mode interference in the connecting cable by as much as one million to one.

3.3. Multiplexing

The amplifier/a.-d.c./output combination represents a single recording channel, and in order to share this

between a larger number of sources a multiplexer is required, which will either scan the several inputs in fixed sequence (prospecting) or in a sequence determined by the digital computer program (analysis). For this reason the use of rotary scanning devices is restricted to simple systems and are of little value in the type of data acquisition system we are considering here. Relay and solid-state switches are usually employed and provide scanning speeds of up to 50 or 20 000 channels per second respectively.

3.4. Special Features

It will be readily appreciated that such a simple data acquisition system would have limited applications even if the input signals were restricted to thermocouple voltages. In practice it is usual to include simple compensation, data reduction and out-of-limit detection features, so that the maximum logical power can be derived from the equipment employed. This is particularly important when a digital computer is not connected to the data acquisition system, since it enables the data acquisition system to produce indicative data which are suitable for alarm detection or operator guidance applications. If the data are fed directly to a digital computer, then the systems organization will probably arrange for these functions to be provided by the computer.

Among the more common features should be mentioned:

- (a) cold-junction compensation for thermocouple signals,
- (b) linearization of instrument characteristics,
- (c) alarm-level detection, both high and low,
- (d) peak level detection of signals from instruments such as mass spectrometers,
- (e) scaling of input signals, which usually involves adjusting the gain of the amplifier either automatically or under program control to achieve a wide dynamic range of measurement,

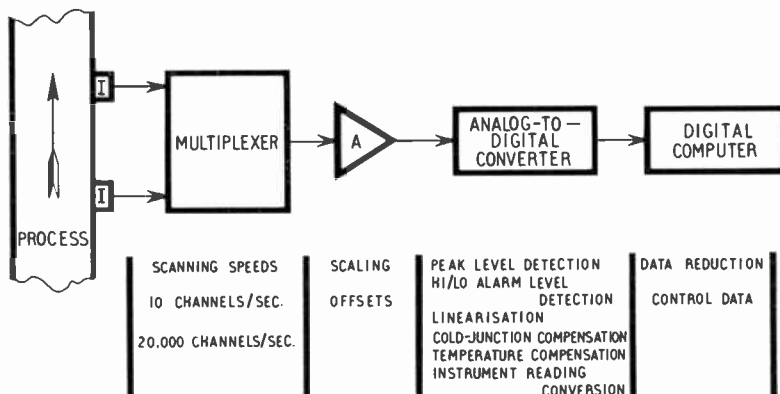


Fig. 1. Basic data acquisition system.

- (f) offsetting of input signals, which do not vary much, by fixed potentials under program control to achieve maximum accuracy of measurement,
- (g) simple data-reduction facilities such as
 - (i) temperature compensation of instrument readings
 - (ii) square-rooting to convert pressure drop across an orifice to fluid flow rate.

This list is by no means exhaustive, but summarizes the more common features. Some of the methods used to provide these are considered in detail later.

3.5. Nature of Input Signals

Although thermocouple signals constitute approximately 60% of signals found in process plants, it is necessary for data acquisition systems to handle other forms of data also. These can be summarized as follows:

- (a) electrical analogue signals in which either the current or voltage represents the parameter to be measured. In this class are included signals from thermocouples, strain gauges, and certain telemetry systems,
- (b) electrical digital signals which can be further sub-divided as:
 - (i) code modulation, such as are obtained from shaft digitizers and from the data acquisition system itself,
 - (ii) pulse count modulation, as emitted by integrating watt-hour meters in power stations, and special types of data acquisition systems incorporating ramp-type a.-d.c.'s which are described later,
 - (iii) condition or status indicators which are usually contacts attached to valves which indicate whether the valves are open or closed.
- (c) pneumatic analogue signals which are basic to the usual type of instrumentation, telemetry and local controllers to be found on contemporary process plants. Systems vary but the range of pressures is usually standardized to 3–15 lb/in².

4. Modular Approach in Data Acquisition System Organization

When dealing with a wide range of input signals such as those enumerated above, it is clear that a systems organization more complex than that of the basic data acquisition system will be necessary. Furthermore there is a strong danger that it might be necessary to design a special system for each application, and many systems have been individually tailored in the past.

On the other hand there are certain similarities between these systems, and one can easily classify the various input signals into well-defined groups and this encourages us to consider a modular approach to systems organization with its attendant advantages of manufacturing convenience, flexibility and ease of expansion.

Then to allow full freedom of interconnection between individual modules we need to standardize the conditions at the junctions. This involves more than agreement on voltage and impedance levels, however, and includes the type of control, form of addressing and the nature of the data signals themselves. We are led to the concept of a boundary across which flows the data in a standard form, and we call this boundary an "interface".

In defining the interface requirements we must ensure that the full capabilities of current technology are considered lest, in setting our sights too low, we prevent the systems designer from taking full advantage of the equipment used, and in setting them too high we make equipment costs exorbitant in terms of the performance obtained.

Such considerations lead to the concept of the "technological knee", which is the break-point which occurs in most performance/cost curves for the various units, above which the increase in performance achieved is small for the extra expense incurred. Clearly for each interface we should seek to define parameters, slightly below the "technological knee" values.

More specifically, for each interface we should seek to define, where applicable :

- (a) capacity,
- (b) the nature of the signals (voltage, current, etc.),
- (c) the code employed (for digital data),
- (d) signal level,
- (e) signal impedance,
- (f) the nature of the control signals,
- (g) the resolution or staticizing time,
- (h) the accuracy of the interconnection.

Since similar units are commoned at the interface, it is seen that some form of switching is necessarily involved, and the multiplexing requirement should be incorporated within the interface definition. For this reason, a separate multiplexer unit as shown in the basic data acquisition system does not appear in Fig. 2, which indicates one form of systems organization for a modular data acquisition system. This is based on a minimum of four major interfaces:

- (1) The transducer interface
- (2) The analogue interface
- (3) The digital interface
- (4) The output interface.

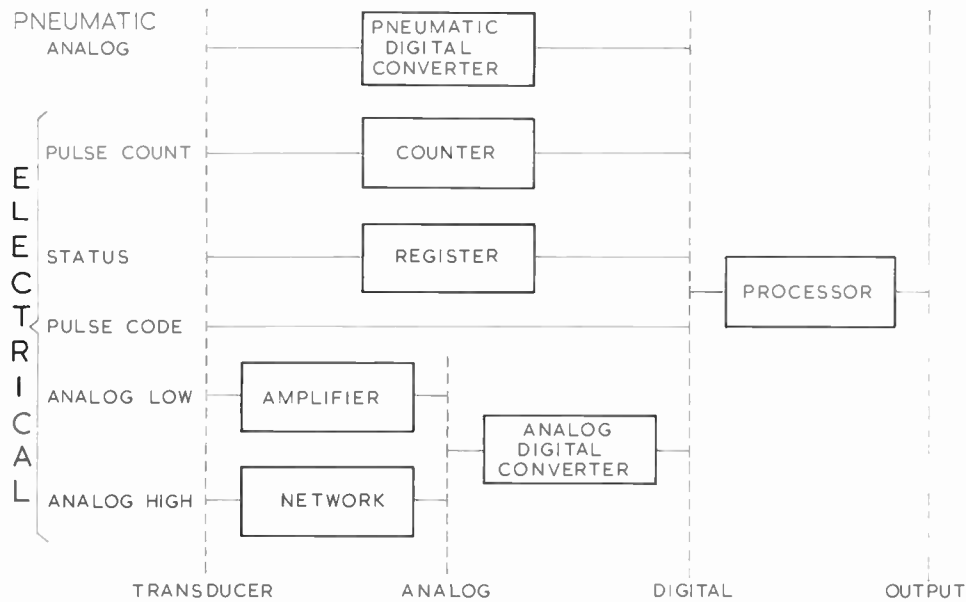


Fig. 2. Modular data acquisition system.

4.1. *The Transducer Interface*

This is the boundary between the process source instruments and the data acquisition system and will incorporate circuit elements to match the various signals listed above to one of the standard interfaces within the system. It is not expected that this interface will have only one standard form, but will have special sections within which signals will be standardized according to nature and magnitude. Some simplification of signal variations could be achieved here. For instance current signals could be converted to a standard voltage range by suitable terminating resistances. Potentiometers could be provided to adjust all voltage signals to a common range. Rejection filters for serial mode noise in thermocouple leads could also be included in this interface connection.

To avoid unnecessary degradation of signal/noise ratio, it will be advisable to separate low- and high-level signals and handle low-level signals in a special multiplexer before amplification and transfer to the analogue interface. Many types of multiplexer are available ranging from electromechanical devices operating at 50 channels per second to solid state elements capable of speeds of 20 000 channels per second. These speeds represent the two "technological knees" for the associated techniques. There is probably a need for both in view of cost differences and performance requirements, but the interface conditions for each type should be standardized. The definition of the interfaces should allow for the amplifier gain to be programmable, i.e. adjusted by control signals from an external source. Neither should it make the use of offsets difficult. The form of addressing will

also need definition, basic modules of 100 being particularly convenient for decimal addressing. Direct addressing of points by external controllers such as computers, and for the alternate modes of prospecting or systematic scanning should be allowed for and common mode rejection for certain types of signal specified. Terminating impedances should be carefully chosen to avoid excessive variation in potentiometric effect with varying lengths of instrument leads, and hang-over effects due to line capacitance—a value of 100 000 ohm is suggested as a useful compromise.

4.2. *Analogue Interface*

This interface must be defined to meet the needs of the a.-d.c. Source impedance should be low to reduce quantize errors in the a.-d.c. and a suggested value is 5 ohms \pm 10%. Speed, input voltage range and accuracy of the a.-d.c. are interrelated and the "technological knee" values are suggested as:

Accuracy	0.1%	0.01%
Voltage range	1 V	10 V
Conversion time	20 μ s	200 μ s

These two alternatives may have to be incorporated in the analogue interface to meet the two frequently conflicting requirements of performance and cost.

4.3. *Digital Interface*

All signals will be brought to this interface which represents the unprocessed data gathered by the data acquisition system. Analogue signals, both electrical and pneumatic, will be converted to this interface through suitable converters. Pulse count modulation signals are converted to coded form through suitable

counters. Condition or status signals will be connected directly through suitable level changing circuits or registers. Similarly keyboard or punched card sensed data will be connected to this interface.

Logical signal levels and power capacity will need to be defined. Suggested figures are 6 V level and 1000 ohms terminating impedance.

Choice of code will need careful consideration. A decision will first have to be made between pure binary or binary-coded decimal (or accept each as alternatives), and within the latter 8421 or 4221 (or other codes) chosen. The former is more conventional in data processing systems, but the latter leads to worthwhile simplification in a.-d.c. design.

The standard output interface has to cater for many requirements and is consequently expensive. The digital interface, if carefully defined, could provide a useful low-cost alternative.

4.4. Output Interface

This interface represents the major output from the data acquisition system. Since it has to connect to a wide range of output devices, such as computers, tape stores, it is important to conform to the standard input/output interface for the interconnection of peripheral equipment and data processing systems, which it is hoped will be established as a national standard in the near future.

The requirements of the digital and output interfaces are quite similar, but the existence of the two allows for simple processing to be provided. Alarm level detection, data reduction are but two facilities which could be provided.

Other types of output connection are provided through this interface and need no further definition. Thus a data transmission system associated with the data acquisition system would be regarded as a normal output unit.

5. Basic System Elements

Having discussed the general problems of systems organization and the advantages of modularity let us next consider briefly some of the engineering techniques employed in the basic elements.

5.1. Switching Elements

These are used in the multiplexers incorporated in most interfaces, and apart from the diode gates and logical devices similar to those found in data processing systems, they fall into two general classes:

- (a) electro-mechanical, which are characterized by:
 - low cost
 - high switching efficiency, measured in terms of contact resistance
 - contact flexibility

- compact size
- low switching speed
- low thermal e.m.f.
- low leakage.

- (b) solid state, which achieve nearly three orders of magnitude of improvement in speed over the electro-mechanical switching elements at the sacrifice of reduced performance for every other parameter.

The *sealed relay* probably represents the best all-round switching element available today when its performance is related to cost and provided it is made under a high standard of quality control. In its usual form it is a compact device, enclosing up to four changeover gold-plated contacts in an inert atmosphere (usually nitrogen). These contacts have resistances of the order of 50 milliohms, and a thermal e.m.f. of approximately 10 microvolts. Unfortunately their useful life is only about 5 million operations, which necessitates replacement every few years, and for this the whole unit has to be scrapped, which is its major weakness. This has led to the development of relays in which only the contacts are sealed, and inserted in suitable arrangements into unprotected solenoids to provide the necessary switching configurations.

In the *reed relay*, the contacts are formed of gold-plated magnetic alloy and are brought together under the influence of an external magnetic field usually provided by the solenoid, although permanent magnets can also be used. Switching speeds of up to 100 operations per second are possible with a life of well over 100 million operations. Contact resistance is again approximately 50 milliohms, although considerable care is necessary to hold the thermal contact e.m.f. down to 10 microvolts due to temperature gradients within the bulky control solenoid. Undoubtedly the reed relay is the best switching element to use when it can be afforded.

Another form of sealed unit is the *mercury-wetted contact*, in which the active contact is made between two globules of mercury brought to the contact area through the hollow wire of the contact bar by capillary action, or in more recent types by grooves cut in the bar of the armature. The thermal e.m.f. is usually high, of the order of 30 microvolts, due to considerable temperature gradients associated with the high power needed in the drive coil. The sealed enclosure is filled with hydrogen under considerable pressure, and care is needed in handling to avoid the unit exploding, which it can do with rather devastating results. The major disadvantage of the relays is the need to mount the units vertically because of the reservoir pool of mercury.

Solid-state switches usually take advantage of the low saturation e.m.f. of overdriven inverted junction transistors. A typical arrangement is shown in Fig.

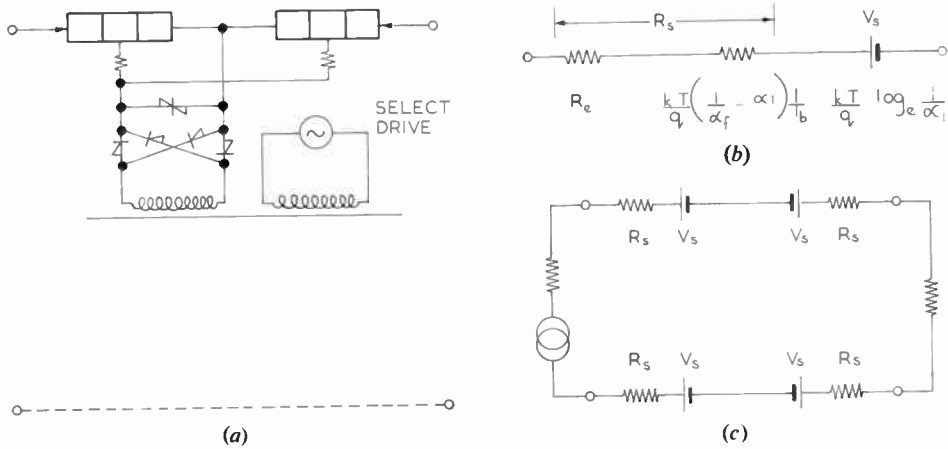


Fig. 3. Solid-state switch.

3(a) where only half of the switch is shown for ease of interpretation.

When heavily overdriven, the collector-to-emitter saturation voltage of a junction transistor is reduced to approximately 10 mV. As shown in Fig. 3(b) it depends on the reciprocal of the inverse current gain. Consequently if the transistor is inverted, the saturation voltage will depend on the forward alpha, which being larger than the inverse, gives an even smaller saturation voltage, of the order of 2.5 mV.

By connecting two similar units back-to-back, the saturation voltage of one can be made to cancel the other, as shown in Fig. 3(c) and the resulting difference e.m.f. is reduced to approximately 20 microvolts.

The success of this compensation depends upon the careful choice of transistor type and matching of individual units after suitable ageing.

In particular, low saturation resistance, low cut-off leakage current and low emitter resistance are looked for in determining the choice of transistor type, and silicon transistors have been found to be the most suitable. From the expressions in Fig. 3(b) it can be seen that it is essential to stabilize the base drive to close limits, and also to maintain all units at the same temperature by mounting in a heat sink, a gradient of 1 deg C being acceptable.

The switch is "opened" by applying suitable drive power through conventional gating to the drive transformer, and the base drive for each unit is derived through a bridge rectifier and defined by stabilizing the drive voltage and applying it through defining resistors to mask variations in the collector-base resistance. Switching speeds of up to 20 000 operations per second can be achieved, but the basic switching element is expensive compared with the slower electromechanical devices.

5.2. Amplifiers

For this type of application amplifier performance requirements are very severe and special configurations have been developed to meet them. Basically a gain stability and linearity of at least 0.01 % is necessary, and involves a high loop gain in the feedback loop, with its attendant stabilization problems, although some relief is obtained by restricting the bandwidth of the amplifier to match the switching speed of the associated multiplexers. For electro-mechanical switches operating at 10 channels per second a bandwidth of 100 c/s is normally sufficient. The basic gain itself should be capable of variation by an external control signal from a computer or similar program device. Frequently the amplifier gain is made the reciprocal of an instrument constant, so that an a.-d.c. operating in a fixed input voltage range becomes direct reading in the terms of the measured parameter. For example, a chromel-alumel thermocouple has a "constant" of approximately 40 microvolts per deg C, so if its signals are amplified by 250 and fed to a 3-digit a.-d.c. operating from 0 to 10 V, then a direct reading of degrees Celsius is obtained.

Considerable care in design is needed for the input circuits. Not only is a low d.c. zero drift (less than 5 microvolts) and high input impedance (more than 1 megohm) essential, but for thermocouple and similar low level sources, a well-balanced input is essential in order to reduce the effect of common mode interference injected into the connecting leads between transducer and amplifier, and common mode rejection performance of a million to one is frequently necessary.² Low zero d.c. drift is also necessary to reduce errors, and this coupled with the balancing and earth loop problems has caused one designer to include an isolating transformer between the output and input of his d.c. amplifier, which involves

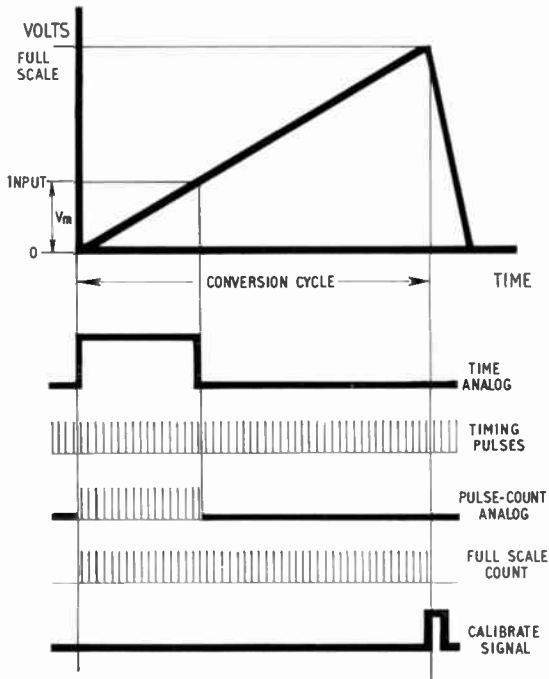


Fig. 4. Time encoding.

modulating and demodulating the d.c. feedback signal with an a.c. carrier for transmission through the transformer itself.

A low and controlled output impedance is also essential to feed the a.-d.c., since this impedance is frequently included in the quantize network of the converter and changes will affect the accuracy of that unit.

5.3. Analogue-to-Digital Converters³

Of the various types of a.-d.c., only two have found general application in data acquisition systems.

Time encoding is illustrated in Fig. 4.⁴ The input signal is compared with a standardized ramp function and the cross-over point of equality determined. Thus a time interval analogue of the input is produced and can be converted to digital form by gating pulses from a standard oscillator to a counter, which gives the digital output. It is difficult to produce a stable saw-tooth generator, and so provision is usually made to calibrate each cycle by reference to a standard voltage and to apply a suitable correction voltage to the generator. This type of a.-d.c. is capable of accuracies of 0.01%, but suffers from the disadvantage that the conversion time increases with increase in accuracy, and is limited to a maximum of 1000 conversions per second. The method is particularly useful where the instruments are widely dispersed, since the gated pulses can be sent through a simple data transmission system and counted in the central equipment. The loss of a pulse produces an error

equivalent to the resolution of the system, but the system is wasteful of both time and bandwidth. Techniques exist for the manipulation of pulse count modulated numbers, but they are rather expensive and slow.

Successive approximation is illustrated in principle in Fig. 5 and a typical system arrangement is shown in Fig. 6. Binary-weighted voltages are successively compared in decreasing order of magnitude with the input voltage and either discarded or retained if their contribution to the total causes it to exceed or be less than the input.

Such converters can achieve accuracies of 0.01% at speeds of 5000 conversions per second or of 0.1% at 50 000 conversions per second. They represent the most economical form of a.-d.c., when all factors such as cost, speed and accuracy are considered. Usually they operate in either pure binary code or in binary-coded decimal using 8421 notation, although certain simplification in design is possible using the 4221 notation; code conversion from this, to the more usual 8421 is readily achieved.

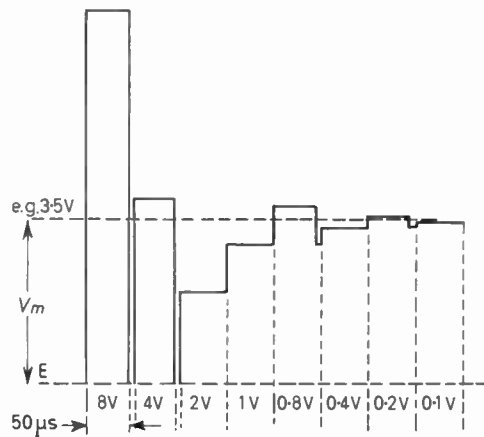


Fig. 5. Sequential balancing.

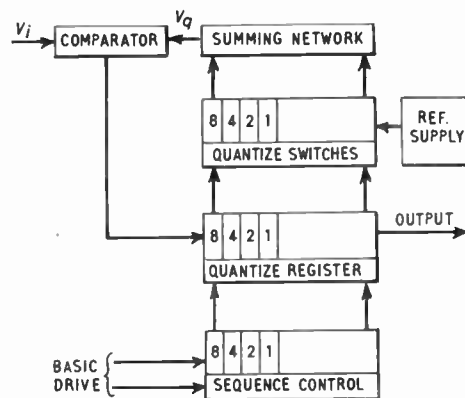


Fig. 6. Block diagram of an analogue-to-digital converter.

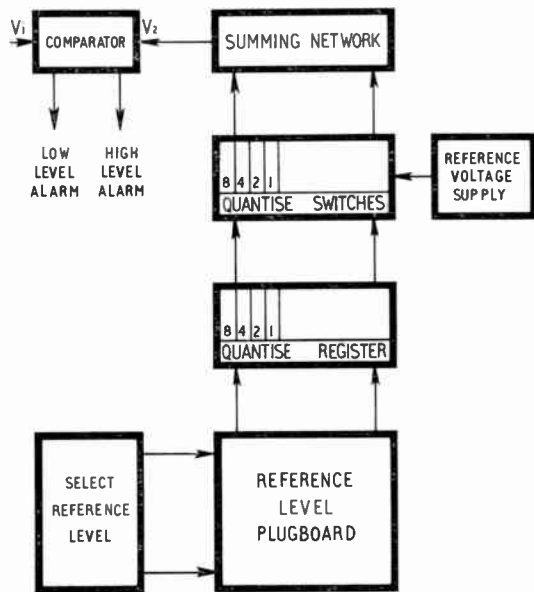


Fig. 7. Alarm level detection.

5.4. Output Equipment

Data from the data acquisition system will normally be fed directly to a computer or to temporary storage devices such as:

- (a) magnetic tape units, which will accept data at speeds up to 62 500 characters per second;
- (b) paper tape punches at speeds up to 150 characters per second;
- (c) card punches at speeds up to 18 characters per second serially or up to 1000 characters per second buffered;
- (d) data transmission systems which can send data at speeds of up to 250 characters per second over the telephone network.

6. Data Operations within the Data Acquisition System

Manipulation of the data acquired by the data acquisition system is normally best done in an associated digital computer. On the other hand there are some applications where such a computer is not justifiable and a need exists for certain types of operations to be carried out on the data within the data acquisition system.

Some of these techniques will be described in this Section, and it will be seen that all take advantage of the logical power of the successive approximation type of a.-d.c. Similar facilities could also be provided by the time encoding type of a.-d.c., but will not be described since they can be readily deduced from the other methods.

6.1. Alarm Level Detection

The most common requirement is to detect whether the value of a parameter has passed outside pre-determined limits, sometimes a high and a low limit. Often it is most convenient to set these limits in digital form either by switches within the data acquisition system or from the computer.

A simple method of achieving this is shown in Fig. 7, where it is seen the digital reference level is selected from a plugboard and set into the a.-d.c. register. From this register it is quantized and offered as the equivalent analogue voltage to the a.-d.c. comparator, which compares it with the input signal and indicates, via suitable logical output connections, whether a high- or low-level alarm condition exists.

6.2. Linearization of Transducer Characteristics

Many transducer characteristics are non-linear and require some form of compensation or linearization before a tolerable accuracy can be obtained. Normally the a.-d.c. simulates the transducer characteristic by a straight line approximation, whose slope is determined by the reference voltage, and can consequently be varied by this means. Furthermore, by injecting suitable voltages into the quantize network we can produce offsets in the straight-line simulation.

Figure 8 shows how a typical non-linear characteristic can be approximated with two straight-line simulations. The a.-d.c. can readily be switched to provide either of the two straight-line approximations, once it has been determined which is appropriate, i.e. on which side of the "break-point" between the

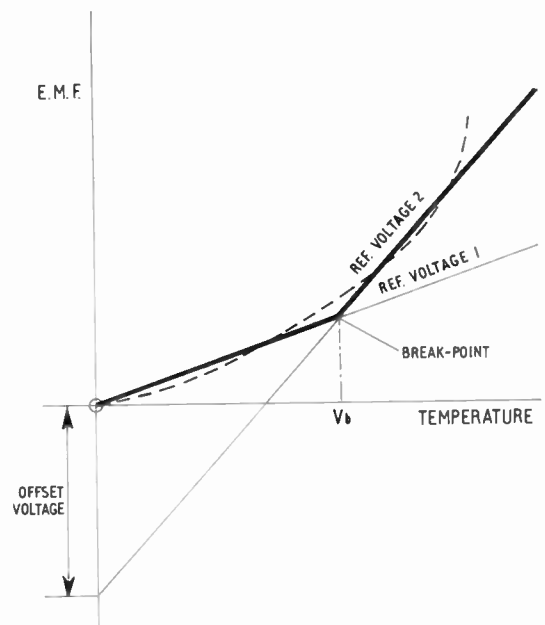


Fig. 8. Instrument characteristic linearization.

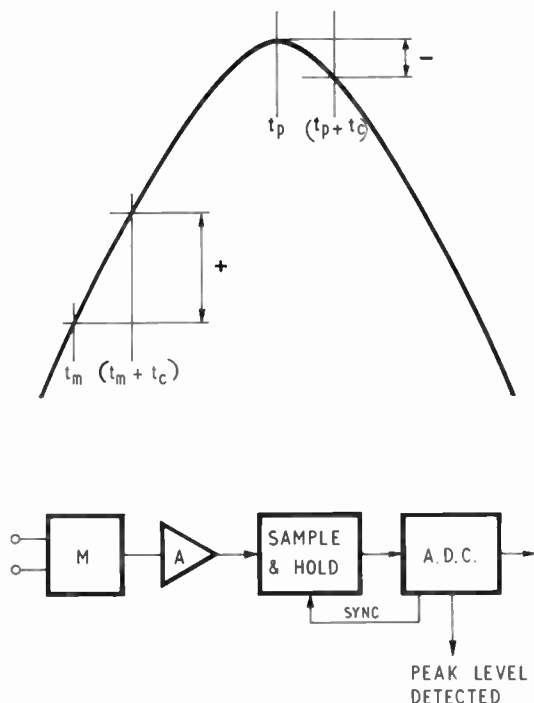


Fig. 9. Peak-level detection.

two, the particular parameter value lies. This decision is made using the same method as for alarm level detection, the “break-point” value being set into the converter from a reference register.

The basic method can be readily extended to any number of straight-line approximations.

6.3. Peak-level Detection

Some instruments, such as mass spectrometers, produce a time varying output, of which only the value of peaks is at all significant and the time (or conditions) at which they occur. By a simple modification to the internal logic of the a.-d.c., such peaks can be detected. The method uses the ability of the a.-d.c. register to remember the instantaneous value of the

input voltage and compare it a short time, t_c , later with the new value, as indicated in Fig. 9. If the new value exceeds the earlier value, then the a.-d.c. knows that the input is still increasing, so it replaces the old value with the new in the register, and after another interval, t_c , repeats the comparison. The peak value is detected by the new value being equal to or less than the previous value.

Should the variation of the input voltage during the a.-d.c. conversion be excessive, then a sample-and-hold circuit will be needed.

6.4. Alternative Amplifier—A.-D.C. Configuration

In an earlier section, the amplifier and a.-d.c. were treated as separate elements and their individual requirements discussed in detail. In this section some facilities have been discussed which involve modifications to the basic a.-d.c. Figure 10 shows a novel amplifier—a.-d.c. configuration which takes the amplifier into the control loop of the converter, and the quantizer into the feedback path of the amplifier. The resulting configuration has several advantages over the separate connection by easing the gain stability and linearity requirements of the amplifier and providing simple methods of scaling and varying the reference voltage.

By including the quantizer and range setting network in the feedback path of the amplifier, a variable gain amplifier of P/D is formed, provided A is very much larger than 1. Thus the voltage presented to the comparator is $(P/D) \cdot V_i$; and D is varied until

$$(P/D) \cdot V_i = V_r, \text{ the reference voltage}$$

$$\text{whence } V_i = (D/P) \cdot V_r.$$

Thus the system is equivalent to the normal a.-d.c./amplifier combination with scaling. The scaling network can be regarded as an extension of the quantizer and can be set up from external sources such as a computer.

The major advantages of the arrangement are:

- (a) the amplifier gain and stability requirements are eased, the loop gain being less than in the

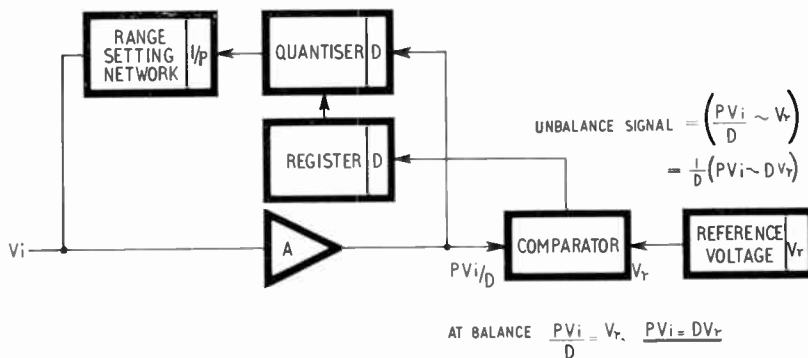


Fig. 10. Alternative a.-d.c. configuration.

separate amplifier because D is always less than 1; scaling is also more simply achieved,

- (b) the error signal at the comparator is larger than in the more conventional a.-d.c. since D is less than 1,
- (c) the reference voltage is now only lightly loaded, and so its value can be varied by simple means, such as a quantizer unit, to provide the variable slope characteristics needed in the above operations.

7. Conclusion

Only some of the possible techniques and systems arrangements which can be used in data acquisition systems have been described in this paper. Many others exist both for special purpose applications and general systems use, but it is hoped that those described will serve to illustrate the general trend.

The wide application of these systems will depend upon the achievement of the essential flexibility and expandability in design. For this, the adoption of standard interfaces such as those described in the paper is essential, and it is encouraging to note that

this approach is under active consideration by one trade association at the moment.

8. Acknowledgments

The author wishes to acknowledge the considerable help he has received from discussions with his colleagues in several laboratories of IBM engaged in this work and also to IBM United Kingdom Ltd. for permission to publish this paper.

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3. A. K. Susskind, "Notes on Analog-to-Digital Conversion Techniques" (Wiley, New York, 1957).
4. H. Fuchs and D. Wheable, "An all-electronic four-digit digital voltmeter", *J. Brit.I.R.E.*, **20**, pp. 541-7, July 1960.

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News from the Sections . . .

West Midland Section

The first meeting of the session to be held in Birmingham took place on 11th October when Mr. D. W. Robinson of the National Physical Laboratory read a paper on "Measurement of Traffic Noise". Defining his subject as the measurement of the external noise of individual vehicles, Mr. Robinson said that two schools of thought existed as to the type of noise measurement required. The first school suggested that the noise produced by vehicles running at ordinary town speeds should be measured, while the second school, regarding full throttle working as most offensive, considered that this alone should be measured.

Mr. Robinson suggested that any noise measuring system should have three basic requirements.

- (a) A system of physical measurement as opposed to individual judgment.
- (b) A set of limits on the physical scale.
- (c) A method of enforcement.

The paper dealt with these topics in considerable detail and was followed by a lively discussion.

K.L.H.

North Western Section

The North Western Section held its first formal meeting of the 1962-63 session on 4th October at the College of Science and Technology, Manchester. The paper which Mr. R. N. Jackson of the Mullard Research Laboratories read on "Colour Television Tubes" formed a useful continuation to the meeting held a month earlier at the A.B.C. Studios on the SECAM Colour Television System (See October *Journal*).

After years of speculation a colour television service was now expected to commence in Great Britain in about two years time and Mr. Jackson said that the question of the display device to be used in the receiver was vital to this service.

It was not surprising that the cathode-ray tube had received great attention as the means of displaying colour television pictures in view of its undoubted supremacy in the field of monochrome television. The fundamental factors to be considered in the design of colour cathode-ray tubes were features such as choice of screen structure, colour selection method and single or three electron guns. A brief review was given of the four principal known colour tubes, i.e. "Banana" tube, "Apple" (and "Zebra") tubes, "Chromatron" (or Lawrence) tube, and the "Shadowmask" tube.

Mr. Jackson concluded that of these four only the Shadowmask tube was at present in a state where it could be used as the basis for production. Its performance had been much improved over the past few

years and it could now be said to meet many of the requirements for a colour television service although the cost of a receiver based on this tube was still high.

He suggested that for the future we might expect a widening of the Shadowmask tube deflection angle from the present 70 degrees to 90 degrees with consequent reduction in neck length. This improvement would probably be at the expense of some slight increase in receiver circuit complexity. The possibility of using rectangular tubes in place of the present round tube envelope was under study, although this, of course, would introduce additional manufacturing and circuit problems.

There is still a possibility that a single-gun tube may be produced to compete with the Shadowmask tube but any such tube must have at least a 90-degree rectangular bulb and produce a high quality picture. The "picture on the wall" device is undoubtedly many years hence.

F.J.G.P.

South Wales

The first meeting of the new session was held on 3rd October at the Welsh College of Advanced Technology when Messrs. A. J. Sadler, B.Sc., and D. M. Martin, B.Sc., presented a most interesting paper on Silicon Controlled Rectifiers. These four-element transistor type devices were shown in conception and principle to be very little removed from a generation of semiconductor devices with which we are by now more or less familiar.

"S.C.R.'s" offer a higher efficiency at medium and low voltages than the thyatron (or mercury arc), they have a recovery 3-4 times faster, have no heater, are compact and robust (30-40 kW from a 2 in. cube), and need a controlling power which is very small indeed compared with the output power. All these properties are surely those for which design engineers are seeking when high efficiency, reliability and compactness are in most applications regarded as essential. There is a voltage limitation of course—the rectifiers must be used in series for h.v. operation, but for very large currents they can be used in parallel with no difficulty.

The discussion was mainly concerned with applications and varied from high power contactless switch circuits to the possibility of variable frequency variable output supplies. Time was once again unfortunately too short for the audience and the meeting had to be closed with questions left unasked. Maybe this is stimulating as it leaves one trying to work out the answers for oneself! The major query left of course was when would a good range of these devices be readily and cheaply available.

C.T.L.

Radio Engineering Overseas . . .

The following abstracts are taken from Commonwealth, European and Asian journals received by the Institution's Library. Abstracts of papers published in American journals are not included because they are available in many other publications. Members who wish to consult any of the papers quoted should apply to the Librarian, giving full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the journal unless otherwise stated. Translations cannot be supplied. Information on translating services will be found in the Institution publication "Library Services and Technical Information".

COMMUNICATIONS IN MINES

A recent German paper has discussed the properties which an r.f. balanced line system must have in order to permit communication between transportable stations which are inductively coupled to the line system within the tunnels of a mine. The effect of the surrounding earth on the transmission loss between transmitter and receiver is determined for various limiting cases. The required features of the receivers are outlined and the conditions limiting a carrier-frequency transmission of speech are mentioned.

"The conditions for carrier-frequency communication between transportable stations in mines", F. Buchholz. *Nachrichtentechnische Zeitschrift*, 15, No. 7, pp. 309-22, July 1962.

TUNNEL DIODE CIRCUITS

In an oscillator using a tunnel diode with a parallel LC tuned circuit, a variety of voltage waveforms from sinusoidal to relaxation type can be obtained with different LC combinations.

An Australian paper deals with the explanation of these wave shapes in terms of the L/C ratio of the tuning circuit, operating point, load, etc. It includes a description of general tunnel diode current voltage characteristics, their plotting and typical ranges of characteristics that are commercially available.

The basic oscillator circuit is described, with the requirements for oscillation. From a non-linear analysis of the circuit, the graphical process of obtaining the limit cycle for oscillation by the method of isoclines in phase space is outlined. Further graphical processing enables the voltage waveform for different $L-C$ combinations to be obtained and compared with a range of observed waveshapes.

Using the results of this analysis and the current voltage characteristics of the tunnel diode, a reasonable qualitative account can then be given of the observed waveforms. The paper concludes with a brief mention of the problem of synchronization to tunnel diode oscillators.

"Waveforms in tunnel diode oscillators", I. S. Docherty and R. E. Aitchison. *Proceedings of the Institution of Radio Engineers Australia*, 23, No. 5, pp. 304-10, May 1962.

COMMON CHANNEL INTERFERENCE IN TELEVISION

Two B.B.C. engineers have contributed a paper to the European Broadcasting Union describing measurements of the ratios of the amplitudes of wanted and interfering signals which produce a particular subjective grade of interference. The measurements were made in order to ascertain the required protection ratios for all the possible

conditions of mutual interference between stations operating on any of the three television standards at present in use in Western Europe; 405 lines, 625 lines and 819 lines with 50 fields per second. The results show that if the frequency difference between the wanted and interfering carriers can be maintained with a stability of better than 2 c/s (precision offset), each of the three wanted signal standards requires the same protection. In these conditions there is an advantage of up to 6 dB in having negative modulation for the wanted signal. If the offset frequency cannot be kept constant to within a few cycles per second, the required protection is smaller for the systems with the greater number of lines per field. A comparison between 525 lines 60 fields/second and 625 lines 50 fields/second is included.

"Common-channel interference between television signals of the same or different standards", G. F. Newell and E. W. Taylor. *European Broadcasting Union Review*, No. 72-A (Technical), pp. 56-67, April 1962.

STEREOPHONIC BROADCASTING

The March 1962 issue of *L'Onde Electrique* (the journal of the Société Française des Electroniciens et des Radio-électriciens) contains 12 papers on stereophonic broadcasting and associated matters. Details are as follows:

"Stereophonic broadcasting in the European setting", J. Matras (pp. 155-71).

"Stereophony and musical perception", M. P. Philippot (pp. 172-75).

"Principles involved in a quantitative study of directional perception in stereophony", H. Mertens (pp. 176-82).

"The placing of sounds in space", R. Condamine (pp. 183-86).

"Stereophony and broadcasting", R. Clouard (pp. 187-90).

"Sound pick-up and listening conditions in stereophony", L. Chatenay (pp. 191-99).

"From pure stereophony to monophony", R. Condamines (pp. 200-7).

"Theoretical plans for the distribution of frequencies in band II in the transmitter network grouped in transmission centres with its application to stereophony", J. Arnaud (pp. 208-18).

"Stereophony and magnetic recording", R. Buffard (pp. 233-40).

"A study of stereophonic broadcasting techniques", A. Keller (pp. 219-24).

"Stereophony in Italy", G. Castelnuovo (pp. 225-32).

"Stereophonic disks", R. Lafaurie (pp. 241-51).

Onde Electrique, 42, pp. 155-251, March 1962.

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